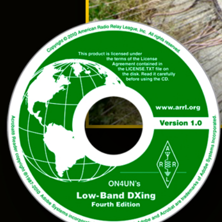
ON4UN's
Low-Band DXing

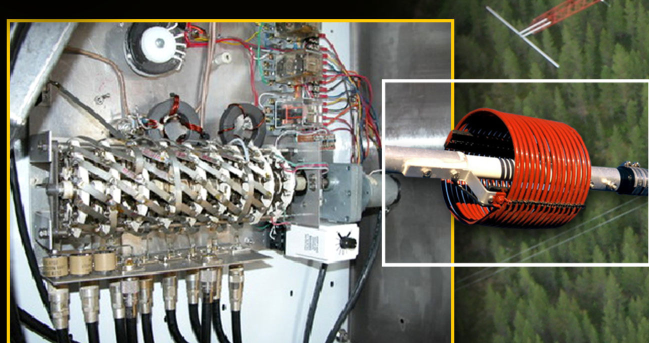
Fifth Edition

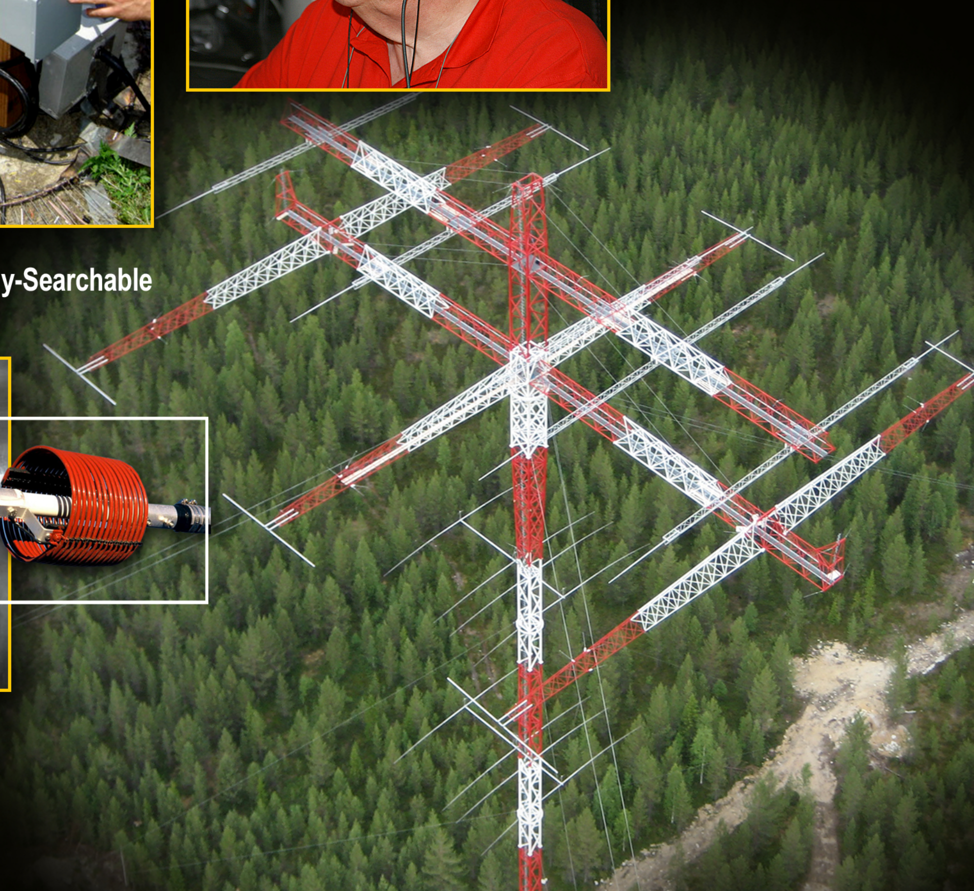
Antennas, Equipment and Techniques for DXcitement on 160, 80 and 40 Meters

John Devoldere, ON4UN



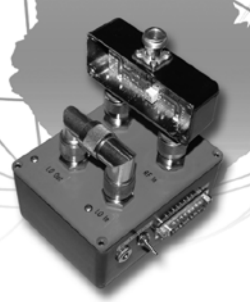
INSIDE: Inludes the Fully-Searchable Book on CD-ROM





ON4UN's Low-Band DXing

Fifth Edition



Antennas, Equipment and Techniques for DXcitement on 160, 80 and 40 Meters

John Devoldere, ON4UN

Production

Michelle Bloom, WB1ENT
Nancy G. Hallas, W1NCY
Jodi Morin, KA1JPA
Maty Weinberg, KB1EIB
Mark Wilson, K1RO

Technical Illustrations

David Pingree, N1NAS
Carol Michaud, KB1QAW

Cover Design

Sue Fagan, KB1OKW

About the Cover
(clockwise, beginning at lower right)

- Low band Yagis at Radio Arcala (OH8X): 3 elements on 160 meters, 5 elements on 80 meters and 4 over 4 elements on 40 meters—all on a 300 foot rotating tower. (photo by Juha Hulkko, OH8NC).
- The high-Q loading coil used in VE6WZ's shortened 80 meter Yagi.
- A motor driven rotary switch for direction switching in the K9DX Nine Circle array for 160 meters.
- The N2PK vector network analyzer (VNA). Chapter 11 describes how a VNA can be used for precise measurements during construction of phased arrays. (upper left corner)
- The feed point and matching network at the base of one of the elements in K3LR's 3-element vertical Yagi for 160 meters.
- Author John Devoldere, ON4UN.

Printed in the USA

ISBN: 978-0-87259-856-0

5th Edition
First Printing

Contents

A more detailed Table of Contents is
included at the beginning of each chapter.

Note: Chapter 16, Literature Review, is included with the PDF version of the book on the CD-ROM but is not part of the printed book.

Foreword

Do you enjoy a good challenge? Does your adrenaline flow at the thought of chasing elusive DX under the most difficult and unpredictable conditions? Do you like exploring technical issues and creating things with your own hands? If so, low band DXing is for you. Few areas of Amateur Radio offer the combination of technical and operating challenges presented by DXing on the low bands, and in particular on 160 meters.

It's been 25 years since author John Devoldere, ON4UN, published the original *Low Band DXing* to collect and share knowledge of techniques for success on 160, 80 and 40 meters. Over the years the book has evolved to keep pace with our growing knowledge of propagation, equipment, operating techniques and receiving and transmitting equipment for the low bands. This edition is no different, with significant new additions and updates throughout the book, and rewritten chapters on receiving antennas and phased arrays.

This book is a shining example of the Amateur Radio tradition of experimentation, collaboration and sharing of ideas and results. The many amateurs who contributed to this book are mentioned throughout the text.

John, ON4UN, is himself a recognized expert on low band DXing and his operating achievements speak for themselves. On 80 meters John has the highest number of DXCC countries confirmed worldwide (he is holder of the DXCC 80 meter award #1 with 357 countries confirmed on that band). When this book was published, on 160 meters John had the highest country total outside North America with 312 countries confirmed.

We hope that this new edition of *Low Band DXing* will inspire you to try something new — whether that's getting on 160 meters for the first time or upgrading to your existing station.

ON4UN has said that this will be his final edition of *Low Band DXing*. Thank you, John, for sharing your knowledge, enthusiasm and experience with us over the years, and for helping to stimulate our use of the low bands.

David Sumner, K1ZZ
Chief Executive Officer
Newington, Connecticut
October 2010

Preface

First of all I would like to thank the readers who signaled the inevitable typos and other errors in the 4th edition. You were very helpful. Also many thanks for your encouragement and suggestions.

Some readers wanted this book to be more of a book for beginners, giving step-by-step instructions on how to build the antennas, starting from a simple dipole. Let me be blunt: this is not a general antenna book or a beginner's book on antennas. The ARRL publishes the highly technical *ARRL Antenna Book*, as well as a wide variety of publications aimed at beginners and those requiring simple or small antennas. The *Low Band DXing* book is meant to complement these publications with a narrow focus on a fairly specialized field — the low bands.

This book explains the basic principles of all of the technologies (antennas, equipment, propagation and so on) required to become successful on the low bands. On top of that, *Low Band DXing* goes into detail on subjects that are typical for DXing on the low frequencies. Receiving antennas are one example, as well the very special propagation phenomena we witness on the low bands.

One of the nice things about every new edition is the opportunity it gives me to discuss various topics with a number of eminent experts acting as mentors for each of the chapters of this book. As in the past, many have been found willing to coach, support, advise and help me with the chapters. I am indebted a great deal to these fine gentlemen and true friends. They were my perfect critics, coaches, counselors and godfathers during the many, many months of hard work preparing this new edition. Thank you Roger (ON6WU), Robye (W1MK), Bill (W4ZV), George (W2VJN/7), Uli (DJ2YA), Klaus (DJ4AX), George (K2UO), Lew (K4VX) and Frank (DL2CC). Finally my thanks also go to the late L.B. Cebik, W4RNL, who reviewed one of the chapters of the new book just a week before becoming a Silent Key. Without all of you, the work would have been so much harder, and there would have been only half of the pleasure and satisfaction!

I am especially grateful to all authors and low band DXers/contesters who let me quote from their work, use figures from their publications or refer to their statements on various Internet reflectors. Your contributions were essential.

Some contributors also brought up new items that were expanded and included in this new edition. Examples include the new voltage feed method for arrays, brought up by Pekka, OH1TV, and the optimized hybrid coupler feed systems developed by Robye, W1MK, and Greg, W8WWV. Thank you Pekka, Greg and Robye! All of these are truly new items, never before published.

Finally, my thanks go to Steve, WB8IMY, at ARRL who twisted my arm and convinced me to write this "final" edition, and to Mark, K1RO, who helped edit this 5th edition, and to the production staff at ARRL for bringing all of the pieces together.

I hope that this new book will be instrumental in further promoting this wonderful hobby, where we move right in the steps of the pioneers of Amateur Radio. We do it at roughly the same wavelengths, we love experimenting, innovation and ingenuity, the sense of discovery, adventure, excitement and joy. It's everything that makes Amateur Radio unique. Unique in a way that satisfaction does not come automatically but as a result of dedication, education, knowledge, perseverance and patience.

I would like to dedicate this book to everyone who has made it possible for me to write the series of *Low Band DXing* books, a story that started 25 years ago. First of all, I think of my loving wife Frida, who has actively supported my wonderful hobby for half a century.

Frida Devoldere

But I would also like to dedicate this book to those "godfathers" of the different chapters in the different editions of the book, who were willing to spend time and effort to make this book into what it has become. One person I would like to mention in particular is indefatigable Robye Lahlum, W1MK. This book would not have been what it's become without Robye.

Finally I want to dedicate this edition of my book to Gaston Geirnaert, ex ON4GV, who became Silent Key in late 2009 while I was working on the last chapters of this edition. Gaston was my uncle. He introduced me to Amateur Radio over 55 years ago. He was my Elmer and gave me my first CW key. Without him this book would never have been.

To all of you, my readers, enjoy the low bands and have fun! I enjoyed writing for you.

John Devoldere, ON4UN
October 2010

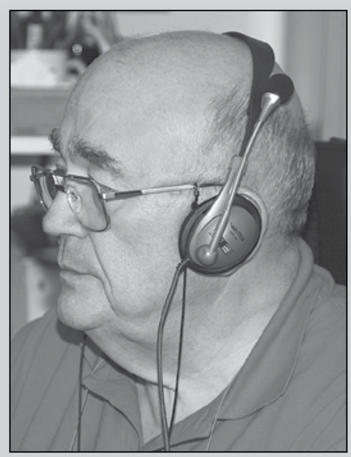
About the Author

John Devoldere was merely 10 years old when he was introduced to Amateur Radio. Ten years later he obtained the call ON4UN. John's interest in technology and science led him to become an engineer and his entire professional career was spent in the telecom world. All along he remained active on the bands, activity that has resulted in nearly half a million contacts in his logs.

In 1962, one year after he received his call sign, he took part in his first contest, the UBA CW contest, which he won. This was the beginning of a near 50 year long Amateur Radio career in which contesting and DXing — especially on the lower HF bands — have played a major role. On 80 meters John has the highest number of DXCC countries confirmed worldwide (he is holder of the ARRL DXCC 80 meter award #1 with over 355 countries confirmed). On 160 meters, at the time this book is published, John has the highest country total outside the US and Canada with 312 countries confirmed. John also was the first station world wide to obtain the prestigious 5B-WAZ award from *CQ*.

John wrote a number of technical books and articles concerning our hobby, most of which are published by the ARRL. These covered mainly low band antennas, propagation and on-air operating. He also wrote technical software on the subject of antennas, including mechanical design of antennas and towers. He is the co-author of the UBA (Belgian Amateur Radio Society, member of IARU) handbook for the HAREC-license (highest level license).

In 1963, as a very young ham, John got involved in Amateur Radio society affairs and became HF Manager for the UBA for a short period. More recently John served as President of the UBA between 1998 and 2007.



John Devoldere, ON4UN

In 2008 John combined his experience and expertise with that of his friend Mark ON4WW, to write a unique handbook *Ethics and Operational Procedures for the Radio Amateur,* which became the official document representing IARU's point of view on the subject matter. Since then the book has been translated in nearly 30 different languages.

A highlight in John's Amateur Radio career was undoubtedly his induction into the *CQ Contest Hall of Fame* in 1997 and into the *CQ DX Hall of Fame* in 2008, honors which until then had been bestowed upon only a handful of non-American hams.

Guide to ARRL Member Services

ARRL, 225 Main Street ♦ Newington, Connecticut 06111-1494, USA

Public Service

Advocacy

Education

Technology

Membership

Tel: 860-594-0200, Mon-Fri 8 AM to 5 PM ET (except holidays)
FAX: 860-594-0303
e-mail: hqinfo@arrl.org
ARRLWeb: www.arrl.org

JOIN or RENEW or ORDER Publications
tel. **Toll Free 1-888-277-5289** (US)
International callers
tel. +1 (860) 594-0355

VISITING ARRL HEADQUARTERS AND W1AW
Tours Mon-Fri at 9, 10, 11 AM; 1, 2, 3 PM
W1AW guest operating 10 AM to noon, and 1 to 3:45 PM (bring your license).

INTERESTED IN BECOMING A HAM?
www.hello-radio.org
e-mail: newham@arrl.org
tel. 1-800-326-3942

News Center

ARRLWeb: www.arrl.org

ARRL Letter:
www.arrl.org/arrlletter

Public Relations/Advocacy

Government Relations and Spectrum Protection:
www.arrl.org/regulatory-advocacy
e-mail: govrelations@arrl.org

Public and Media Relations:
www.arrl.org/media-and-public-relations

Membership Benefits

Membership Benefits (all):
www.arrl.org/membership

Awards: www.arrl.org/awards

Contests: www.arrl.org/contests

FCC License Renewal / Modification:
www.arrl.org/fcc-license-info-and-forms

QSL Service: www.arrl.org/qsl-bureau

Regulatory Information
www.arrl.org/national

Technical Information Service
www.arrl.org/technology
e-mail: tis@arrl.org
tel. 860-594-0214

Contributions, Grants and Scholarships

ARRL Development Office:
www.arrl.org/donate-to-arrl
e-mail: mhobart@arrl.org
tel. 860-594-0397

- ARRL Diamond Club/Diamond Terrace
- Spectrum Defense Fund
- Education & Technology Fund
- Planned Giving/Legacy Circle
- Maxim Society

ARRL Foundation Grants and Scholarships:
www.arrl.org/the-arrl-foundation

Public Service

Public Service Programs:
www.arrl.org/public-service

Amateur Radio Emergency Service® (ARES®):
www.arrl.org/ares

ARRL Field Organization:
www.arrl.org/field-organization

Clubs, Exams, Licensing and Teachers

Find an Affiliated Club: www.arrl.org/find-a-club

Mentor Program:
www.arrl.org/mentoring-online-courses

Find a Licensing Class:
www.arrl.org/find-a-class

Support to Instructors:
www.arrl.org/volunteer-instructors-mentors

Find an Exam Session:
www.arrl.org/finding-an-exam-session

Volunteer Examiner Coordinator (VEC):
www.arrl.org/volunteer-examiners

Publications & Education

QST — Official Journal of ARRL:
www.arrl.org/qst
e-mail: qst@arrl.org

QEX — Forum for Communications Experimenters:
www.arrl.org/qex
e-mail: qex@arrl.org

NCJ — *National Contest Journal*:
www.arrl.org/ncj
e-mail: ncj@arrl.org

Books, Software and Operating Resources:
tel. 1-888-277-5289 (toll-free in the US);
www.arrl.org/shop

Advertising:
www.arrl.org/business-opportunities-sales
e-mail: ads@arrl.org

Continuing Education / Online Courses:
www.arrl.org/online-courses

The American Radio Relay League, Inc.

The American Radio Relay League, Inc. is a noncommercial association of radio amateurs, organized for the promotion of interest in Amateur Radio communication and experimentation, for the establishment of networks to provide communication in the event of disasters or other emergencies, for the advancement of the radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

ARRL is an incorporated association without capital stock chartered under the laws of the State of Connecticut, and is an exempt organization under Section 501(c)(3) of the Internal Revenue Code of 1986. Its affairs are governed by a Board of Directors, whose voting members are elected every three years by the general membership. The officers are elected or appointed by the directors. The League is noncommercial, and no one who could gain financially from the shaping of its affairs is eligible for membership on its Board.

"Of, by, and for the radio amateur," the ARRL numbers within its ranks the vast majority of active amateurs in the nation and has a proud history of achievement as the standard-bearer in amateur affairs.

A *bona fide* interest in Amateur Radio is the only essential qualification of membership; an Amateur Radio license is not a prerequisite, although full voting membership is granted only to licensed amateurs in the US.

Membership inquiries and general correspondence should be addressed to the administrative headquarters: ARRL, 225 Main Street, Newington, Connecticut 06111-1494.

About the ARRL

The seed for Amateur Radio was planted in the 1890s, when Guglielmo Marconi began his experiments in wireless telegraphy. Soon he was joined by dozens, then hundreds, of others who were enthusiastic about sending and receiving messages through the air—some with a commercial interest, but others solely out of a love for this new communications medium. The United States government began licensing Amateur Radio operators in 1912.

By 1914, there were thousands of Amateur Radio operators—hams—in the United States. Hiram Percy Maxim, a leading Hartford, Connecticut inventor and industrialist, saw the need for an organization to band together this fledgling group of radio experimenters. In May 1914 he founded the American Radio Relay League (ARRL) to meet that need.

Today ARRL, with approximately 155,000 members, is the largest organization of radio amateurs in the United States. The ARRL is a not-for-profit organization that:

- promotes interest in Amateur Radio communications and experimentation
- represents US radio amateurs in legislative matters, and
- maintains fraternalism and a high standard of conduct among Amateur Radio operators.

At ARRL headquarters in the Hartford suburb of Newington, the staff helps serve the needs of members. ARRL is also International Secretariat for the International Amateur Radio Union, which is made up of similar societies in 150 countries around the world.

ARRL publishes the monthly journal *QST*, as well as newsletters and many publications covering all aspects of Amateur Radio. Its headquarters station, W1AW, transmits bulletins of interest to radio amateurs and Morse code practice sessions. The ARRL also coordinates an extensive field organization, which includes volunteers who provide technical information and other support services for radio amateurs as well as communications for public-service activities. In addition, ARRL represents US amateurs with the Federal Communications Commission and other government agencies in the US and abroad.

Membership in ARRL means much more than receiving *QST* each month. In addition to the services already described, ARRL offers membership services on a personal level, such as the Technical Information Service—where members can get answers by phone, email or the ARRL website, to all their technical and operating questions.

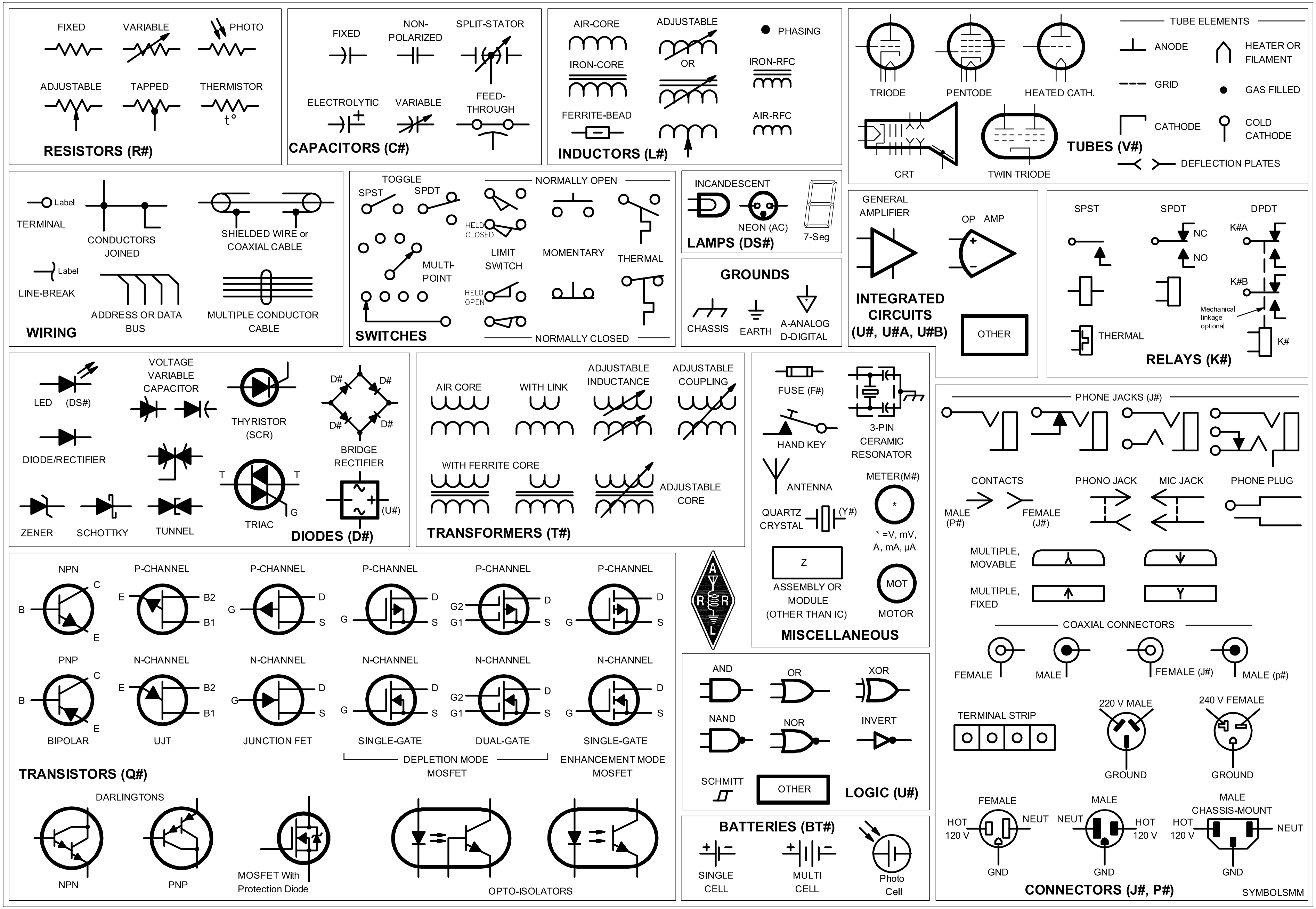
Full ARRL membership (available only to licensed radio amateurs) gives you a voice in how the affairs of the organization are governed. ARRL policy is set by a Board of Directors (one from each of 15 Divisions). Each year, one-third of the ARRL Board of Directors stands for election by the full members they represent. The day-to-day operation of ARRL HQ is managed by an Executive Vice President and his staff.

No matter what aspect of Amateur Radio attracts you, ARRL membership is relevant and important. There would be no Amateur Radio as we know it today were it not for the ARRL. We would be happy to welcome you as a member! (An Amateur Radio license is not required for Associate Membership.) For more information about ARRL and answers to any questions you may have about Amateur Radio, write or call:

ARRL—The national association for Amateur Radio
225 Main Street
Newington CT 06111-1494
Voice: 860-594-0200
Fax: 860-594-0259
E-mail: **hq@arrl.org**
Internet: **www.arrl.org/**

Prospective new amateurs call (toll-free):
800-32-NEW HAM (800-326-3942)
You can also contact us via e-mail at **newham@arrl.org**
or check out *ARRLWeb* at **www.arrl.org/**

Common Schematic Symbols Used in Circuit Diagrams



Contents

CHAPTER 1

Propagation

I am happy and proud to announce that for this 5th edition of this book, I found Carl Luetzelschwab, K9LA, willing to be my critic, my counselor, my helping hand, in one word a perfect godfather for the low band propagation chapter. Thank you, Carl, your expertise has been very welcome to update and further improve this important chapter. Carl wrote a number of great articles related to 160 meter propagation, articles that can be read on his Web site **mysite.ncnetwork.net/k9la.** I can highly recommend these articles as complementary reading material on this subject of this chapter.

Carl Luetzelschwab, K9LA, was licensed as WN9AVT in October 1961. He upgraded to General in May 1962, and traded in WA9AVT for K9LA in the mid 1970s when the FCC allowed Extra Class licensees to pick 1 × 2 call signs.

Carl received his BSEE and MSEE degrees from Purdue University in 1969 and 1972, respectively. Carl is an RF design engineer by profession with over 30 years of experience, and has done mostly transmitter and power amplifier designs for Motorola and Raytheon (formerly Magnavox). He currently lives in Ft Wayne, Indiana, and enjoys propagation, contesting, DXing and antenna work. His wife is Vicky, AE9YL, and they have had the good fortune to have participated in a couple of DXpeditions (YK9A in 2001 and OJØ in 2002).

K9LA's interest in propagation goes back to his MSEE days at Purdue, where his MSEE project dealt with group delay issues in the F_2 region. He currently writes a monthly propagation column for *WorldRadio* and is the propagation columnist for the *National Contest Journal* (*NCJ*). He also contributes propagation features to many other Amateur Radio publications. Select features are available on his Web site.

Carl holds 160 meter DXCC #960, and enjoys trying to understand propagation on 160 meters. He has publicly stated on the Top Band reflector that we're not likely to figure out 160 meters in our lifetime. He says that's fine with him, as it makes 160 meters more interesting!

1. INTRODUCTION

I will try to answer the following questions in this chapter:

- When can I work DX on the low bands?
- Are propagation mechanisms and propagation conditions on 40, 80 and 160 meters the same or similar?
- Are conditions very variable?
- Are conditions predictable?
- Is there any good propagation-prediction software?
- Are there any other tools to help me catch the difficult ones?
- What are crooked paths, when do they happen and what causes them?
- What directions should I aim my antennas?

In the last 10 years we have seen more publications than ever before covering propagation, antenna modeling and digital communications. It should come as no surprise that the availability of powerful PCs is a major reason for this evolution. Speaking of antennas and digital communications, the more that is being published on these subjects, the more we all benefit.

I cannot, however, say the same thing about publications concerning propagation, especially related to the low bands. The number of variables influencing low-band and especially 160 meter propagation appears to be so vast that scientists have only discovered the proverbial tip of the iceberg. So far, no one has come up with a forecasting system that works.

At best, we seem to be able to correlate — after the fact — some of the known parameters with actual observations. But the full "why and how," the global picture, is still missing. But don't let this scare you off. The elements of mystery and discovery on the low bands makes for half the excitement and fun there!

A number of publications (Ref 101, 103, 104, 105 and 167) cover the basic principles of radio propagation by ionospheric refraction, primarily on HF. Let me recommend in particular Robert Brown's (NM7M, now a Silent Key) excellent books: *The Little Pistol's Guide to HF Propagation* and *The Big Gun's Guide to Low-Band Propagation* (Ref 167 and 179). These books are must-reads for anyone who wants to have a more than casual understanding of propagation.

Both of these books are out of print.

The existence of books of this caliber makes it easy for me, since I will not have to explain the basics of the propagation mechanisms. This chapter on low-band propagation is thus not meant to be a general-study book on propagation. I have written it for the dedicated low-band DXer, who tries to understand the "how" of propagation (not necessarily the "why") to help him work an elusive new country or maybe to generate a better contest score.

This chapter is mainly based on observations, your observations and mine. You will soon find that the basic rules that govern propagation on the low bands are rather simple. You need a path in darkness, one that exhibits (very often) sunrise and/or sunset peaks. These are the simple but very important ground rules.

Testimonies of literally hundreds of low band DXers are used to identify what may seem like odd propagation phenomena. I will mention possible or probable mechanisms that govern these phenomena. These can be widely accepted, or in some cases more speculative and yet-to-be-proven mechanisms. What is important is that you are able to recognize "odd" phenomena and circumstances and that you know how to take advantage of them. That's what this chapter is really all about.

I will try to cover propagation on the three low bands (40, 80 and 160 meters), explaining similarities, but also pinpointing important differences. Understanding the basic mechanics is essential. If you realize that on 160 meters the opening over an 18,000 km path will occur maybe one day a week, and then only during a specific time of a "good" year, and that the opening will last maybe three to five minutes, you will realize how important it is to know when you have to try to make that contact.

A spectacular example to illustrate this was my QSO on 160 meters with ZL7DK in early March 1998. 1998 was a good, low sunspot year. I knew I had a three to five minute window both in the morning as well as in the evening. After observing the two windows for almost two weeks, one morning the weak signals from Chatham Island finally came through and I was able to work them. That morning "I had the skip." Other mornings their signals made it into England, France or Germany, without any spillover into Belgium. Why? In addition to just noting these facts, I will try to describe and explain a mechanism that seems to fit these facts.

Over the years numerous HF propagation programs have popped up. While these have proven their use for predicting propagation on the higher bands under quiet geomagnetic field conditions (the MUF-related bands), I have never had much use for them on the low bands, certainly not on 160 meters.

In the world of commercial HF broadcasting and HF point-to-point communications, the challenge consists in finding the optimum frequency or maybe the best angle of radiation (to select the right transmitting antenna) that will give the most reliable propagation, as a function of the time of day. In low-band DXing, the problem is quite different. The challenge is to determine the best time (month, day and hour) to make a contact on a given (low-band) frequency, with a given antenna setup, between two specific locations.

Cary Oler and Ted Cohen (N4XX) wrote in their excellent article "The 160-Meter Band: An Enigma Shrouded in a Mystery" (Ref 142): "Top Band is one of the last frontiers for radio propagation enthusiasts. It involves regions of the Earth's environment that are very difficult to explore and are poorly understood. These factors have led to our failure to predict propagation conditions with any level of accuracy. They also account for our inability to explain some of the puzzling mixtures of conditions that make this one of the most interesting and volatile bands available to the Amateur service."

Bill Tippett, W4ZV, hit the nail on the head when he wrote: "If 160 were perfectly predictable, we would all become bored with it and take up another hobby. Let's just enjoy it as it is because we'll never be able to figure it out!" So, don't expect this chapter to predict all kinds of exotic openings on 160 for you!

2. WHEN CAN WE WORK DX ON THE LOW BANDS?

Let's have a look at the following time cycles:

- The 11-Year Sunspot Cycle
- The 27-Day Sun-Rotation Cycle
- The Seasonal Cycle
- The Time of Day

2.1. The Sunspot Cycle

It is well-known that radio propagation by ionospheric refraction is greatly influenced by the sunspot cycle. This is simply because ionization is caused mainly by ultraviolet (UV) radiation from the sun, and therefore is highly dependent on solar activity.

Solar activity can influence HF propagation in three major areas:

- MUF (maximum usable frequency)
- Absorption from D layer and E layer
- The occurrence of magnetic disturbances

The sun's activity is usually expressed in terms of the Smoothed Sunspot Number (SSN) or the Solar Flux Index (SFI). **Fig 1-1** shows conversion from 2800-MHz solar flux to Smoothed Solar Number (SSN). You can get the SFI on WWV

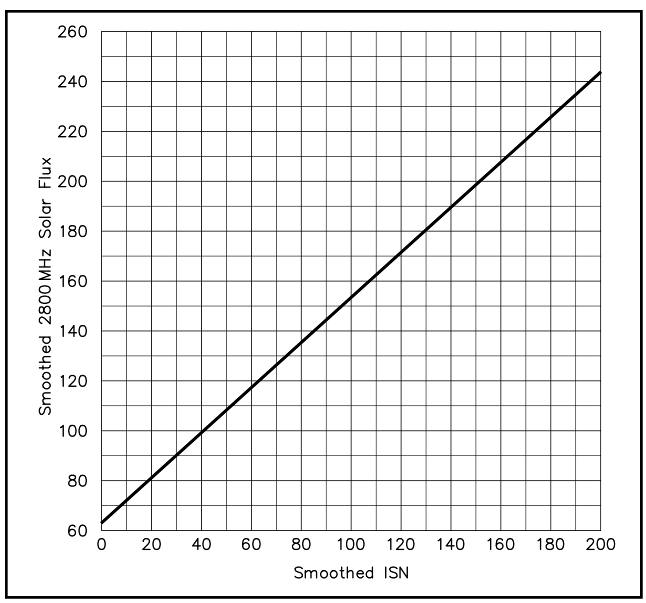


Fig 1-1 — Conversion from 2800-MHz solar flux to Smoothed Solar Number (SSN). (Courtesy *The ARRL Antenna Book*, 20th Edition.)

(2.5, 5, 10, 15 and 20 MHz), but if you have a computer with a permanent connection to the Internet, you can use a very nice program called *IonoProbe*, written by VE3NEA (**www.dxatlas.com**).

IonoProbe is a *Windows* application that monitors the space weather parameters essential for HF radio, including SSN/SFI, A_p/K_p, X-ray/Proton flux and auroral activity. *IonoProbe* downloads near-real time satellite and ground-station data, stores that information for future use and displays it in a user-friendly way. Time-critical parameters, such as X-ray flux, proton flux and auroral index, are updated every 15 minutes. An alarm can be set up to notify you of a geomagnetic storm within a few minutes after its start (**Fig 1-2**).

Fig 1-3 shows the previous five cycles (monthly mean and smoothed values) as well as Cycle 23 in more detail plus the forecasts for Cycle 24 in terms of smoothed values. (See **sidc.oma.be** and **www.swpc.noaa.gov/SolarCycle/**). Note that the smoothed sunspot number in Cycle 19 peaked almost twice as high (200) as Cycle 23 (just over 100). Old-timers will remember the spectacular 10 meter signals we enjoyed in 1957-1958 at the peak of Cycle 19.

You can also view a Solar Terrestrial Activity report in the form of a chart on **www.dxlc.com/solar/**. (See **Fig 1-4**.)

2.1.1. The MUF

The *critical frequency* is the highest frequency at which a signal transmitted straight up at a 90° elevation angle is returned to Earth. There are critical frequencies for the E region, the F_1 region, and the F_2 region. The critical frequency is continuously measured in several hundred places around the world by devices called *ionosondes*. At frequencies higher than the critical frequency, all energy will travel through the ionosphere and be lost in space (**Fig 1-5**). The critical frequency varies with sunspot cycle, time of year and day, as well as geographical location. Typical values for the F_2 region are 9 MHz at noon and 5 MHz at night. During periods with low sunspot activity, the F_2 region critical frequency can be as low as 3 MHz. During those times we can witness dead zones on 80 meters at night with higher elevation angles.

Table 1-1

α (°)	1/sin(α)	α (°)	1/sin(α)
0	3.37	50	1.27
10	2.94	60	1.14
20	2.28	70	1.06
30	1.78	80	1.01
40	1.47	90	1.0

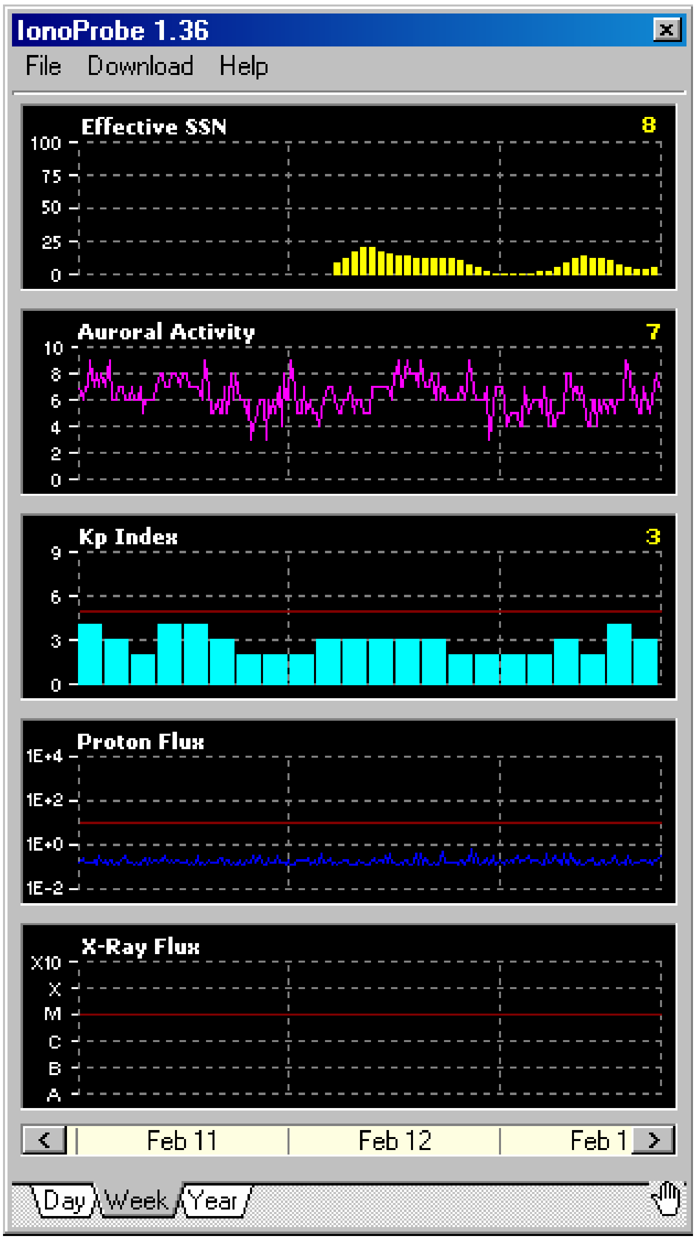


Fig 1-2 — A number of ionospheric parameters can be displayed together, for a timeframe of 1 day, 1 week or 1 year. For example, the Effective SSN, the K_p Index, Auroral Activity, X-Ray Flux and Proton flux are shown here for the last week. These are updated continuously using the *IonoProbe* program by VE3NEA. (Courtesy of VE3NEA.)

Fig 1-6 shows a world map with the monthly median critical frequencies for midwinter and a low Smoothed Sunspot Number (SSN=20) at 0000 UTC. **Fig 1-7** shows the same map for midsummer with a high Smoothed Sunspot Number (SSN=240) at 1200 UTC.

At frequencies slightly higher than the critical frequency, refraction will occur for a relatively high wave angle and all lower angles. As we increase the frequency, the maximum elevation angle at which we have ionospheric refraction will become lower and lower. At 30 MHz during periods of high solar activity, such angles can be of the order of 10° and even as low as 1° (Source: *VOACAP* by NTIA/ITS; see **www.uwasa.fi/~jpe/voacap/**).

The relation between MUF and critical frequency is the *wave elevation angle*, where:

$$MUF = F_{crit} / \sin(\alpha)$$

where α = angle of elevation.

Table 1-1 gives an overview of the multiplication factor (1/sin α, also called secant α) for the F_2 region at 300 km for a number of elevation angles (α) in a spherical Earth-ionosphere system. For the situation where the critical frequency is as low as 2 MHz it can be seen that any 3.8-MHz energy radiated at angles higher than 35° will be lost into space. This is one reason for using an antenna that concentrates its energy at a low radiation angle for the low bands. The calculation is done in spherical

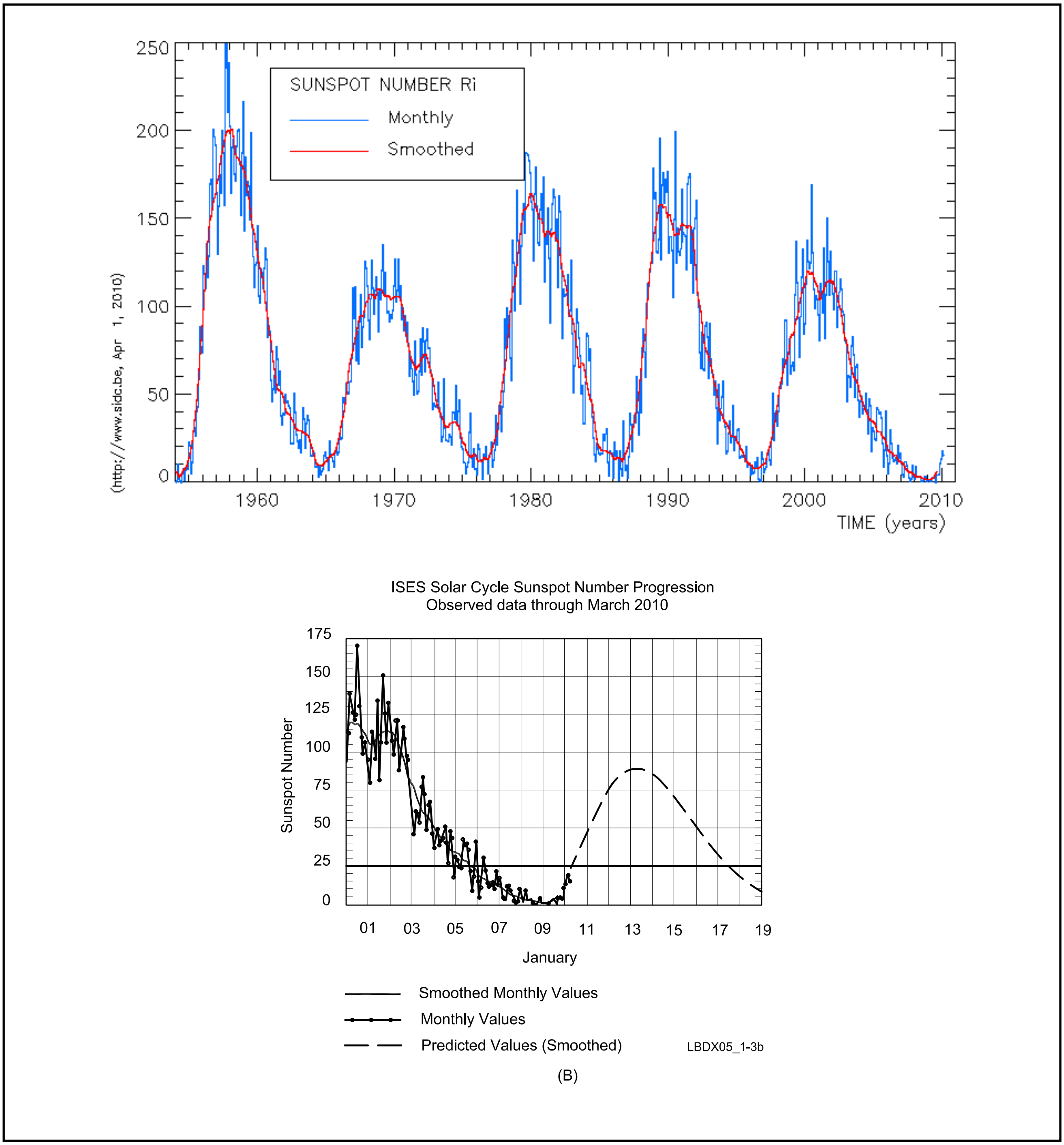


Fig 1-3 — Top: evolution of Sunspot Cycles 19 through 23, showing smoothed and monthly values (courtesy of Solar Influences Data Analysis Center, Royal Observatory of Belgium). Bottom: Sunspot Cycle 23 and the prognosis for Cycle 24. (Courtesy of NOAA/SWPOC Boulder, CO, as of 6 Apr 2010.)

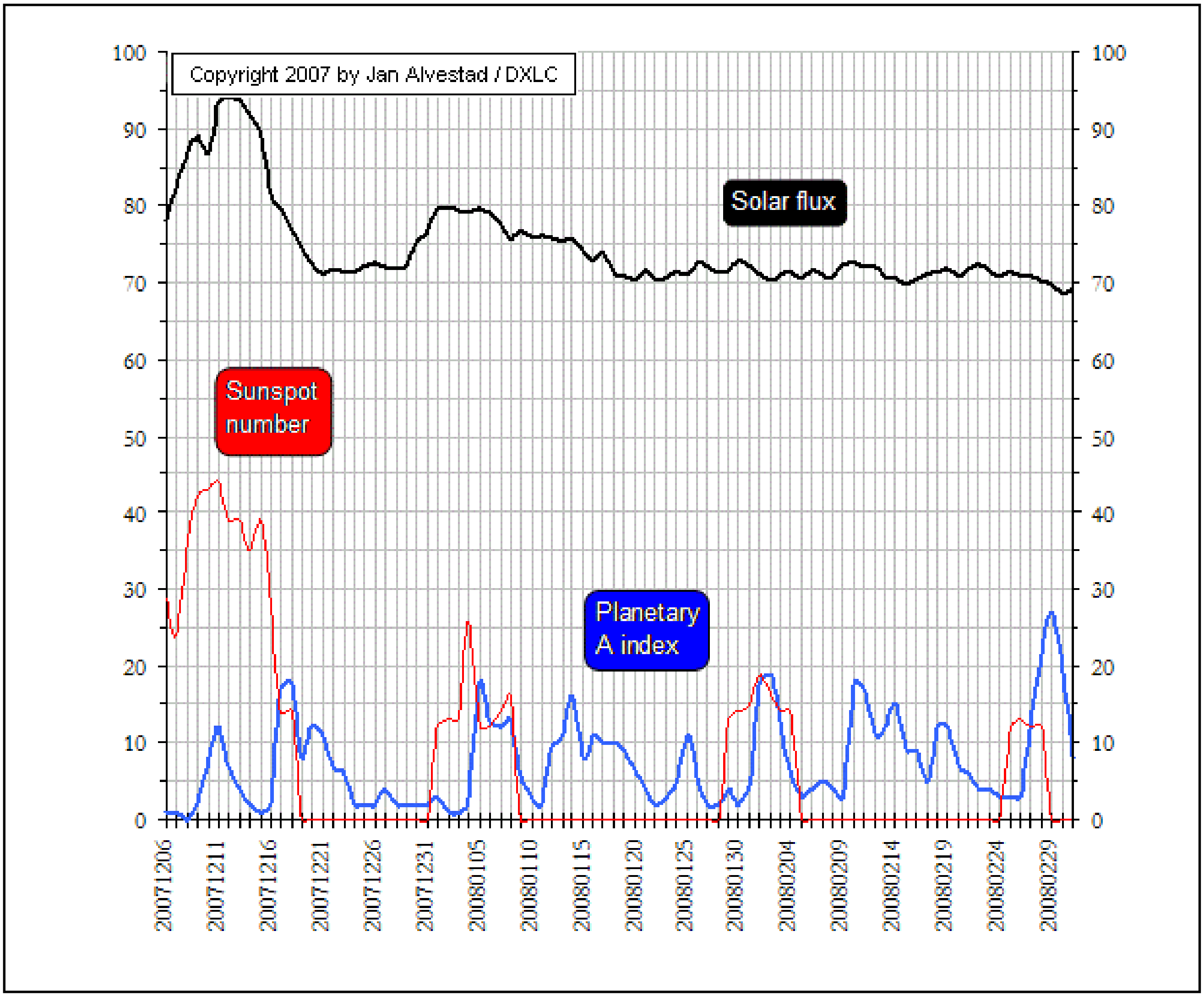


Fig 1-4 — The chart of the Solar Terrestrial Report shows the day-by-day evolution of the solar flux, the sunspot number and even the planetary A-Index. (Courtesy of Jan Alvestad.)

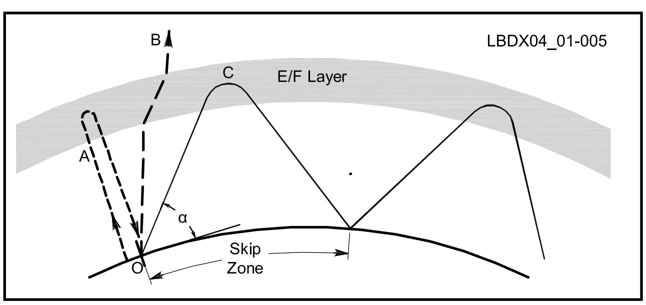


Fig 1-5 — Ionospheric propagation: At A, we see refraction of the vertically transmitted wave — this means that the frequency is *below* the critical frequency for that refracting layer. At B, the angle is too high and the refraction is insufficient to return the wave to Earth. At C, we have the highest angle at which the refracted wave will return to Earth. The higher the frequency, the lower the angle will become. Note the skip zone, where there is no signal.

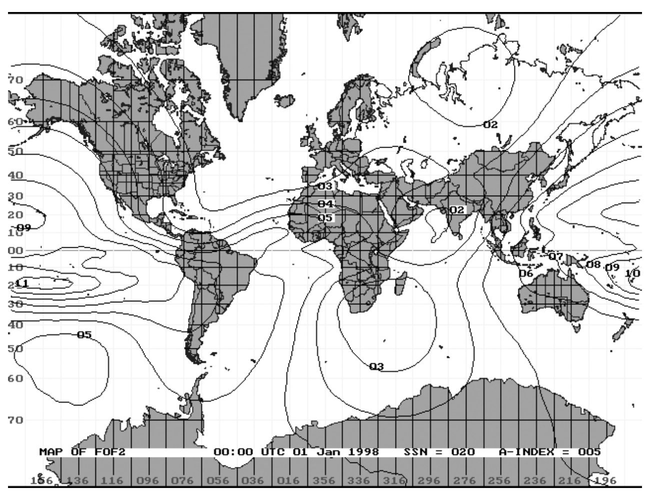


Fig 1-6 — Mercator-projection world map with the critical f_0F_2 frequencies for midwinter and a low Smoothed Sunpot Number (SSN = 20) at 0000 UTC. (Map generated by *Proplab-Pro* software.)

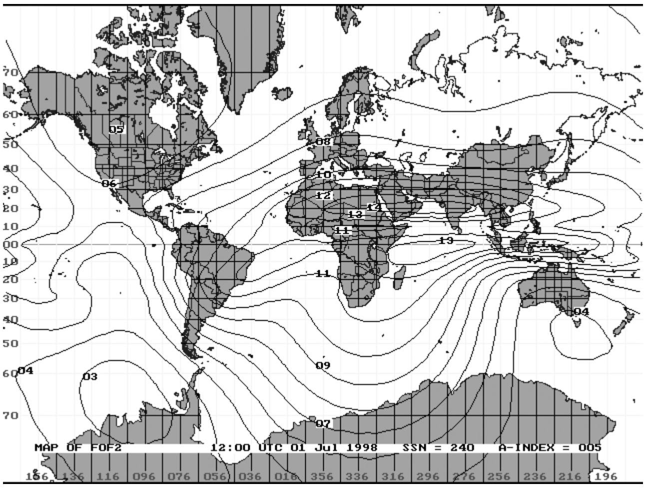


Fig 1-7 — Map for midsummer with a high SSN of 240 at 1200 UTC.

trigonometry as both the Earth and the ionosphere are spherical.

The MUF is the highest frequency at which reliable radio communications by ionospheric propagation can be maintained over a given path. This is also called the "classical MUF." Another common use of the term MUF refers to the *median statistical value*. Fifty percent of the time, the actual MUF observed on any given day will be higher than the median (and 50% of the time it will be lower). If you take 85% of the (median) MUF, the path should support signal propagation 90% of the time. This is also called FOT or the *frequency of optimum traffic*. At 115% of the (median) MUF, it should only support signal propagation 10% of the time. This is called the *highest possible frequency* or HPF.

Finally, just because signal propagation is supported

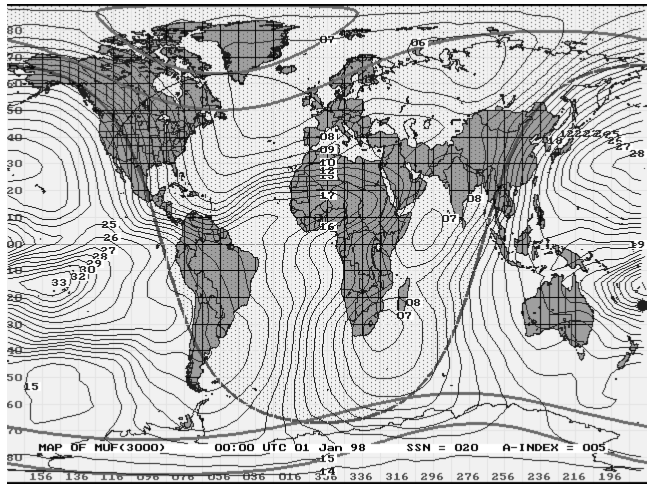


Fig 1-8 — Mercator projection of the world showing the 3000-km equal-MUF lines for January 1 (0000 UTC) for an SSN of 20 (A-Index=5). Note that there are areas in the Northern Hemisphere where during the night the MUF is below 7 MHz. Also note the great-circle (short) path between Europe and California.

does not mean that we will be able to communicate. For that, you have to consider the (S+N)/N, which is a function of the modulation type used and the noise field at the receiver, among other things.

The MUF changes with time and with specific locations on the Earth, or to be more exact, with the geographic location of the ionospheric refraction points. An MUF is always for a given signal path. The MUF for a path with multiple refraction points will be equal to the lowest MUF along the path. **Fig 1-8** shows a typical 3000-km MUF chart on a Mercator projection

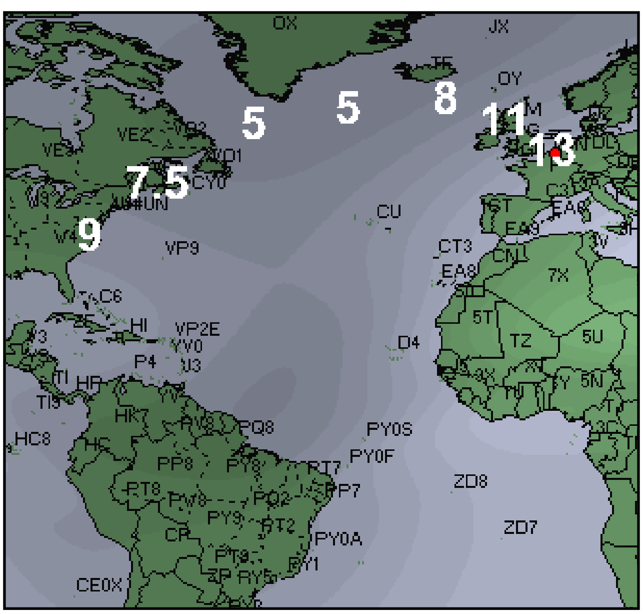


Fig 1-9 — MUF frequency map for January 1, 0800 UTC (SSN=20, K=0). The different shades of gray indicate different levels of MUF frequency. The darker the shade, the lower the frequency. The MUF on the path between Europe and the USA is below 7 MHz on a whole stretch of the path south of Greenland. (Map generated with *DX Atlas*, with additions by ON4UN.)

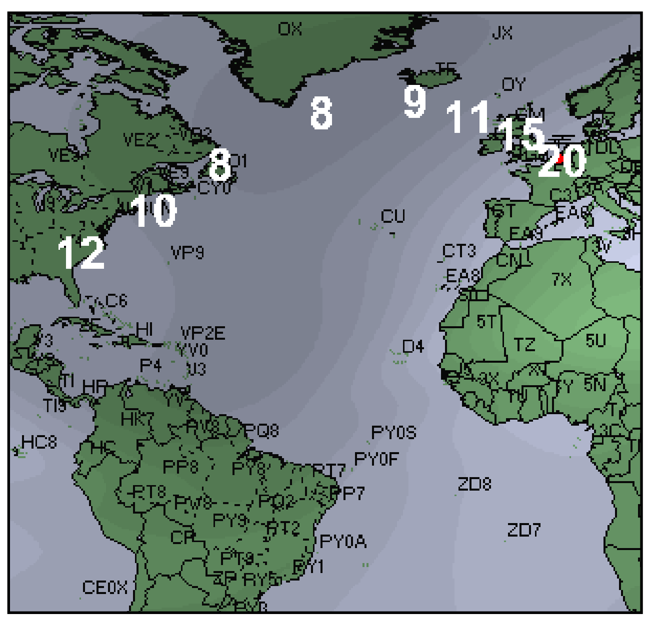


Fig 1-10 — This map is similar to the one in Fig 1-9; the only difference is that the SSN = 100. This time the path between Europe and the USA East Coast is open on 40 meters. (Map generated with *DX Atlas*, with additions by ON4UN.)

map produced by *Proplab-Pro*. (See Section 5.2.2.)

Each point of the MUF map is the midpoint of a 3000-km path. If an MUF of 7 MHz is shown for a given location, then you can expect an MUF of 7 MHz for any 3000-km path for which the mid-point is the given location. In other words, the MUF for a signal transmitted 1500 km away from the given location and for which the path goes through the ionosphere above that location is the value shown at that location on the map.

It is generally accepted that the optimum communication frequency (FOT) is about 80% to 85% of the MUF. On much lower frequencies, the situation is less than optimum, as the absorption in the ionosphere increases and the atmospheric noise generally increases. We now have computer programs available that will accurately predict MUF and FOT for a given path and a given level of solar activity under quiet geomagnetic field conditions. These programs are very useful in predicting propagation on the higher bands, as well as for 40 meters. Their usefulness is limited on 80 and they are even less useful on 160 meters, since propagation on that band is not ruled by MUF.

During low sunspot cycle years, the MUF is often below values sustaining 7 MHz long-distance propagation. For example, during low sunspot years the 7 MHz path between Europe and the USA very frequently closes down during the night and contacts are only possible near sunset (eastern end) or near sunrise (western end of the path). Even 80 meters can sometimes suffer from this phenomenon. The low sunspot years are therefore not always the best years for low-band propagation — despite the widely held belief of many lowbanders.

Fig 1-9 shows a so-called *frequency map*, generated using the *DX Atlas* program (see Section 5.2.1). This frequency map (which really is an MUF map) allows you to quickly assess the frequencies you can use into a given area of the world. This example is for a low sunspot number (SSN=20) on January 1 at 0800 UTC (Europe sunrise). Note that the map says there is 3.5 MHz propagation, but no 7.1 MHz propagation, between the USA and Europe, confirming what I showed earlier. **Fig 1-10** is for an SSN of 100, which guarantees enough ionization for a 40 meter path between Europe and the USA. In both frequency maps the geomagnetic A-Index (see Section 3.2.5.) was entered as 0, which means there would be virtually no solar-induced geomagnetic activity to disturb the path.

Fig 1-11 shows an "Oblique Azimuthal Equidistant" projection (commonly called a *great-circle map*) from *Proplab-Pro* that allows us to see the MUF values encountered along a great-circle path between a QTH (the center of the map) and

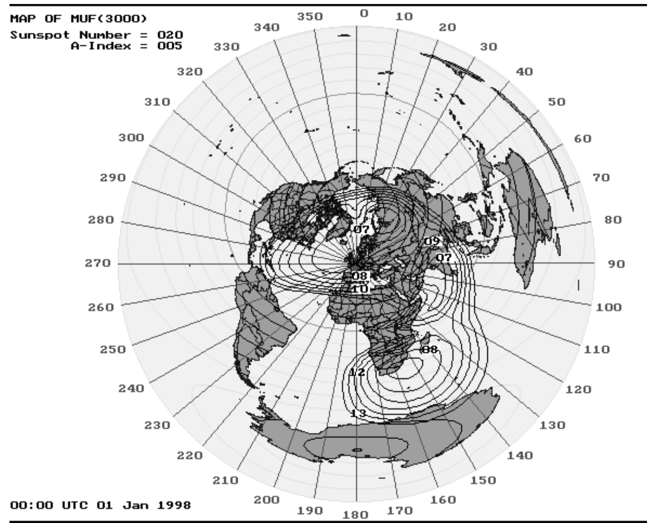


Fig 1-11 — Map showing 3000-km, equal-MUF lines, displayed on a great-circle projection (Azimuthal Projection). In this example only the MUF lines up to 10 MHz are shown, in order not to clutter the map. All the radial lines departing from the center QTH (Belgium) are great-circle lines, indicating the straight-line beam headings to all target locations. This same map projection (without the MUF lines) can be used to show beam headings from a particular QTH to DX around the world. (Map generated by *Proplab-Pro* software.)

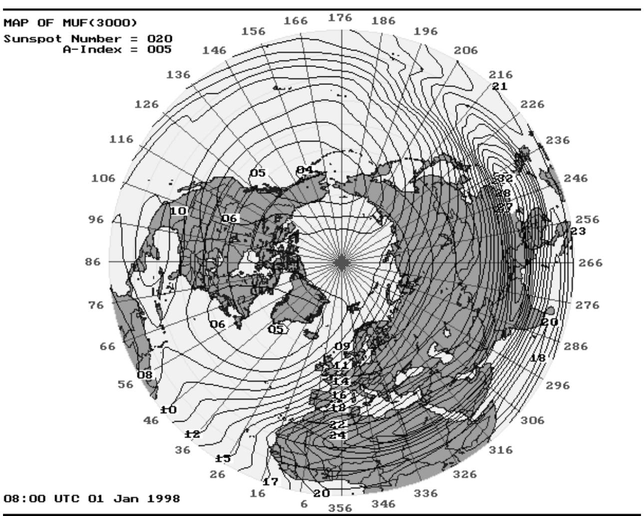


Fig 1-12 — The same 3000-km equal-MUF information as shown in Fig 1-11, but now displayed on a Polar Azimuthal Equidistant projection (globe view from above). This example is for 0800 UTC (sunrise in Western Europe). Notice the low MUF over the North Atlantic and North America. This shows why 40 meters can often go dead between Europe and the US during low sunspot years. (Map created by *Proplab-Pro* software.)

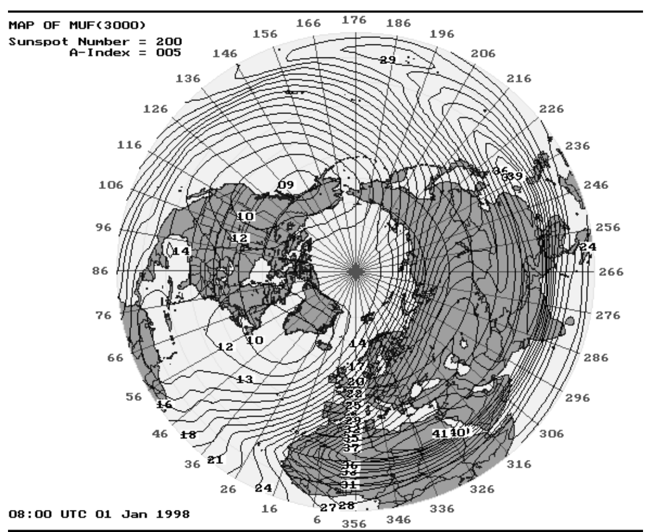


Fig 1-13 — The same 3000-km equal-MUF map shown in Fig 1-12, but for a very high sunspot number of 200. Note that even 10 MHz will remain open all night long between the USA and Europe. (Map created by *Proplab-Pro* software.)

a target QTH. In this example the great-circle map is centered on Western Europe. From these maps we can see that the MUF is lower during local winter and much lower at night than during the daytime.

Another very useful map projection is the "Polar Azimuthal Equidistant" projection in *Proplab-Pro*. This map shows either the Northern or the Southern Hemisphere and is very suitable for analyzing polar paths. **Fig 1-12** shows an example of such a map with equal-MUF lines for midwinter at 0800 UTC (sunrise in Western Europe) for a sunspot number of 20. Note again the low MUF zones between Europe and the USA. **Fig 1-13** shows the same map for a sunspot number of 200, indicating that 40 as well as 30 meters will remain open all night long between Europe and North America at this level of solar activity.

Important: Please note that the MUF has nothing to do with propagation on 160 meters, since the maximum usable frequencies are always greater than 1.8 MHz, even at solar minimum in the middle of the night. Propagation prediction programs based on MUF have no value in predicting propagation on Top Band.

2.1.2. Attenuation Through D Layer Activity

During the day, the lowest ionospheric layer in existence is the D layer, at an altitude of 60 to 90 km. **Fig 1-14** shows how low-angle, low-frequency signals are absorbed by the D layer. The D layer absorbs signals, rather than noticeably refracting them, because its critical frequency is well below 160 meters and the electron-neutral collision frequency is much higher than in the higher ionospheric layers. Collisions between electrons and neutral ions are responsible for absorption. The density of neutral, non-ionized particles, which make up the bulk of the mass in this region, is 1000 times greater in the D layer than in the E layer. (See Ref 121.) For a low-frequency signal to propagate through any layer without large losses, the number of neutral atoms should be small. Statistically speaking, a free electron in the D layer during the day would collide with nearby neutral atoms about 10 million times per second! The "collision frequency" is high, resulting in high levels of signal absorption.

During a typical day, the level of ionization of the D layer is due to ionization of nitric oxide and follows the solar zenith angle. It is also greatly influenced by the level of solar X-ray flares. During the night, the ionization level of the D layer drops dramatically but some very small level persists. The remaining ionization in the D layer helps determine the attenuation during the night on 160 meters. Small variations in D layer ionization can cause large fluctuations in signal absorption on Top Band. This is especially important in multi-hop propagation modes, where the signal has to traverse the D layer twice for each hop.

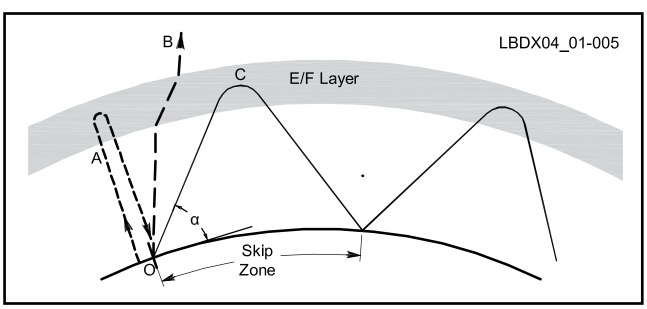


Fig 1-14 — D layer absorption. The high-angle signals pass through the D layer and are reflected by the E/F layer. Low-angle signals are absorbed. This explains the need for a high-angle antenna to work short distances during daytime.

The absorption level is inversely proportional to the arrival angle of the signal, so high-angle signals pass through the D layer relatively unattenuated. This is one reason our high-angle (low to the ground) dipoles work so well for local traffic on 80 meters.

I think this mechanism may also play a role in the often-reported phenomenon where, for periods shortly after sunrise, high-angle antennas often take over from low-angle antennas for working very long (eg, long-path) distances. (See also Section 2.4.4.4.) How might the sunspot cycle affect this phenomenon? When sunspot activity is low, the formation of the D layer is slower; D layer build-up before noon is less pronounced, while the evening recombination of the layer occurs faster. This is because there is generally less energy from the sun to create and sustain the high ionization level of the D layer. This means, in turn, that at a sunspot minimum, absorption in the D layer will be less than at a sunspot maximum, especially around dusk and dawn.

The absorption mechanism of the D layer has been studied repeatedly during solar eclipses. Reports (Ref 121, 125 and 180) show that during an eclipse, D layer attenuation is greatly reduced and propagation similar to nighttime conditions occurs on short-range paths. Elaborate testing shows that the "memory" of the D layer is very short, and once the irradiation of the sun rays stops, the D layer ionization drops to its very small residual value in less than a minute. This means that propagation between two stations that are fully in darkness will only be influenced by the variation in the remaining D layer density to a very small degree. Most of the attenuation will come from absorption at the bottom of the E layer (see also Section 2.1.3). Things are different when operating in the gray line (see Section 2.4.4.1), where signals have to punch through an already much ionized D layer. During high sunspot years, gray line signal enhancement will be less pronounced because of greater D layer absorption, thus producing fewer chances for signal ducting (propagation without intermediate hops).

2.1.3. The Influence of the E Layer on the Attenuation Mechanism

During the night the D layer that causes high absorption during the day is almost totally absent. Why then is nighttime absorption usually so much higher on 160 compared to 80 meters? The answer is that the absorbing region moves up from the D region in daylight to the lower E layer region (85-95 km high) in darkness (Ref 169). Even in the dark ionosphere there is still sufficient ionization in the lower E-region (and still a high-enough electron-neutral collision frequency) to cause such absorption, which is much more pronounced on 160 than on 80 meters. The typical absorption (in the ionosphere) for a single hop under these circumstances is 11 dB on 160 meters and 5 to 6 dB less on 80 meters (Ref 169). This limits the multi-hop communication range on 160 meters to approximately 10,000 km, assuming vertical monopoles, 1500 W transmit power, a typical receiving sensitivity at both ends, and a low noise environment on both ends.

This really fits very nicely with our day-to-day observations on Top Band: from Western Europe, Japan is about 9500 km. In winter time, working JAs on Top Band can be done almost every day with high power and a good vertical and Beverage antennas. The same holds true for countries like VR2, XU, etc. They are easy to work, almost day after day if well equipped stations are involved at both ends. JD1/O (Ogasawara), which is approximately 11,000 km, is a very different story. This seems to confirm that a different propagation mechanism may be involved once you pass the 10,000 km limit.

We will see later (Section 2.4.4.4.) that many longer paths on 160 are by ducting mode, rather than multi-hop modes. Ducting eliminates the losses connected with the E layer as well as losses incurred through ground reflections.

2.1.4. Magnetic Disturbances

Auroral activity is one of the important low-band propagation disturbances and is still largely a field of research for scientists. Amateurs living within a radius of a few thousand miles from the magnetic poles, however, know all about the disastrous consequences of aurora! The most favorable periods during the 11-year sunspot cycle (that is, when there is the least geomagnetic activity) occur during the up-phase of the cycle. Aurora acts like a big tall wall, thousands of kilometers long, located in the polar regions, a wall through which no radio signals can penetrate. The phenomenon of aurora will be covered in more detail in Section 3.2.

2.1.5. Low-Band Propagation During High Sunspot Years

For years, the generally accepted notion was that low-band DXing was not favored during high sunspot years, but not everyone shares this opinion any longer. Since there must be enough ionization to sustain some sort of propagation mechanism (refraction, ducting, etc), higher levels of solar activity should in theory be advantageous for low-band DXing, as well as for DXing in general. On the low bands sunspots affect mainly the 40 meter band, and sometimes 80 meters during the bottom of the sunspot cycle.

In the past there were fewer DX signals on the low bands during high sunspot years. To a large degree this was due to the absence of other DXers to work. The relative lack of specific interest in low-band DXing kept run-of-the-mill DXers away from the low bands when 20 meters was open day and night, and the higher bands were open during the day. Multiband and specific low-band awards increased emphasis on low-band operating during contests. Further, the growth of an elite group of Top Band and 80 meter DXers (stations that do not work on "VHF" bands like 20 and higher) has been instrumental in raising the activity on the low bands all through the cycle and all throughout the year.

Whereas 40 years ago none of the DXpeditions would waste their time operating on 80 or 160 meters, nowadays every DXpedition worth the name does include 40, 80 and 160 meters in their operating schedules, and all through the sunspot cycle. Even in the middle of the summer in the high sunspot years we can often hear several stations calling CQ DX on Top Band. Remember, when it's summer in the Northern Hemisphere, it is winter down south.

Still, the sunspot cycle has some impact on low-band propagation, although not as drastic as on 10, 12 and 15 meters, where the MUF rules propagation. Low-band propagation, particularly on 160 meters, is rather "digital." Either the waves are refracted or ducted in the ionosphere or they are not, and they are lost into space.

On 160 meters, the main limiting mechanism is one of *attenuation*, since 1.8 MHz is always lower than the MUF (the amount of absorption is inversely proportional to the square of the frequency). There are a number of ways in which Top Band signals are attenuated during their travels. If enough signal survives being attenuated by all these mechanisms and we can hear the signal above the local noise floor, then we say we have propagation. Distance plays an important role on 160 meters. N6TR recognized this when he introduced the Stew Perry (W1BB) Top Band contest, where scoring goes by distance, just like in some VHF/UHF contests. We know that the sun is involved in various mechanisms causing the cumulative path loss.

1) *Aurora*: It appears that we enjoy the geomagnetically quietest years during the minimum and the rising phase of the sunspot cycle (Ref 142). See Section 3.2. and **www.spacew.com/swim/bigstorm.html**.

2*) Ionization levels in the E and F layers during the night*: On 160 and 80 meters, when the entire path is well into darkness, propagation is primarily by multiple hops in the F layer, with little or no D layer absorption. However, the remaining ionization of the E layer takes its toll in attenuation (up to 11 dB ionospheric loss per hop on 160 meters). The sun's activity (as witnessed by the sunspot numbers) will influence E layer absorption. The difference may not be a spectacular number of dB, but we often operate on the verge of what is possible (very low S/N ratio), where a few dB can make the difference. (See Section 2.4.4.)

3) *D layer*: The remaining ionization of the D region during the night plays a role, especially in the multi-hop propagation. (See Section 2.4.4.) During gray line periods absorption in these regions is more substantial than during the night, hence high levels of sunspots do influence gray line propagation.

4) *Ionospheric ducting*: Conditions conducive to an ionospheric ducting mechanism are more easily met during low sunspot years. (See Section 2.4.4.5.)

We all know that good propagation, especially over long distances (>7000 km) occurs much more frequently during low sunspot years than during high sunspot years. On 160 meters during high sunspot years, from Belgium I can reach to the US Midwest when conditions are good. But I never work California during the high sunspot years. As a rule, good openings between Western Europe and the West Coast of the USA happen *only* during low sunspot years when geomagnetic field activity is lowest.

2.2. The 27-Day Solar Cycle

The sun rotates around its own axis in approximately 27 days. (Since the sun is essentially a fluid, the higher latitudes rotate more quickly and the lower latitudes rotate more slowly.) Sunspots and other phenomena on the sun can last several solar rotations. This means that we can expect similar radiation conditions from the sun to return every 27 days. DXers (both on the low bands and the higher HF bands) look forward to a repeat of very good conditions 27 days after the last very good ones have occurred — and their expectations are often met. This is probably the only somewhat reliable propagation prediction system for 160 meters that we have at this time! If conditions are good today, the *lack* of bad things is why propagation may possibly be good 27 days from today. If you just have had outstandingly good conditions on 160 meters, always mark your calendar for 27 days later. There is a fair chance you may have the same or similar good conditions again.

However, if conditions today are very bad (maybe due to a solar flare), there's no telling whether conditions will be bad in 27 days, since a solar flare does not repeat every 27 days. A recurring coronal hole, however, would most likely repeat in the next 27 day period, since these tend to hang around for at least several solar rotations. Unfortunately, during the early years of a solar cycle this predicting system may not be very reliable because the sunspots don't hang around for even one full revolution most of the time. As the cycle matures, the 27-day recurrence becomes more important.

2.3. The Seasonal Cycle

We know the mechanism that originates our seasons: the *declination* of the sun relative to the equator (in other words, the Earth's axis of rotation is not perpendicular to the plane on which Earth orbits around the sun). This tilt reaches a maximum of 23.5° around December 21 and June 21 (see **Fig 1-15**). This coincides with the middle of the Northern Hemisphere winter propagation season and the middle of the summer propagation season. At those times the days are longest (or shortest) and the sun rises to the highest (or lowest point) at local noon in the non-equatorial zones.

On the equator, the sun will rise to its highest point at local noon twice a year, at the equinoxes around September 21 and March 21. These are the times of the year when the sun-Earth axis is perpendicular to the Earth's axis (sun declination is zero), and when nights and days are equally long at any place on Earth (*equi* = equal, *nox* = night). On December 21 and June 21, the sun is still very high at the equator (90° – 23.5° = 66.5°). The maximum height of the sun at any latitude on Earth is given by the expression:

Height = 90° – north latitude + 23.5° (with a maximum of 90°)

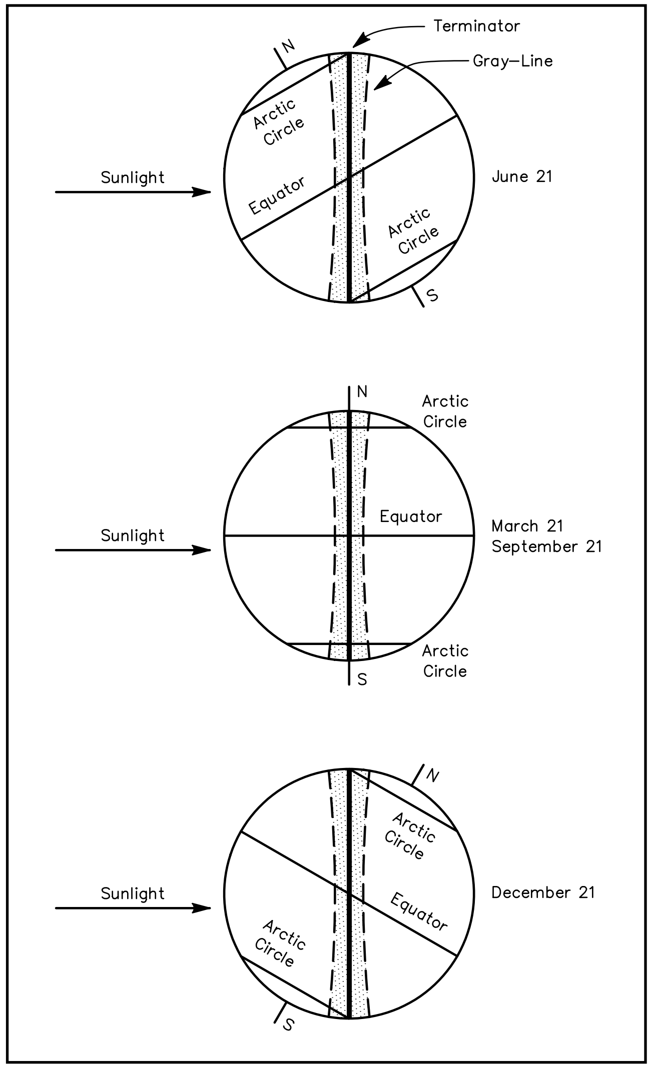


Fig 1-15 — These drawings show the declination of the sun at the different positions (solid vertical lines) at different times of the year. The gray line is represented as a zone of variable width to emphasize that its behavior near the poles differs from that near the equator.

In other words, the sun never rises higher than 23.5° at the poles, and never higher than 53.5° where the latitude is 60°. The seasonal influence of the sun on low-band propagation will be complementary in the Northern and Southern Hemispheres. Any influence will be most prominent near the poles and less pronounced in the equatorial zones (±23.5° of the equator).

But how do the changing seasons influence propagation?

1) The longer the sun's rays can create and activate the D layer, the more absorption there will be during dusk and dawn periods. During local winter in areas away from the equator, the sun will rise to a much lower apex and the rate of sunrise will be much lower. Accordingly, D layer ionization will build up much more slowly.

2) If the sun rises quickly (local summer in areas away from the equator) the configuration of the D, E and F layers necessary to set up a wave-ducting mode will last for a much shorter time than in winter. *Gray line propagation* will thus last longer in the winter than it will during summer.

3) In non-equatorial areas, many thunderstorms are generated in the summer. Electrical noise (QRN) easily masks weaker DX signals and can discourage even the most ardent DX operator. On north-south paths (US-to-South America or Europe-to-Africa), the Northern Hemisphere summer is usually the best season, since QRN is likely to be of less intensity than the QRN during the Southern-Hemisphere summer.

4) When the nights are longest during local winter, you will have more time (longer nights) for DX openings. Indeed, you must be in darkness or in the gray line zone (see Section 2.4.4.) not to suffer from excessive D layer absorption on 80 meters and even more so on 160 meters.

5) The occurrence of aurora is most pronounced around the equinoxes (March-April and September-October).

2.3.1. Winter (15 October to 15 February in the Northern Hemisphere)

Winter is characterized by higher daytime MUFs and lower nighttime MUFs, shorter days, lots of darkness, sun rising slowly, longer gray line duration (see Section 2.4.4.1) and fewer discharges from lightning strikes (QRN) from local thunderstorms. This period is best for all stations located in the Northern Hemisphere during the winter. Conversely, this condition will not exist in the Southern Hemisphere. Therefore the winter period in the Northern Hemisphere is ideal for east-to-west and west-to-east propagation between two stations both located in the Northern Hemisphere. Typical paths are US-to-Europe, US-to-Japan, US-to-Asia, etc.

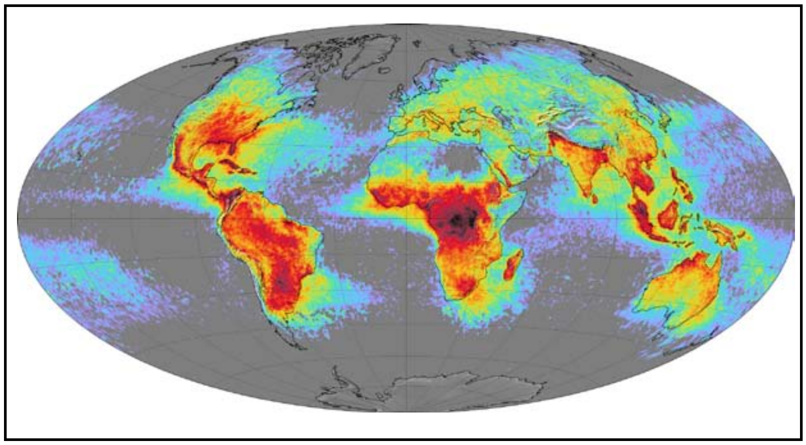


Fig 1-16 — Average worldwide lightning activity. Black areas: more than 70 lightning flashes per square km per year. Light gray (most of Europe): less than three flashes per year. (Source: NASA.)

2.3.2. Summer (15 April to 15 August in the Northern Hemisphere)

Summer is characterized by lower daytime MUFs and higher nighttime MUFs, longer days, faster-rising sun, increased D layer activity at dusk and dawn, and higher probability of QRN due to local and distant thunderstorms. These factors create the worst conditions for east-to-west or west-to-east propagation in the Northern Hemisphere. However, good QRN-free openings to the west can be possible just around local sunrise at the eastern end of the path. While most amateurs may be fighting the local QRN in the Northern Hemisphere in summertime, our friends down under are enjoying excellent winter conditions. This means that summertime is the best time for transequatorial propagation (for example, from Europe to southern Africa or North America to the southern part of South America).

2.3.3. Equinox Period (15 August to 15 October and 15 February to 15 April)

During these periods the ionospheric conditions are fairly similar in both the Northern and the Southern Hemispheres (with the symmetry about the geomagnetic equator, not the geographic equator): similar MUF values, days and nights approximately 12 hours long on both sides of the equator, reduced QRN, etc. Clearly, this is the ideal season for "oblique" transequatorial propagation on the NE-SW and NW-SE paths. Typical examples are Europe to New Zealand and West Coast US to SE Asia (NA morning) or East Coast NA to Indian Ocean (NA evening).

2.3.4. Propagation into the Equatorial Zones

In principle, all seasons can produce good conditions for propagation from the Northern or Southern Hemisphere into, but not across, the equatorial zone. On 80 and 40 meters the only real limiting factors can be the MUF distribution along the path and especially the amount of QRN in the equatorial zone itself. Unfortunately, there is no rule of thumb concerning the electrical storm activities in these zones. From Europe, we work African stations and stations in the southern part of South America on 160 meters mainly during the months of June, July and August. A similar situation exists between North America and the southern parts of Africa and South America.

There are a number of Web sites where you can check for current thunderstorm activity. See **Fig 1-16**.

2.3.5. Low Bands are Open Year-Round

It is clearly incorrect that DX on 80 and 160 meters can only be worked during the local winter, a popular belief not so very long ago. If that were true we (in the Northern Hemisphere) never would work countries located in the Southern Hemisphere. The equinox period is the best time of year for equatorial and transequatorial propagation, while in the middle of our Northern Hemisphere summer (if QRN is acceptably low for us), we often work rare DX stations from down under or from the equatorial zones. Even good east-west openings can happen in the middle of the local summer, so long as there is a darkness path and the QRN level makes listening for the signals at all possible. Of course the operators on both sides must be willing to try, and not take for granted that it won't work. Over the years I have worked quite a few rare ones over an east-west path in summertime. Here are a few examples:

XYØRR (Burma) was worked on September 3, 1991, on 160 meters, shortly before his sunrise. After a QSO on 80 meter SSB, we moved to 160 meters, where 579 was exchanged. After the QSO XYØRR called CQ a few times, but nobody else came back. Those convinced that summer time is not a good time for 160 meters were wrong once again. The little story behind this contact is that I normally don't have a "long" Beverage up for that direction during the summer. There happened to be 2.5-meter (8-ft) tall corn growing in the field in that direction. That day I spent a memorable couple of hours putting an insulated wire right on top of the cornfield. You must try that sometime for fun!

Another striking example is what happened during the DXpedition of Rudi, DK7PE, to S21ZC in early August 1992. The first night he was on 160 meters, I was in the middle of a local thunderstorm (S9 + 40 static crashes) and no chance for a contact. The next day, the QRN was down to S7, and a perfect QSO (579) was made over quite a long path (comparable with a path from ET3PMW to the US East Coast in June, or from the West Coast of Africa to California). See also Section 3.3.

2.4. The Daily Cycle

We know how the Earth's rotation around its axis creates day and night. The transition from day to night is very abrupt in equatorial zones. The sun rises and sets very quickly; the opposite is true in the polar zones. Let us, for convenience, subdivide the day into three periods:

- Daytime: from after sunrise (dawn) until before sunset (dusk).
- Nighttime: from after sunset (dusk) until before sunrise (dawn).
- Dawn/dusk: around sunrise and sunset (gray line periods).

2.4.1. Daytime

Around local sunrise, the D layer builds up under the influence of radiation from the sun. Maximum D layer ionization is reached shortly after local noon (the D region is under direct solar control via the solar zenith angle). This means that from minimum absorption (due to the D layer) before sunrise, the absorption will gradually increase until a maximum is reached just after local noon. The degree of absorption will depend on the height of the sun at any given time. The "memory" of the D layer is very short, in contrast with the memory of the F-layer. This has been proven by several experiments during solar eclipses (Ref 180).

For example, near the poles, such as in northern Scandinavia, the sun rises late and sets early in local winter. The consequence will be a late and very gradual buildup of the D layer. In the middle of the winter the sun may be just above the horizon for regions just below the Arctic Circle, situated at 90° – 23.5° = 66.5° above the equator. Or the sun may actually be below the horizon all day long for locations above the Arctic Circle. Absorption in the D layer will be minimal or nonexistent under these circumstances. This is why stations located in the polar regions can actually work 80 meter DX almost 24 hours a day in winter (provided there are quiet geomagnetic conditions, see Section 3.2). Contacts between Finland or Sweden and the Pacific or the US West Coast are common around local noon in northern Sweden and northern Finland in winter on 40 and 80 meters and even on Top Band.

I suppose that this is not a good example of "typical" daytime conditions, since in those polar regions they never actually have typical daytime conditions in midwinter but remain in dusk and dawn periods all day long!

I've mentioned before that during typical daytime conditions, when the D layer ionization is very intense, low-angle signals will be totally absorbed, while high-angle signals will get through and be refracted in the E layer (160 and 80 meters). Only at peak ionization, just after noon, may the absorption be noticeable on high-angle signals. The signal strength of local stations, received through ionospheric refraction, will dip to a minimum just after local noon. As stated before, to obtain good local coverage on 80 meters during daytime you must have an antenna with a high vertical angle of radiation (the high angle also means the wave will return to Earth not far from the transmitter). This can easily be obtained with a low 80 meter dipole. On 160 meters, many believe middle-of-the-day propagation is essentially limited to ground-wave signals. On the contrary, theory and actual QSOs show that stations running 1000 W can communicate over distances up to 1500 km around local noon (Ref 184). On the opposite end of the low-band spectrum, 40 meters basically stays open for DX almost 24 hours per day in winter, albeit with much-attenuated signals around local noon.

2.4.2. Nighttime (Black-Line Propagation)

After sunset, the D layer rapidly (within minutes) dissipates and almost completely disappears. Watch out — when the sun sets in the west, locations to the east may have been in darkness some time already, and D layer attenuation toward the east will consequently be reduced some time before your sunset.

Good propagation conditions on the low bands can be expected if both ends of the path, plus the area in between, are in darkness. The greatest distances can be covered if both ends of the path are at the opposite ends of the darkness zone (both located near the *terminator*, the dividing line between day and night). During nighttime in a period of low sunspot activity, the critical frequency may descend to values below 3.7 MHz and dead zones (skip zones) will show up regularly. Skip zones are also common on 40 meters during nighttime. Skip zones due to MUF do not occur on 160 meters since the MUF is always higher than 1.8 MHz. In contrast to gray line propagation, Brown, NM7M, calls propagation with one of the stations at the terminator "dark line propagation" (Ref 140).

2.4.3. Midway Midnight Peak

North-south (±30°) paths exhibit a clear propagation peak at local midnight time halfway on the path, both on 80 and 160 meters. When I make schedules on these bands with African or Indian Ocean stations, I will always try to have them at "midway midnight."

Although it has been generally accepted that an east-west path only exhibits a sunrise and a sunset peak, many critical observers have witnessed (at least on 160 meters) a similar halfway midnight peak. I have observed that this is especially true during high sunspot years, when the gray line enhancement seems to be less common than during low sunspot years. This peak certainly does not exist on 40 meters, where the signal peaks are only before and around sunset, and around or after sunrise. Although I have observed this phenomenon several times, it was Peter, DJ8WL (SK), who raised the question on the Internet. The exact mechanism may not be understood, but it probably is connected to the fact that the sun is exactly "on the other side" of the Earth, creating ideal ionization conditions in the E and F regions of the ionosphere on the dark side of the Earth. I will explain how to calculate these peak times, using sunrise-sunset times in Section 5.1.

2.4.4. Dusk and Dawn

As mentioned before, the terminator is the dividing line between the half of the Earth in daylight and the other half in darkness. The visual transition from day-to-night and vice-versa happens quite abruptly in the equatorial zones and much more slowly *near* the polar zones.

The *gray line* is a band between day and night, similar to but *not* the same as the *twilight* zone. There are three twilight zones: in the *civil twilight* zone the center of the sun is between 6° and 0° below the horizon; in the *nautical twilight* zone, these angles are between 12° and 6° below the horizon; and for the *astronomical twilight* these angles are 12° and 18°. The twilight zone does not extend beyond sunrise; the gray line zone does stretch on *both sides* of the exact terminator. Actually, *gray zone* might have been a more appropriate term than gray line.

The period around dusk (sunset) and dawn (sunrise) produces very interesting propagation conditions that are not limited to only the low bands. However, the mechanisms involved can differ very substantially between high bands (10, 15 and 20 meters) and the low bands (40, 80 and 160 meters).

For low-band operators in particular, it is extremely important that they be able to visualize the situation using maps or globes that show the terminator, the great-circle lines and the auroral oval (See Section 4.1.). For a long time many have speculated about what actually produces the enhanced propagation conditions we experience very often on the low bands on E-W paths at either dusk or more frequently at dawn, especially during low sunspot years. It has become widely accepted that these twilight effects are due to the onset of a specific propagation mechanism — one that is characterized by lower loss than the standard multi-hop model. The *ducting* mechanisms involved are discussed in detail in Section 2.4.4.4.

But besides the role of the ducting mechanism at dusk and dawn, there is another reason why we are able to work DX much better during twilight periods. When the sun is rising in the morning, all signals coming from the east (which can often cause a great deal of QRM during the night) are greatly attenuated by the D layer existing in the east. The net result is often a much quieter band from one direction (east in the morning and west in the evening), resulting in much better signal-to-noise ratios on weak signals coming from the opposite direction. This does not necessarily mean that our signals will be received any better, since the better signal-to-noise ratio obviously only influences receiving and not transmitting. This also explains that, if sunrise on the eastern side of the path coincides with sunset on the western end, we have a double enhancement, similar to what would happen if both stations were using receiving antennas with very high directivity.

It is also important to know how long these special propagation conditions exist, in other words, to know how long the effects of the radio-twilight periods last. It is clear that the rate of change in ionization of the D layer depends upon the rate of sunrise (or sunset). There are two factors that determine this rate: the season (the sun rises faster in summer than in winter) and the latitude of your location (the sun rises very high near the equator, and peaks in the sky at low angles near the poles).

It is also clear that propagation where the signals depart or arrive at an angle *perpendicular* to the terminator will enjoy the least signal attenuation, since these paths travel the shortest distance through the attenuating D layer and the bottom of the E layer (which also causes attenuation). Carl, K9LA, reports that while a multi-hop path is limited to roughly 10,000 km in total darkness (see also Section 2.1.3.), extensive ray-tracing exercises have proven that propagation *in* the gray zone is limited to half this distance (5000 km), all other parameters being the same. This validates what we already know: the best place for 160 meter propagation is in the dark ionosphere (Ref 169). For 80 meters the maximum distance for a propagation path along the terminator roughly 9000 km. Here too, the absorption in the lower E layer region limits the distance a signal can travel and still be heard. In mid winter we have such a morning path to Japan along the terminator on 80 meters. I have never seen that happening on 160 meters, though (see Section 4.3.3.1.)

2.4.4.1. The Gray Line or the Gray Zone on the Low Bands

When both ends of a path are in close proximity to the terminator, one side at sunrise and the other at sunset, then we have a so-called *gray line* situation. However, the term gray line propagation is also used where only one side of the path is located near the terminator (usually at sunrise at the eastern end of the path).

The effect of advantageous propagation conditions at sunrise and sunset has been recognized since the early days of low-band DXing. Dale Hoppe, K6UA (SK), and Peter Dalton, W6NLZ (SK), first used the term *gray line* for the zone centered on the geographical terminator (Ref 108). See Fig 1-15.

In the past, some authors have shown *the gray line zone* as a zone of equal width all along the terminator. All map programs (eg *DX Atlas*) showing gray line do so, because what they actually show is the *twilight zone*, and not the *gray line zon*e. This could potentially lead to confusion. The gray line zone is not the same as the twilight zone. R. Linkous, W7OM, recognized that the zone width varies in his excellent article "Navigating to 80-meter DX" (Ref 109).

The duration of enhanced propagation around sunset and sunrise, in other words the width of the gray zone, is heavily influenced by the latitude. In the Arctic and Antarctic regions, the gray line will last several hours (if the location considered is not in permanent darkness at winter solstice). At the equator the switch from day to night and vice-versa takes as little as 20 minutes. This is because at low latitudes the sun's apparent movement is perpendicular to the observer's horizon; in other words the sun rises at an angle of 90° with the horizon. As one gets closer to the Arctic and Antarctic circles, the sun rises at a very low angle with the horizon. This means that at high latitudes it will take much longer to pass through the twilight zone. Above approximately 70° the gray zone lasts day and night around midwinter!

All of this explains why we have a narrow gray line (zone) near the equator and a wide gray line near the poles. The time span during which we will benefit from typical gray line conditions will accordingly be shorter near the equator and longer in the polar regions. Therefore the gray line phenomenon is of less importance to the low-band DXer living in equatorial regions than to his colleagues close to the polar circles. This does not mean that there is less enhancement near the equator at sunrise or sunset; it just means that the duration of the enhanced period is shorter.

Some authors (Ref 108 and 118) have mentioned that gray line propagation always happens along the terminator. On the low bands there have only been occasional instances of such propagation. From the following examples of gray line propagation, it should be clear that propagation almost never happens along the gray line but rather through the dark zone, on a path that is in most cases nearly *perpendicular* to the terminator at both ends of the path. In the zone along the gray line there is substantially *more* attenuation due to the absorption in the D layer (and in the lower E layer region). Gray line propagation on the low bands is a different affair from what often is called gray line propagation on the HF bands, where the propagation path does follow the direction of the gray line.

W4ZV wrote: "Here is what I have observed many times for what I call 'long-path modes': SSW before sunrise and SSE after sunset. Signals on these paths typically peak at midway between sunrise/sunset times at each end of the path, and appear to be optimum when there is approximately 40 minutes of common darkness for 80 meters, and approximately 80 minutes of common darkness for 160. These paths are doing something different since the arrival and departure azimuths are nearly aligned parallel to the terminator at each end of the path. Here is an example: **users.vnet.net/btippett/dx_aid_plots.htm**."

This does not at all mean that the propagation goes inside the gray line however! In all similar contacts that Bill shared the details of, it appears that both his station as well as the Far East station were way out of the gray line zone at the time of the contact.

QSO	*Date*	*UTC*	*Common Darkness (minutes)*	*% into Common Darkness*
Sunrise Paths				
UA9UCO - WØZV	29SEP87	1232	59	61
JJ1VKL/4S7 - WØZV	28DEC91	1335	109	56
3W5FM - W4ZV	06JAN00	1143	107	55
XZ0A - W4ZV	21JAN00	1156	62	50
XU7ACB - W4ZV	05DEC01	1119	93	40
Sunset Paths				
9V1XQ - W4ZV	13JAN96	2309	42	50
S21XX - W4ZV	04FEB97	2320	106	26
Average			83	48

I would certainly not call these "classical" long path QSOs (see Section 4.2), but a crooked path where the arrival and departure azimuth are generally along the direction of the gray line, a direction less than 90° off the real long path direction (the direction opposite to the short path). See Chapter 2, Figs 2-23 and 2-24. **Fig 1-17** shows the daylight-darkness situation of the QSO between W4ZV and XU7ACB.

Gray line propagation occurs right around sunrise or sunset or both combined. On 40 meters the morning peak is always after sunrise. On 80 meters it is from sunrise to shortly after sunrise (typically up to 30 minutes in winter time), and on 160 meters typically shortly before sunrise until sunrise or just a few minutes after sunrise. These figures are for locations at average latitudes (30 to 50°). These sunrise/sunset peaks are more pronounced during low sunspot years than during years of high sunspot activity. Usually the sunrise peak is much more pronounced than the sunset peak.

2.4.4.2. Examples of Remarkable Gray Line Propagation

Many of us remember the unforgettable DXpedition to Heard Island in January 1997 (at the bottom of the sunspot Cycle 22). K9LA (Ref 152) described how US East Coast stations, against all expectations, worked Heard Island on 160 meters day after day, on what was considered to be an extremely difficult path. **Fig 1-18** shows the theoretical great-circle path between Heard Island and New York (mid January at 2300 UTC). Note that the path (heading of 250°) makes an angle of approximately 30° with the terminator on Heard Island. **Fig 1-19** shows the same path, with New York as the center of the azimuthal projection map. This path (heading of 130°) makes a similar sharp angle (25°) with the terminator at the US end of the path. Most, if not all, US East Coast stations who worked VKØIR and who had access to a variety of directive receiving antennas (such as Beverages) noted that the VKØ signals arrived at a heading of approximately 60°, right across Europe. (See also Section 4.3.)

Fig 1-20 shows the path between Heard Island and Spain. The path (beaming 300°) now makes a perfect 90° angle with

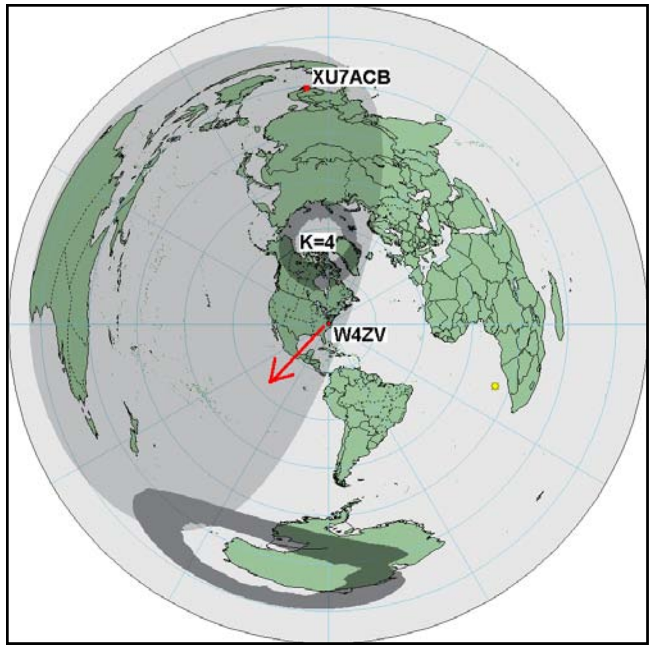


Fig 1-17 — QSO between W4ZV and XU7ACB on December 5, 2001 at 1119 UTC (K=4). Both stations are approximately 40 minutes into darkness, so they are not in the gray zone which on 160 meters is much narrower than that for the latitudes concerned. The auroral oval for K=4 is included, clearly showing that the direct path via the north is all but impossible. (Map generated with *DX Atlas*, with additions by ON4UN.)

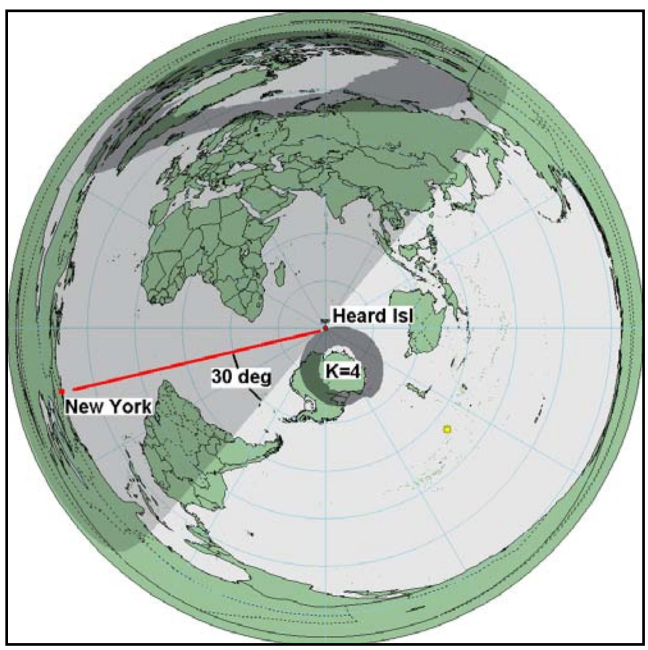


Fig 1-18 — Great-circle short path from Heard Island to New York (early January, 2300 UTC, K=4). The angle compared with the terminator at Heard is quite sharp. (Map generated with *DX Atlas*, with additions by ON4UN.)

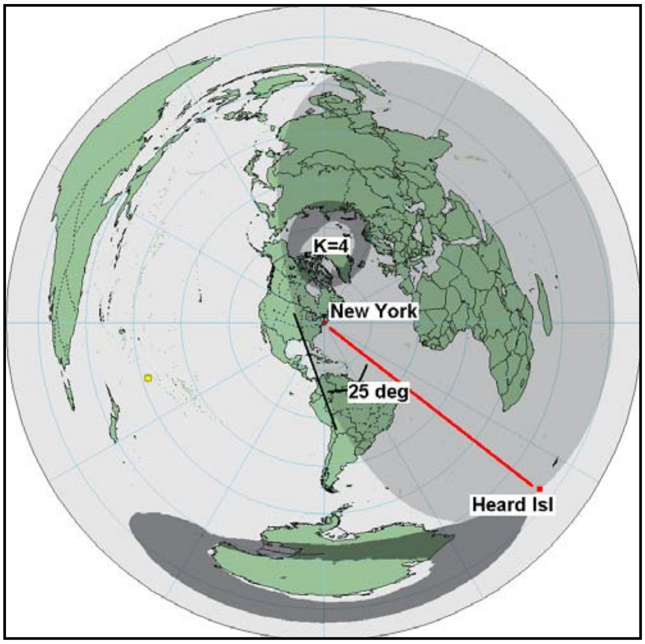


Fig 1-19 — At the other end of the same path, you also see also a sharp angle (25°) compared to the terminator. (Map generated with *DX Atlas* software, with additions by ON4UN.)

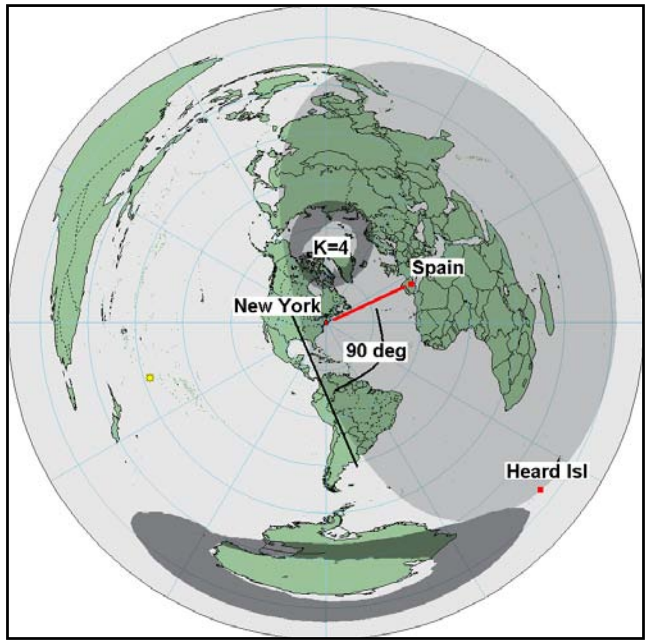


Fig 1-20 — The path from Heard Island to Spain makes for a perfect 90° angle with the terminator, as it does with the bent path from the US East Coast, over southern Europe, to Heard island. US East Coast stations reported that the Heard Island signals came in a path over Europe, at least for the first days of the DXpedition. (Map generated with *DX Atlas* software, with additions by ON4UN.)

the terminator on Heard Island. Similarly, the path between the US East Coast and Spain (beaming 65°) also makes a perfect 90° angle with the terminator at the US end of the path (**Fig 1-21**). I am convinced that the signals at Heard Island traveled across Europe to the US East Coast. This is supported by testimonies from US stations and it again confirms that enhanced gray line conditions most often go together with a signal azimuth that is nearly perpendicular to the terminator.

K1GE confirmed that this has happened with other stations from the Indian Ocean as well. I was listening every day during the Heard Island DXpedition and witnessed that the signals faded out completely in Europe at exactly the same time they faded out in North America. This seems to confirm that, indeed, the path to the US was right across Europe. What makes this path skew to more northerly regions is explained in Section 4.3.1.

To be fully correct, I must admit that a QSO between the US East Coast and Heard Island at 2345 UTC is only half a gray line QSO. However, stations a little further inland in the USA who worked VKØIR just prior to that did it on a double-sided gray line path.

The VKØIR expedition to Heard Island was a living testimony to how the width of the gray line depends on the latitude of the station involved. Many remember how VKØIR (located at 53° south latitude) was worked almost every day on 160 meters until more than 30 minutes after local sunrise, while 80 meter QSOs were made as late as 0050 UTC, which is 1½ hours after sunrise on Heard Island during that DXpedition.

Another striking example of gray line enhancement involved a QSO I had on 80 meters with Kingman Reef, a particularly difficult path late in the Northern Hemisphere winter season from Europe. I made QSOs with Kingman Reef and Palmyra around May 1, 1988. If we analyze sunset and sunrise times for that date, we see that sunset on those islands

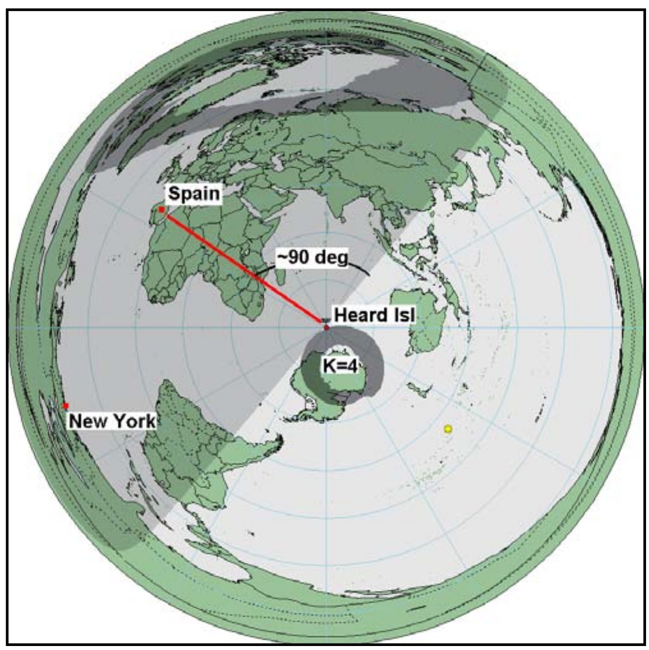


Fig 1-21 — The crooked path from Heard to the US East Coast also was launched on a path at right angle to the terminator at Heard Island toward Europe where it seemed to be bent back toward the US East Coast. (Map generated with *DX Atlas* software, with additions by ON4UN.)

is roughly 40 minutes after sunrise in my location. This means that there is theoretically no opening, but we can force things a little and take advantage of the gray line: split the 40 minutes in half and try a QSO 20 minutes before sunset in Palmyra (or Kingman Reef) and 20 minutes after sunrise here in Belgium.

Does this sound like a nice 50/50 deal? Certainly not! Those Pacific islands are situated only about 6° north of the equator; Belgium is 51° north. This means that the gray line lasts just seconds out on KH5 and maybe 40 minutes in Belgium. I made the schedules right at Pacific sunset time. On Kingman the QSO was made five minutes after Kingman sunset, on Palmyra four minutes after sunset, and in both cases 40 to 45 minutes after sunrise in Belgium (where the gray line is fairly "wide"). This is a striking example of how knowledge of the mechanisms involved in propagation can help you make a very difficult QSO. Here's proof of how marginal a situation it really was: From Palmyra only one QSO was made with Europe on 80 meters, and only two from Kingman Reef.

Over shorter paths, stations can often be worked on 80 meters for hours after sunrise (or hours before sunset). Contacts between the US East Coast and Europe are quite common in midwinter as much as two hours after sunrise in Europe during low sunspot years.

The width of the gray line on 160 meters is much more restricted than on 80 meters. Even in the middle of the winter, I have seldom worked real long-haul QSOs more than 15 minutes after sunrise on Top Band (there are remarkable exceptions though). I often copy stations quite late after sunrise, but with weak signals on a very quiet band, which is possible because of the absence of noise from the east. This phenomenon has been confirmed many times by Jack, VE1ZZ, who can copy European stations on 160 meters up to three hours before his sunset. This does not mean he was able to contact the many European stations he heard. For that he had to wait almost two hours! Anyhow, at that time of the day the European stations are probably listening to the east. VE1ZZ reported that signals were quite weak and only copiable because there was absolutely no noise whatsoever when this happened.

At this time in the afternoon toward the daylight area, the D layer acts as a shield to ionospheric-propagated noise. This provides a much lower noise floor for the station in sunlight. The stations at the eastern end of the path are subject to high noise levels because they do not have the advantage of the D layer absorption attenuating the atmospheric noise propagated into their area. Just watch your S meter on 160 meters at 2 PM during winter and then again after sunset, and you will probably see a 20 or 25 dB higher atmospheric noise level, provided you are in a quiet location with little manmade noise. A spread of 20 dB is difficult to overcome. This explains the one-way propagation under such circumstances. At the same time, in Europe, there is no D layer screen and signals from the east are 50 dB stronger than Jack's signal before his sunset. This one-way propagation is quite common during quiet magnetic conditions (low A-Index and K-Index). Similar observations have been made from Europe, where UA9/Ø stations are heard long before European sunset, but again, most of the time we have to wait until almost sunset to make contact. I recall one exception during the CQ 160 meter contest in January 2008, when I worked JH4UYB at 1545 UTC, approximately 45 minutes before my sunset. But this is quite unusual.

On 160 meters there are rare times when I can work DX well after sunrise from my location in Europe. This happens when I can work ZL stations on a genuine long path (21,500 km) about 30 to 45 minutes after our sunrise during low sunspot years (around equinox). The few ZLs worked on this long path had to be worked right through a wall of English stations, who were enjoying their sunrise peak at exactly that time.

Sometimes, when conditions are really good with no atmospheric noise, signals can also be heard a very long time after sunrise at the eastern end of the path. GW3YDX reported copying many W6/W7 stations as well as KL7 on 160 meters until more than one hour after local sunrise. Really exceptional was the fact that he copied K6SE at 1130 UTC, three hours after local sunrise! That same day, GM3POI reported hearing KL7RA at 1230 UTC, also on Top Band (in midwinter)! We have, of course, to be careful and not extrapolate these observations to the whole of Europe: the few stations that reported these extraordinary conditions are located at the fringes of Europe, almost at the back door of North America, and quite far north (52 to 53° N).

To me, it is clear that these exceptional QSOs or reports are happening thanks to a *bent (crooked) propagation path* (see Section 4.3.). The signals appear to travel near the North Pole, staying in darkness as long as possible. W8LT has reported the same experience on 80 meters, working EI and G stations between 1000 and 1100 UTC. He confirmed that the best antenna for this propagation was a half-square broadside N/S. To him, this indicated that a crooked path was involved. I believe that this kind of propagation can only occur during a short period around winter solstice and when magnetic conditions are exceptionally quiet, with resulting extremely low auroral-zone absorption.

Similar conditions are quite common on 40 meters during the European winter. With a good Yagi antenna, I can work North America 24 hours a day on 40 meters, when there is no geomagnetic disturbance. At local noon, when the sun is highest, I hear W8s and W9s quite commonly, followed somewhat later by W6 and W7/VE7 stations, all on a direct polar path. The West Coast will keep coming through with the beam pointed approximately 350°, until at 1430 UTC the band will also open on the long path. Shortly later the bent short path will close.

These specific propagation paths and times are applicable only to moderate-distance DX when it comes to 160 meters. Really long-haul propagation on 160 seems to follow the rule of enhanced propagation only occurring at dawn/dusk. During a long period of tests (in November and December) on 160 meters between New Caledonia (FK8CP) and Belgium (distance: 16,000 km) I found that his signals always peaked right around sunrise (from three minutes before, to three minutes after sunrise). This "short peak" is valid only for the very long path to Western Europe. Because of the distance involved we no longer speak about standard multi-hop propagation but about signal ducting inside the ionosphere (see also Section 2.4.4.4). FK8CP reports openings into Asia (UA9, UAØ) from much earlier, until a little later after sunrise.

Close-in DX (2000 km or 1500 miles) can be worked as late as 45 minutes after sunrise on 160 meters, again depending on your latitude. FK8CP reports working DX in the Pacific as late as 50 minutes after his sunrise in the middle of his local summer (which is quite late, considering the latitude of New Caledonia).

On 40 meters, the gray zone is of course "wider" (lasts longer) than on 80 meters (remember the gray line is a zone of variable width, not to be confused with the terminator, which is a line uniquely defined by sun-Earth geometry). In the winter, long-haul DX can be worked on 40 meters until many hours after sunrise (or many hours before sunset), again depending on the latitude of the station concerned. For example, stations at latitudes of 55° or higher will find 40 meters open all day long in winter. Even at my location (51° north), I have been able to work W6 stations at local noontime, about three hours into daylight. At the same time, the band sometimes opens up to the east, so we can say that even for my "modest" latitude of 51° N, 40 meters is open for DX 24 hours a day on better days.

2.4.4.3. Propagation Without Lossy Ground Reflections?

Multi-hop propagation with intermediate ground reflections has long been the traditional way to explain propagation of radio waves by ionospheric refraction. Not too long ago some scientists stated that ducts did not exist and that all propagation had to be by means of multiple hops from the Earth to the ionosphere and then back again. In the last 20 years enthusiastic low-band DXers have made literally thousands of observations of propagation "anomalies" that yielded much stronger signals than could be calculated.

This mass of observations spurred propagation scientists to take a closer analytical look at the available data. Theoretical research enables scientists to calculate path losses due to ionospheric absorption, free-space attenuation (path-distance related) and earth (ground or water) reflection losses. The ionospheric refraction loss on 160 meters during the night is approximately 11 dB per hop (about 4 to 5 dB less on 80 meters), excluding ground reflections (Ref 169).

While the theory of propagation with ground reflections and the knowledge of the attenuation involved at each step is satisfactory to explain short- and medium-range contacts on 160 meters (maximum of about 10,000 km), the losses through ground reflections and ionospheric (E layer) losses are no longer accepted by most experts as adequately explaining some of the high signal levels obtained over very long distances, especially when gray line propagation and genuine long-path situations are involved.

On 160 and 80 meters, when the entire path is well into darkness, propagation is primarily by means of multiple hops in the F layer, with little or no D layer absorption, but with additional attenuation due to the remains of ionization in the E layer. This seems to be the model that fits observations when neither end of the path is in the twilight zone.

Even this model does not explain why there generally is a significant signal enhancement where either or both ends of a very long distance path are in the twilight zone (more specifically, when the eastern end of the path is at sunrise). In the *gray line* zone there should actually be *more* D- and E layer absorption, compared to the full-darkness situation. In fact, during low sunspot years, gray line enhancement is more prominent compared to high sunspot years, because the ionization of these attenuating layers at sunrise and sunset during high sunspot years is more pronounced.

Thus there must be another mechanism involved that compensates for the additional D- and E layer losses in the twilight zone, a mechanism where we can happily end up with an overall loss that is significantly smaller than the full-darkness, multi-hop model with little or no D- and E layer absorption. *Signal ducting* could well be that mechanism behind twilight-zone enhancement.

Sometimes, signals are ducted as though they were confined inside a pipe (waveguide) in the ionosphere, without lossy intermediate reflections from the Earth's surface. Such ducting explains strong signals sometimes heard over very long distances. Gray line enhancement seems to go hand-in-hand with ducting (see Section 2.4.4.4) and this is more pronounced during low sunspot cycle years. Recent propagation-prediction software tools (Ref 153) include models that support three-dimensional ray-tracing and ducting, including geomagnetic effects. *Proplab-Pro* is one such program that explicitly computes 80 and 160 meter ducting modes for many long-distance paths.

I first came across a description of the phenomenon of ionospheric ducting in an article by Yuri Blanarovich (K3BU) in 1980 (Ref 110). More than 20 years later the phenomenon of signal ducting (ionospheric ducting) on the low bands finally seems to have been accepted also by the scientific community.

I have to admit, however, that the multi-hop-only model without ducting can help to explain why paths that are across saltwater generally produce stronger signals, due to the minimal reflection losses at saltwater reflections. (Of course, paths can include combinations of Earth-ionosphere multiple hops, as well as in-ionosphere ducting mechanisms.)

Dan Robbins, KL7Y (SK) worked for years in the field of HF radar. He maintained that HF signals really do bounce off the Earth, and that the losses on Earth reflections (especially from saltwater) are mostly insignificant — usually much less than the losses due to ionospheric refraction, at least for frequencies below the MUF.

Textbooks, including *Ionospheric Radio* by Davies, provide charts that indicate that sea-water reflection loss is a fraction of a decibel on 160 meters for all but the lowest angles (3° or less). A land reflection might typically average several dB, so it is easy to see why long paths over water can produce stronger signals on the low bands compared to paths that require multiple reflections over poor ground. This is one area where I believe many propagation programs fail, since they do not know the geography at the reflection point. They plug in an "average value," something like 2 or 3 dB. On an all-water long path with multiple ground reflections the program could be off by 10 to 20 dB. One of the exceptions is the *Proplab-Pro* ray-tracing program mentioned above, which contains an Earth/water/ice geographical database that is used to compute realistic reflection losses.

2.4.4.4. Ionospheric Signal Ducting

Based on experimental observations (Ref 100) and theoretical studies (Refs 131 and 151), researchers have come to the conclusion that some very specific ducting modes were allowing exceptionally strong signals to be heard over very long paths.

Due to the layered structure of the ionosphere, waveguide-like channels (ducts) appear in which radio waves can propagate over long distances. **Fig 1-22** shows the nighttime electron-density distribution, showing a dip in electron density above the E layer peak. This valley is responsible for setting up a waveguide-like 160 meter duct, bounded by the F-layer at the top and the top of the E layer on the bottom (Ref 151). Top Band signals get trapped between the E and F regions

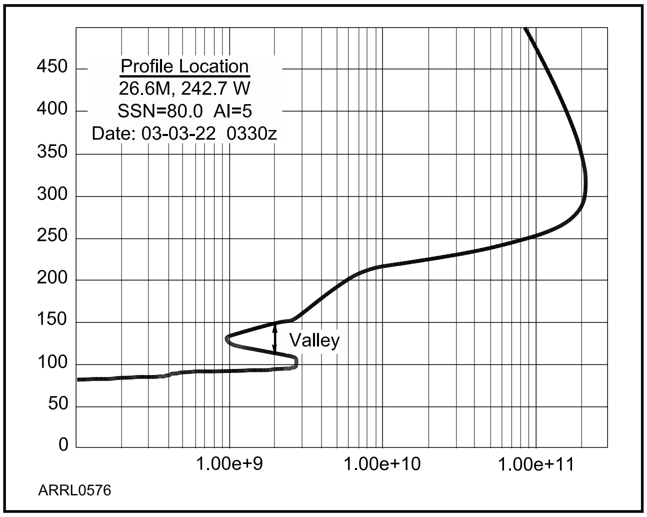


Fig 1-22 — Electron-density profile for a point near Iceland, roughly halfway between Europe and North America on the North Atlantic path. Notice the dip in the electron density between the E and the F layer. This dip is responsible for waveguide-like propagation (ducting), with the bottom-side of the E layer and the topside of the F layer serving as "duct walls."

rather than propagating between the E-region or the F-region and the ground. The trick is to get the signal to enter this duct, and then later to get it out of the duct at the end of the path.

Cary Oler and Ted Cohen (Ref 142) point out that this kind of E-F ducting is most typical for 160 meters because Top Band signals can be refracted more effectively at higher wave angles than signals at higher frequencies can be. From the launching point of such a duct path, refraction occurs when the signal travels through the E layer, resulting in bending of the waves into a lower angle (in other words, there is not a complete reflection). The wave is propagated further at the required angle to start the ducting. Relatively high launch angles from the Earth are required to punch through the D and E layers, so that the wave can finally be reflected up into the E-F region. This may be another explanation why higher wave-angle transmit antennas sometimes (mainly around sunrise/sunset) beat out very low-angle antennas, especially on 160 meters. (See also Section 2.1.2.)

The transition from darkness to daylight causes what is known as *ionospheric tilting*, which can also help bend a signal into (and out of) the duct. In addition, horizontal ionospheric tilting is often required to be able to work into a duct. This condition typically exists at the terminator, provided that the angle between the propagation path and the terminator is close to 90° (Ref 151).

Once inside, how do the signals leave a duct? If the other end of the path is in the gray zone, the local slope of the electron-density contours can alter the local refraction angle of the duct, and the signal angle becomes steep enough to break through the E region back down to Earth. This mechanism favors the relatively high elevation angles often observed for such received signals. The exit of the signal from the duct often produces a spotlight-like illumination of the Earth, making signals very strong in a specific location, while being inaudible at another location only a few hundred miles away.

Why is ducting not an everyday occurrence? The ionosphere is particularly turbulent around sunrise, so the well-defined contours of electron density required for signals to leave the duct are not necessarily there every day. Nick Hall-Path, VE7DXR (Ref 177), summarizes the situation as follows: "I suggest that sunrise enhancement could be caused by ionospheric tilts occurring just before sunrise at the receiver. Those tilts direct signals to the receiver, from a duct between E and F regions. The D-region would tend to absorb such signals rather quickly, however, and no good answer has been offered as to why enhancements occur on some mornings and not on others. Perhaps on mornings when strong signals are heard, there is some retardation of D layer absorption caused more by terrestrial air movements than by solar and geomagnetic influences." This supports the suggestion by R. Brown, NM7M, indicating the importance of the ozone layer in this mechanism (Ref 176).

2.4.4.5. The Influence of Sunspot Cycle on Ducting

Higher sunspot numbers means more ionization. How much more? Brown (Ref 170) reports that going from an SSN of 5 to 100, the electron density at the bottom of the E region increases only modestly, by a factor of merely two. This results, however, in an increase of the critical frequency of the E layer of approximately 30%, which is significant. To penetrate the bottom of the E layer, the wave angle must now be steeper (again explaining why higher angles seem to do well near sunrise and sunset). This steeper angle results in shorter consecutive hops inside the duct, and consequently more loss on the path.

To sustain a ducting path between E layer peak and the F region, the valley between these two regions (see Fig 1-14) must be present all along the path. What could cause this valley to disappear? Scientists suggest that even modest increases in electron density in the ducting region will fill up the valley and halt the ducting mechanism. The required levels are levels that will barely increase signal attenuation. Even very modest levels of auroral activity can be disastrous to this mode of propagation, not only by the creation of extra absorption but mainly by stopping the duct itself.

From his earlier work, NM7M has stated that sources of ionization for the nighttime E valley are starlight, galactic cosmic rays and solar X-rays scattered by the geocorona. Those are listed in order of increasing strength. Starlight and galactic cosmic rays are obviously not directly related to solar activity, although cosmic rays can be affected by a geomagnetic field stirred up after a solar flare or a blitz from a coronal hole. Solar X-rays scattered by the geocorona, however, do increase with increasing solar activity. So the E valley tends to be lowest at times of low SSN and rises with high SSN. This means that there should be more ducting (and better Top Band DX propagation) in times of low SSN, which confirms our observations on the air.

Recent work by NM7M and K9LA ("A Theory on the Role of Galactic Cosmic Rays in 160 meter Propagation") has shed more light on the mechanism of ducting (Ref 183).

NM7M and K9LA now believe that the nighttime valley is maintained primarily by galactic cosmic rays. In essence, the valley wouldn't exist if it weren't for the weak ionization produced by galactic cosmic rays. With galactic cosmic rays out-of-phase with a solar cycle, there can be too much valley ionization at solar minimum, allowing mostly multi-hop propagation to extreme distances. Note that this view is opposite from NM7M's earlier view.

As a solar cycle increases, absorption increases thereby making extremely long distance multi-hop propagation less likely. But the valley ionization produced by galactic cosmic rays decreases, allowing the valley to form and better ducting to occur. Thus it appears that multi-hop propagation to extreme distances is prevalent at solar minimum, and ducting is more prevalent as we move away from solar minimum.

This was clearly demonstrated during the 2008-2009 winter, where SSN of approximately zero prevailed for many months. This resulted in very low geomagnetic activity (aurora), and often excellent propagation over the poles, and that up to approximately 10,000 km, the maximum distance that can be covered by multi-hop propagation. During this same period we however witnessed relatively poor "very long haul" propagation (over 10,000 to 11,000 km), propagation which is believed to happen only thanks to ducting.

Fig 1-23 hints at the correlation between a decrease in galactic cosmic rays (which is believed to result in a better valley) and QSO distance. More work is underway in this area to confirm and better understand the role of galactic cosmic rays in 160 meter propagation.

2.4.4.6. Chordal Hops

Others authors have pictured another very specific way of signal ducting called *chordal-hop propagation*. In this ducting mode waves are guided along the concave bottom of the ionospheric layer acting as a "single-walled duct." The flat angles of incidence necessary for chordal-hop propagation are possible through refraction in the E layer, and because of tilts in the E layer at both ends of the path. Chordal-hop propagation modes over long distances are estimated by some to account for up to 12 dB of gain due to the omission of the ground-reflection losses. Long-delayed echoes or "around-the-world echoes" witnessed by amateurs on frequencies as low as 80 meters can only be explained by propagation mechanisms excluding intermediate ground reflections.

2.4.4.7. High Wave Angles at Sunrise/Sunset

Hams generally accept the notion that low radiation angles are required for DX work on the low bands. Those who can choose between a low-angle antenna and a high-angle radiator (a low dipole, for example) confirm that 95 to 99% of the time, the low-angle antenna is the better one. But there are the occasions where the low dipole will be the winner. This only seems to happen during the gray line period (dusk or dawn) though, and more specifically, after sunrise. W4ZV wrote: "I very often saw post-sunrise conditions favor the inverted-V, even though pre-sunrise almost always favored the vertical. Usually the vertical was about 10 dB better before sunrise, then they would both be equal at exact sunrise, then the inverted-V would be 10 dB better. I believe the post-sunrise peak is high-angle for the following reason. Beverages and verticals are both low-angle antennas and are therefore very complementary. After sunrise, in addition to the vertical being down on transmit, the Beverages would become poor for DX stations (but still good for local USA which was probably still low-angle). I remember when I first worked YBØARA (near Jakarta before he moved to /9 in Irian Jaya) well after sunrise. He was perfectly readable on the inverted-V but was inaudible on any Beverage."

In the 1960s, Stew Perry, W1BB, speculated that at sunrise and sunset the ionosphere acts like a big wall behind the receiving or transmitting location, and it focuses the weak 160 meter signals like a giant, poorly reflective dish on one area at a time just ahead of the densely ionized region in sunlight. This seems like an acceptable explanation, since it appears that losses at low-incident (grazing) angles are very high near the LUF (Lowest Usable Frequency) of a path. This may also explain why very often at sunrise and at sunset high-angle antennas seem to perform better than low-angle antennas.

Another now generally accepted way of explaining the fact that high-angle antennas often have the edge at sunrise/sunset is that a high-angle signal on its way to the F layer can punch right through the absorbing E layer. A lower-angle signal spends too much time passing through the D and the E layer and hence it suffers increased absorption. In other words, over the same distance, a two-hop, high-angle signal can be considerably stronger than a single-hop, low-angle signal due to the effects of the D and E layers.

During the very successful XZØA expedition in 1999 by far the best receiving results at sunrise (working into the USA) were obtained using a horizontal dipole only 6 meters high. Low-angle Beverages were much worse than this low dipole. Over 400 QSOs were made into North America with this receiving antenna, and that was *not* at a sunspot minimum!

Yuri Blanarovich (K3BU) wrote that during one of his Top Band operations from VE1ZZ's QTH: "I had an inverted V at 70 ft and Four Square vertical array, and I was able to crack the 'one-way afternoon' Europeans with the inverted V almost two hours before the Four Square was heard. These verticals have an ocean of radials under them and are sitting at the ocean shore on a small hill." This means the VE1ZZ vertical array can produce low radiation angles, which evidently was not what was needed under those circumstances.

G3PQA notes that his dipole (high-angle antenna) always outperforms his low Beverage (which has very little off-the-side, high-angle radiation) when working ZLs on 80 meters on long path at equinox, when there is a daylight gap

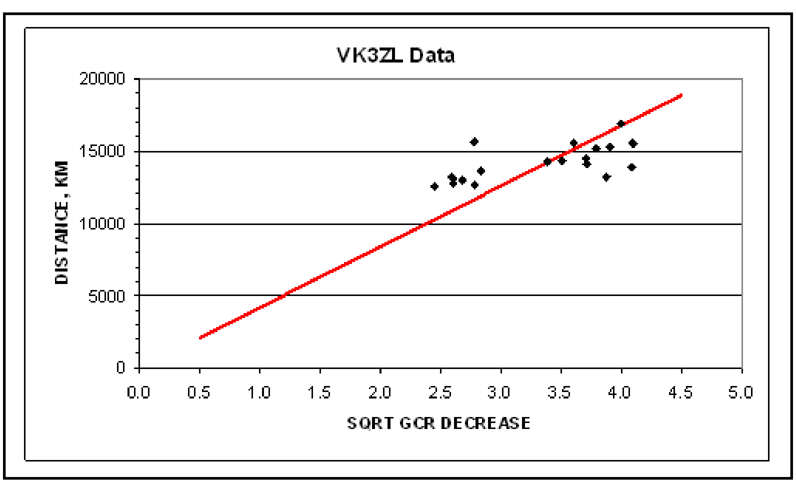


Fig 1-23 — A scatter plot of the distance of VK3ZL's 160 meter QSOs and the square root of the galactic cosmic ray decrease.

and when conditions peak after sunrise. This also seems to confirm that a high angle is required to pierce through the D and E layers to get into some sort of a ducting mechanism.

Similar testimony from Tim Duffy, K3LR, on the Top Band reflector: "I can confirm that a low ½ wavelength dipole is *very* effective on 160 and 80 meters. Mine are 20 feet high for both bands. Before sunset and after sunrise they are often the best RX antennas we have."

Tom, W8JI, who has a dipole at just over 30 meters high and another one 100 meters high, finds that both these dipoles sometimes beat the vertical array. In this case the 100 meter high dipole cannot be called a high angle antenna. Tom wonders if it is not really a matter of polarization (see Section 3.5) rather than angle of radiation.

2.4.4.8. Antipodal Focusing

Most low-band DXers know that, despite the fact that those are the longest distances you can encounter, it is relatively easy to work into regions near the antipodes (points directly opposite your QTH on the globe). By "near" I mean within a radius of approximately 1000 to 2000 km from the geographical antipodes.

In the past a mechanism of ray focusing in near-antipodal regions has often been used to explain the high field strengths encountered at those long distances, although the heterogeneous nature of the ionosphere makes it questionable. Multipath focusing would logically involve very deep fading due to different path lengths from converging signals, which is something that we do not typically witness on these "enhanced" signals coming from close to the antipodes.

But, if it isn't a focusing effect, what's the explanation?

We have an unlimited number of theoretical propagation paths to the antipode, all of equal length, spread over an angle of 180°, going over the dark side of the Earth. Instead of a*dding signals*, which is unlikely (different phase from different path causing deep QSB) the mechanism is maybe one of *selecting signals*. Let me explain.

Assume we have a transmit antenna that transmits equally well in all directions (a vertical). We know that, in order to obtain good signal strength at these long distances, we cannot live with multi-hop propagation but need to propagate our signals via ionospheric ducting (see Section 2.4.4.4.). We have learned that the conditions required for such a duct are rather critical and not always present. Now, if you have numerous theoretically possible propagation paths (as is the case with propagation to the antipode), chances are that for one of these path the conditions for onset of a duct in the ionosphere are better met than for other paths. Chances are definitely much higher for a good duct path than if we were stuck with a single propagation path.

Remember also that we needed very particular circumstances for the signal in the duct to be able to exit the duct, and that these are normally present in the gray zone near the terminator. Now: the antipode is the only place on Earth where the sun is setting at exactly the same time the sun is rising in your place *for each day of the year*. That means that we have real chances to have perfect duct entry and duct exit conditions all year long. In this enhancement concept we are not adding signals from different paths but rather selecting a direction that sustains the ionospheric ducting mode, set up along a real gray line path, with simultaneous gray line enhancement at both ends.

Most of us with access to directive receiving antennas have noticed that when enhanced propagation to (or near) the antipodes is involved, we do not receive the signals equally well from all over 180°, or over a very large angle, which is what we would expect if ray focusing were involved.

In my particular situation, New Zealand is about 1000 km from the actual antipode. On 80 meters the enhanced propagation conditions are very outspoken. The shorter path (19,000 km) is across Asia, beaming east from Belgium, the longer path (21,000 km) is to the west. While one might think that my evening path (into the east) would be the better one (it is shorter), it is not at all the case. Signals always *seem* much stronger on the longer path in the morning. Both are real gray zone paths. The difference is that in my evening the first 5000 km goes all across Europe, where a lot of noise and also radio signals are generated. In the morning the first 5000 km are water, and behind me (to the east), I have a wall set up by the D layer. This is why the morning path is much better for me.

Conclusion: there likely is no such phenomenon as ray focusing involved with the fact that signals coming from the antipodes or from near the antipodes are always better than from areas at a lesser distance but further away from the antipodes. The mechanism involved is likely to be only the mechanism of gray line propagation and signal ducting. A recent series of Propagation columns in *WorldRadio* by K9LA (Ref 185) laid the groundwork for an analysis of the alleged antipodal path between Colorado and Amsterdam Island appears to confirm this view.

3. PROPAGATION VS LOCATION

Working the low bands is very different, depending on whether you live near the arctic regions or near the equator. Let's analyze what causes these differences. In the previous sections, I have referred a number of times to the geographical location of the station. There is a close relationship between the time and the location when considering the influence of solar activity. Location is the determining factor in five different aspects of low-band propagation:

1) Latitude of your station vs rate of sunrise/sunset
2) Magnetic disturbances
3) Local atmospheric noise (QRN)
4) Effects caused by the electron gyrofrequency
5) Polarization and power coupling

3.1. Latitude of Your Location vs Solar Activity

This aspect has already been covered in detail in Section 2.3. The latitude of the QTH will influence the MUF (important on 40 meters and sometimes on 80 meters), the best season for a particular path and the width of the gray line zone.

3.2. Magnetic Disturbances (Aurora)

In his book *Aurora Australis* F. R. Bond wrote: "The aurora (Southern and Northern Lights) is mankind's only visible marker of the interactions taking place in the vast and complicated region of the Earth's magnetosphere." Brown, NM7M, stated: "… and Top Band propagation is another aspect of those interactions" (Ref 140).

Auroral absorption, most often evidenced by the aurora at high latitudes, is a very important factor in the long-distance propagation mechanism on the low bands. It is certainly the most important one for those living at geomagnetic latitudes of 60° or more, as well as for all of us living in more southerly regions when we are trying to work stations on paths that cross areas affected by the aurora. We are interested in what effect this phenomenon (which we hams mostly refer to simply as aurora) has on low-band radio propagation.

3.2.1. Auroral Absorption

Auroral absorption (AA) is very frequent and takes place in the D region due to the influx of very energetic auroral electrons. Visual aurora is due to such electrons that penetrate the ionosphere to altitudes as low as 70 to 100 km (which are heights normally associated with the D layer), and collide with mainly oxygen and nitrogen molecules. The ionization density of the affected areas in the ionosphere is very high and absorption of signals on 1.8 MHz can exceed 35 dB. Auroral absorption is relatively brief in duration, occurring during the times of visible auroral displays. Absorption regions tend to be elongated in longitude and narrow in latitude, just like the aurora display itself. They usually occur on the equator side of intensely bright visual auroral displays.

AA events are always accompanied by geomagnetic activity due to ionospheric current systems. Hence, the interest in the records of auroral-zone magnetometers for predicting times of low magnetic activity (or conversely, periods of high auroral absorption).

3.2.2. Coronal Mass Ejections (CMEs)

Sporadic outbursts of plasma, called Coronal Mass Ejections (CMEs), represent the release of considerable matter/mass from the sun's corona. They are the sources of blasts of solar wind that can disrupt the geomagnetic field, giving rise to auroral ionization and shutting down propagation on the low bands.

Only the plasma from a CME that goes out of the sun in the direction of the Earth may possibly hit the geomagnetic field and cause a magnetic disturbance. CMEs off the backside of the sun do not bother us, since they represent material ejected into space in directions that never can result in an encounter with the Earth.

A closely related phenomenon, a coronal hole, can also cause a geomagnetic disturbance. This essentially is a high-speed wind stream.

3.2.3. Aurora

The plasma coming from the solar corona is called interplanetary plasma. Magnetospheric plasma is plasma that is trapped within the Earth's magnetic field. The solar wind consists mainly of low energy protons and electrons. Magnetic activity here on Earth results from the impact of the solar wind on the magnetosphere.

The solar wind blowing by the Earth's magnetic field acts like a gigantic dynamo, where huge electrical currents are generated. This energy is often pent up in the Earth's magnetosphere. At times the energy is violently released, accelerating electrons in the tail regions of the Earth's magnetosphere. These electrons, since they are charged particles, are constrained to follow the magnetic field lines of the Earth. And since these field lines penetrate the Earth in the high-latitude regions, these electrons end up with trajectories that take them into the high-latitude ionosphere, where they collide with constituent particles and ionize the lower regions of the ionosphere. This process also releases photons of light, which we see as auroral activity. The increased electron density and disturbed ionization patterns contribute to increases in auroral absorption and can cause signals to begin experiencing multipathing and fading.

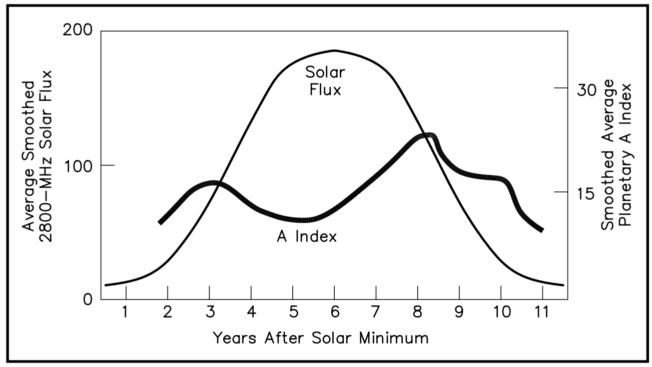


Fig 1-24 — This graph shows the geomagnetic activity (measured A-Index) as a function of the solar cycle. It appears that the geomagnetic activity is lowest during the upswing of the solar cycle.

So far as low-band propagation is concerned, the auroral belt at a height of approximately 100 km acts much like the D layer does during the day; it absorbs all low-band signals trying to go through the belt. Sustained periods of low auroral activity appear to be most common during the rising phase of the solar cycle. **Fig 1-24** shows the relation between the solar flux and the A-Index over a typical solar cycle.

The auroral belt is centered on the magnetic poles. The magnetic North Pole lies about 11° south of the geographic North Pole and 71° west of Greenwich. The magnetic South Pole is situated 12° north of the geographic South Pole and 111° east of Greenwich. The intensity of the aurora determines the diameter, the width and the ionization level of the auroral belt. At very low activity the auroral oval retracts to a major-axis dimension of approximately 3500 km, with a belt width of only a few hundred km. During a very heavy aurora the belt can grow to a major-axis dimension of more than 8000 km, with a belt width of more than 3000 km. Ionization in the auroral oval is usually not constant all the way around. In general the ionization is minimum at the local noon meridian and maximum at local midnight. Of course the oval is a statistical description and does not describe how the ionization is distributed or how energetic it may be. In other words, the local intensity of the aurora is not the same in all points of the oval and at all times of the day. It is obvious that the intensity at local noon is of very little interest to us low-banders since our signals typically propagate only in darkness.

The Earth rotates around the axis going through the geographic poles, while the auroral oval is more or less centered around the magnetic poles (see **Fig 1-25**). This means that the position of the often irregular-shaped oval changes position continuously with respect to the Earth rotating underneath it.

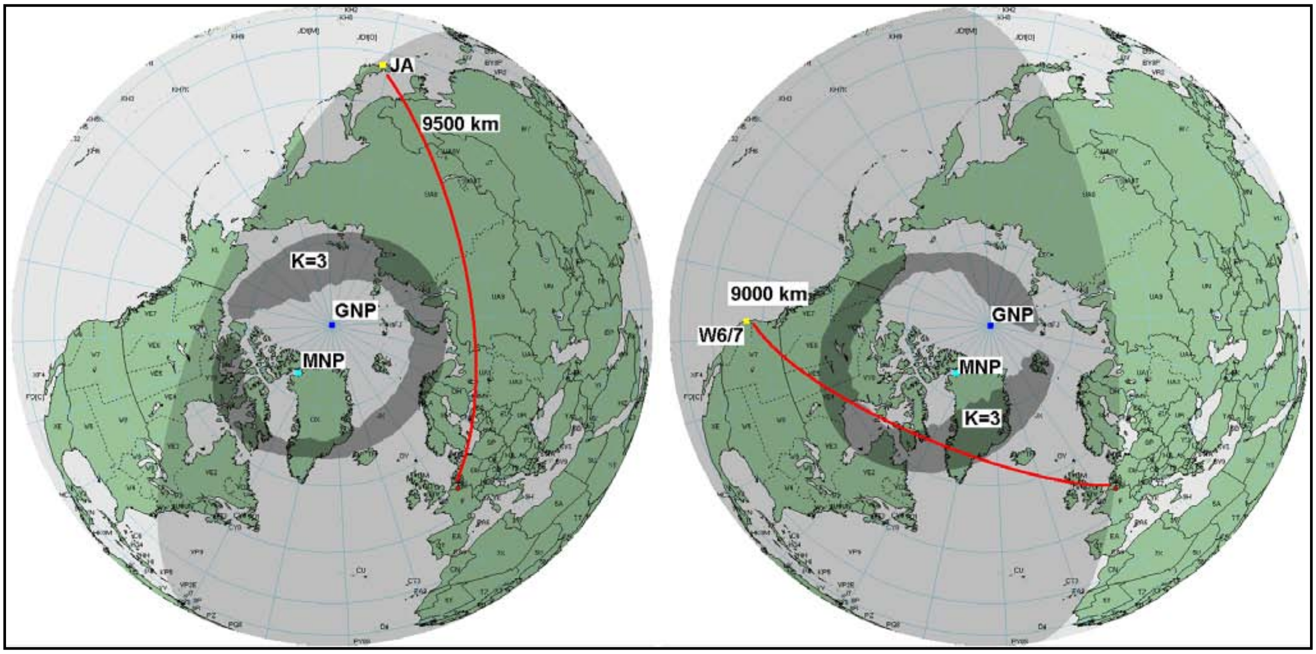


Fig 1-25 — "Top" view of the Earth showing the geographic North Pole (GNP) and the magnetic North Pole (MNP). On the left is the situation at sunrise in JA in mid-winter. On the right is the path between Europe and W6 at sunrise in Belgium. The auroral oval is shown for K=3. Notice that the auroral oval changes with time (minimum at local noon, maximum at local midnight). The path from ON to JA barely passes by the donut (for K≥6 the path goes into the donut), while the ON-W6 path goes for approximately 4000 km through the auroral donut.

3.2.4. Effects Caused by the Auroral Oval on Low-Band Propagation

The auroral belts (also called auroral ovals or auroral donuts) have a profound impact on propagation. If the low-band path over which you are communicating goes along or through the auroral oval, the result is usually degraded propagation caused by strong absorption of the signal. On the higher bands (20 meters and up) fast selective fading (multipathing) is a common sign of aurora. I have very seldom heard this on Top Band, and only infrequently on 80 meters, where these episodes always seem to be of short duration.

During exceptionally quiet geomagnetic conditions (K indices of zero for at least eight hours) the auroral zone might shrink to a major axis of about 40% as compared to when geomagnetic conditions are heavily disturbed, while its width might be reduced to a few hundred kilometers. The ionization levels in the shrunken oval can be extremely low during extended fully quiet conditions. Under such circumstances most polar paths will either pass along this small and almost undisturbed area and signals will suffer hardly any degradation.

During disturbed conditions, however, the auroral oval can very rapidly grow to an average size of some 8000 km. Under such conditions all paths that cross or touch this extended oval will be affected by severe absorption in the D and E regions and by other instabilities of the auroral ionosphere.

Cary Oler and Ted Cohen (Ref 142) state that when the auroral zone is contracted, it is possible for Top Band signals to pass through the auroral zone without suffering heavy absorption by skirting underneath the auroral oval. During periods of very quiet geomagnetic activity, the width (not the diameter!) of the auroral belt is only a few hundred km. On the other hand, radio signals reflected from the E layer can travel over distances of 500 to 2000 km through the lower atmosphere, on their way from or to Earth for a propagation hop. This means that with proper geometry, low-band signals can literally skip underneath and through the auroral zone into the polar ionosphere inside the auroral belt, where the ionosphere is more stable. They then continue from the polar ionosphere back into the ionosphere at latitudes below the auroral belt, without ever coming in contact with lossy region of the belt itself.

I have often found that propagation into or through polar regions favors the use of low-angle antennas much more than propagation into or across equatorial zones. I suppose this is so because low-angle hopping has more chances to skip underneath the auroral oval, hence suffering less attenuation than would be the case with higher-angle hopping.

Besides "undershooting" the auroral donuts, a common way for stations outside the oval to deal with aurora is to launch their signals in a non-great-circle route, called a "bent path" or a "crooked path" away from the auroral oval. A signal launched directly at the auroral oval will bend away from it because the enhanced ionization in the auroral zone creates horizontal ionization gradients. These horizontal gradients can refract signals in the horizontal plane (see Section 4.3.2.).

For stations inside the auroral belt, propagation to the world outside the belt is all but impossible once a geomagnetic disturbance has set in. It has been reported, though, that stations inside the auroral oval can hear quite well, but they do not seem to get out at all. For example, VY1JA experienced much frustration during the 1998 November Sweepstakes contest hearing strong stations that couldn't hear him. I suspect that the launch angle to skip under the oval was not right at VY1JA's end.

For stations just outside the auroral belt near the North Pole, usually the only (marginal) opening is directly to the south. Frequently, these stations seem to enjoy better propagation toward the equator than stations 1000 or 2000 km further south when the aurora is on.

On at least one occasion on 160 meters, I experienced propagation conditions similar to those on VHF during an extremely heavy aurora. Around 1600 to 1800 UTC on February 8, 1986, I heard and worked KL7 and KH6 stations on 80 meters, at the same time that auroral reflection was very predominant on VHF and 28 MHz. From Europe, this was on a path straight across the North Pole and the signals had the buzzy sound typical for auroral reflection. This seemed to indicate to me that under exceptional conditions (the aurora was extremely intense) aurora can be beneficial to low-band DXing. This particular aurora generated an A-Index of 238. K-Index values were reported between 8 and 9. This was one of the largest geomagnetic storms since 1960. A similar situation existed in January 1987, when in Europe we could work KL7 stations during several days on 160 meters.

Will, DJ7AA, reported a similar happening (February 18, 1998): "Around 0130 I heard K1UO with a very big signal out from nothing working a SP3 station. I went on 1835 and one CQ brought me a huge pile with really big signals banging in here, even from call areas like W5 or WØ. I wonder what NAØY was running, I think he was the loudest WØ ever heard here at my place. Interesting, all signals having a little flutter sounded like aurora, and they were all coming in over my Beverage to South America, about 3-4 S-units stronger than on my big 500-meter Beverage to 320°.... At 0300 the band

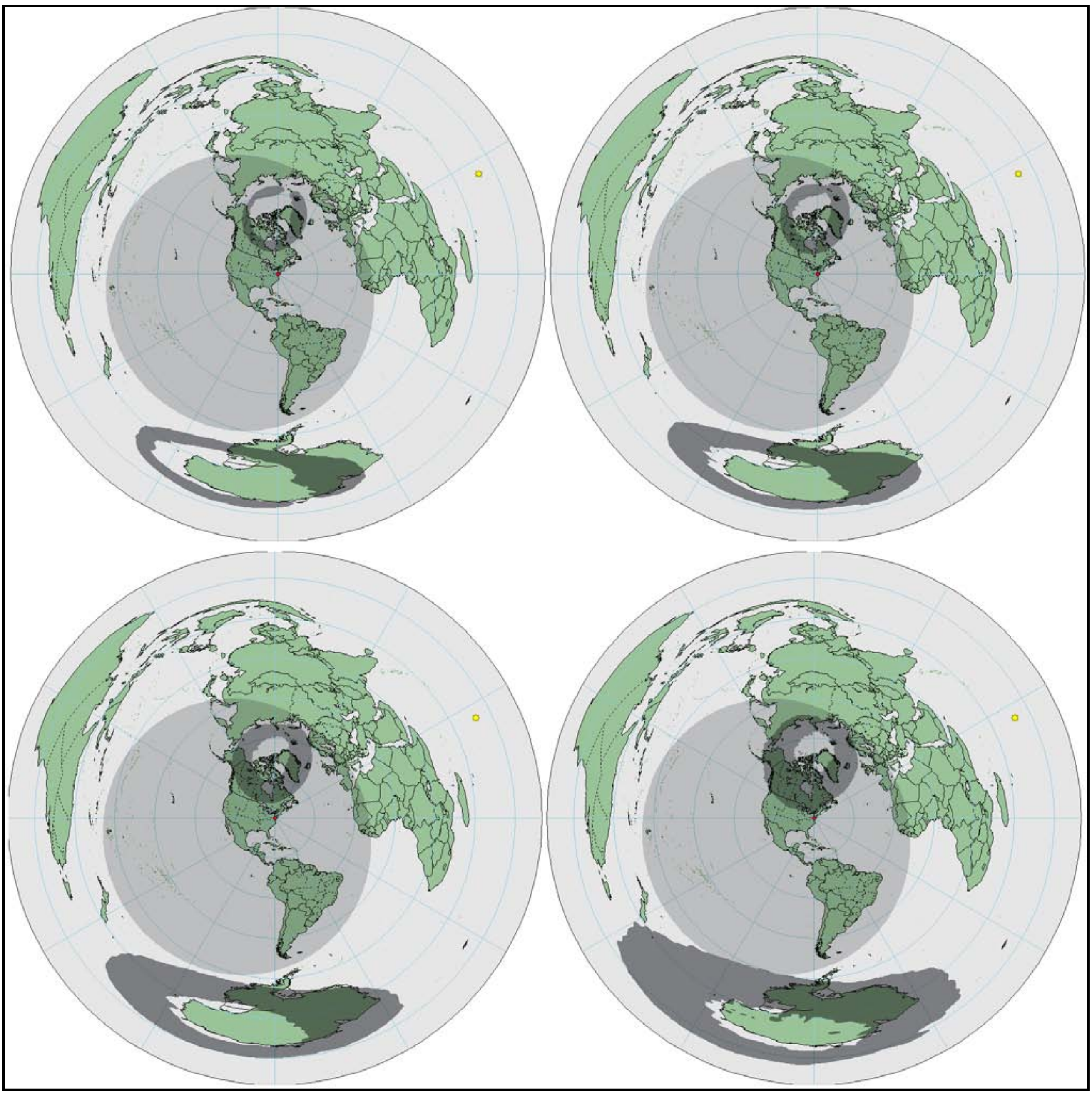


Fig 1-26 — Great circle map centered on Washington DC showing day and night zones for January 1 at 0700 UTC, and for varying values of magnetic K-Index (K=3, 4, 6 and 9).

died. When checking the NOAA home page I saw a very big auroral zone over the Northern Hemisphere at this time, while WWW said: Major Storm..."

There is almost always a temporary enhancement of conditions right after a sudden rise in the K-Index. Except for polar paths, however, it seems that a low K-Index doesn't help for most 160 meter propagation.

Enhanced propagation conditions shortly after a major aurora appear quite regularly. I witnessed a striking example on 80 meters on November 12, 1986, only nine hours after a major disturbance. N7AU produced S9 signals via the long path for more than 30 minutes, just before sunset in Belgium. Normally, long-path openings occur to the US West Coast from Belgium only between the middle of December and the middle of January, and even then the openings are extremely rare this far west in Europe. During the November opening, I heard N7UA calling CQ Europe with signals between S6 and S9 for almost an hour. The propagation was very selective, since only Belgian stations were returning his calls! A few days earlier DJ4AX was heard working the West Coast and giving 57 reports while the W6/W7 stations were completely inaudible in Belgium, only 200 miles to the northwest.

I presume that an ionospheric-ducting phenomenon was responsible for such propagation. This means that very specific launching conditions had to be present at both sides of the path. It appears that duct "exit" conditions are very critical and thus area selective — more so for longer path lengths. It also seems that auroral disturbances can occasionally create and enhance such critical conditions.

3.2.5 A-Index and K-Index

The most common way to quantify the level of geomagnetic activity is through the three-hour K-Index and the daily A-Index (Ref 158).

3.2.5.1. The Local K-Index

The K-Index indicates the magnitude of irregular variations in the magnetic field over a three-hour period. This index is calculated from the actual measured value at each observatory station. There are a number of these observatories worldwide. Since magnetic-field measurements vary greatly depending on location, the raw measurements are normalized to produce a K-Index specific to each observatory.

The K-Index scale is quasi-logarithmic, increasing as the geomagnetic field becomes more disturbed. K indices range in value from 0 to 9 (0 = dead quiet, to 9 = extremely disturbed). The K-Index that we often monitor on radio station WWV is an index derived from magnetometer measurements made at the Table Mountain Observatory located just north of Boulder, Colorado, and hence is referred to as the "Boulder K-Index." Every three hours, new K indices are determined and the broadcasts are updated. (See also Section 3.2.7.1.) For more details on the derivation of the K-Index, see "Where Do K and A Come From?" at **mysite.ncnetwork.net/k9la/**.

3.2.5.2. The Local A-Index

The underlying concept of the A-Index is to provide a longer-term picture of geomagnetic activity using measurements averaged over some timeframe. The A-Index is the mathematical average of the a-indices (the small "a" means it's a linear equivalent of the three-hour K-Index) over the last 24 hours.

The overall A-Index is an averaged quantitative measure of geomagnetic activity derived from the three-hour K-Index measurements. For each three-hour K-Index, a conversion is made to the a-index using a conversion table. (See **Table 1-2**.) The A-Index is the average of the last eight a-indices.

A indices are always linked to a specific day. Therefore, estimated A indices are issued during the day itself. For example, the Boulder A-Index (in the WWV announcement) is the 24-hour A-Index derived from eight of the three-hour K indices recorded at Boulder. The first estimate of the Boulder A-Index is at 1800 UTC. This estimate is made using the six observed Boulder K indices available at that time (0000 to 1800 UTC) and the best-available prediction for the remaining two K indices. At 2100 UTC, the next observed Boulder K-Index is measured and the estimated A-Index is reevaluated and updated if necessary. At 0000 UTC, the eighth and last Boulder K-Index is measured and the actual Boulder A-Index is produced. For the 0000 UTC announcement and all subsequent announcements the word "estimated" is dropped and the actual Boulder A-Index is stated.

A- and a-indices range in value from 0 to 400 and are derived from K indices based on the table of equivalents. The A-Index and K-Index for Boulder, Colorado, are broadcast by WWV on 2.5, 5, 10, 15 and 20 MHz every hour at 18 minutes past the hour. They are also available on the Web from **www.swpc.noaa.gov**.

3.2.5.3. Geomagnetic Activity Terms in English Instead of Numbers

As an overall assessment of natural variations in the geomagnetic field, six standard English terms are used in reporting geomagnetic activity. The terminology is based on the estimated A-Index for the 24-hour period directly preceding the time the broadcast was last updated. These are listed in **Table 1-3**.

3.2.5.4. Planetary A and K Indices

The Geophysical Institute in Goetteningen, Germany averages the data from 12 observatories (10 in the Northern

Table 1-2

a index	*Corresponding K*
0-2	0
3-5	1
6-10	2
11-20	3
21-35	4
36-61	6
62-102	6
103-166	7
167-268	8
>269	9

Table 1-3

A-Index Range	*Category*
0-7	Quiet
8-15	Unsettled
16-29	Active
30-49	Minor Storm
50-99	Major Storm
100-400	Severe storm

Hemisphere and two in the Southern Hemisphere) to give planetary values, A_p and K_p (the subscript p stands for *planetary*).

Table 1-4 shows an example of K, A, K_p and A_p indices from a Boulder report from June 24, 1997. It lists the Daily Geomagnetic Data from Fredricksburg, Virginia and College, Alaska, as well as the Estimated Planetary values from NOAA. You will see differences between the observations (at Fredricksburg and College) and the Estimated A_p and K_p values.

The K values are listed for three-hour intervals. Both the A-Index and K-Index are available from various sources on the Internet (see Section 3.2.7.1).

Both indices are equally important: K_p tells you if a geostorm is in progress, and A_p indicates whether the storm has just started or it has been developing for a while.

Of course the oval is a statistical description and does not describe how the ionization is distributed or how energetic it may be. In other words, the local intensity of the aurora is not the same in all points of the oval and at all times of the day. Also note that the K and A indices are for occurrences in the ionosphere at mainly E region altitudes — there's not much information about the F region in these indices.

3.2.6. Viewing the Aurora from the Satellites

The only source of really reliable information is to use real-time measurements superimposed on statistical maps, which now are available from various sources on the Internet. Today we have satellites that produce these data and can give us much more information than what we've had before. Views of both North and South Poles and the auroral oval are available at **www.swpc.noaa.gov/pmap/**. These are updated when the NOAA Polar-Orbiting Operational Environmental Satellite (POES) satellite passes by about every hour. The satellite maps out the auroral zone for that pass.

The POES images are based on particle-sensor readings the spacecraft makes as it passes over the polar regions. Instruments on board continually monitor the power flux of the protons and electrons that could produce aurora in the atmosphere. These readings are valid only for those longitudes where the spacecraft passes overhead. The readings may be considerably different at other positions along the auroral oval. This is why SEC must examine the results of 100,000 other polar passes in order to form a statistical picture of what is most likely happening elsewhere. This means that what the maps actually shows is based on data obtained from previous passes. It is not a real-time picture, but a combination of real-time data and best-fit extrapolations taken from a huge database.

Fig 1-27A shows a typical POES-generated polar view during a very quiet geomagnetic spell. The black line shows

Table 1-4

Date		*Fredricksburg*		*College, AK*		*Estimated*
		Local K indices		*Local K indices*		*Planetary K_p indices*
June	*A*		*A*		*A_p*	
16	8	1-2-1-2-2-3-3-1	3	1-0-0-3-0-1-1-0	5	1-1-0-2-2-2-3-1
17	6	1-2-2-1-1-2-2-2	1	0-1-1-0-0-0-1-0	5	0-2-2-1-1-2-2-2
18	3	1-1-1-1-1-1-1-1	0	0-0-0-0-0-0-0-0	4	0-1-1-1-1-2-1-2
19	11	2-2-4-2-2-2-2-3	6	1-2-3-3-2-0-0-1	10	3-2-4-3-2-2-2-2
20	6	2-1-2-2-2-2-1-2	2	0-1-2-0-2-0-0-0	5	2-1-1-1-2-2-1-2
21	2	0-0-0-0-0-1-1-2	2	0-0-0-3-0-0-0-0	3	0-0-0-1-1-2-1-1
22	15	2-4-3-3-3-3-3-2	4	1-2-3-1-1-0-1-0	9	1-3-3-2-3-2-2-2

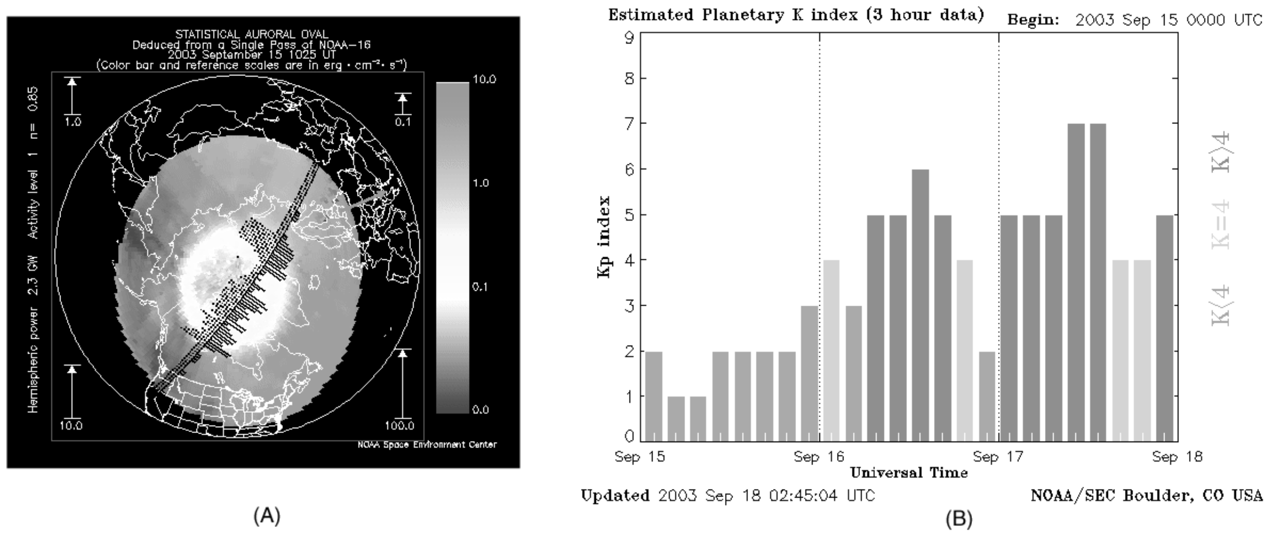


Fig 1-27 — At A, North Polar view generated by the POES satellite, during a magnetically quiet period (K_p=1-2). Inside the bright, light-shaded auroral zone "donut" dark areas are areas of high ionization. Darker areas outside the donut indicate lower levels. The black lines show the orbit of the satellite making the measurements, and the dots on either side represent the measurements done in the directions of the stacked white dots. At B, interplanetary K_p for time period in A. (Source: NOAA Web pages.)

the track of the satellite making the measurements, and the bars represent the flux of precipitating electrons (how many) while the dots represent the energy of the precipitating electrons (how low they go in the atmosphere). The arrow in the upper-right quadrant shows the local noon meridian, where the width of the oval is usually smallest. Note that the local noon meridian is not related to propagation on the low bands, since propagation at local noon is impossible anyhow due to D layer absorption. Fig 1-27A shows that the total power in the Northern Hemisphere during quiet geomagnetic conditions is 2.3 GW (that's Gigawatts = billions of watts), and what NOAA calls the auroral "Activity level" is 1, with a good confidence factor n = 0.85. (When the confidence factor approaches n = 2, NOAA is indicating that their statistical model is considered inadequate for this particular timeframe because the satellite is not covering an area sufficiently well to make accurate maps.)

Fig 1-27B shows a short history of the Planetary K (K_p) indices over the same period of days shown in Fig 1-27A and **Fig 1-28**. The K_p rose to 6 on September 16, 2003, and peaked at 7 on September 17, 2003, indicating a magnetic storm was in progress and indicating that aurora should be possible. Fig 1-27 shows the POES-generated polar view for September 17, 2003, during that magnetically upset period. Here, the total Northern Hemisphere power rose to 113.7 GW, with an Activity level of 10. The auroral oval did indeed intensify greatly and did spread to lower latitudes, especially across Northern Europe and Northern Asia.

There are many other pictures available taken from various spacecraft. **Table 1-5** lists the conversion from Total Hemisphere power, as reported by the NOAA, to the more familiar planetary K_p values and to the latest NOAA Aurora Activity Index. For a detailed discussion of these maps, read "A Look Inside the Auroral Zone" available at K9LA's Web site: **mysite.ncnetwork.net/k9la/**.

Table 1-5

Power (Gigawatts)	*K_p index*	*NOAA Aurora Activity Index*
0-2	0	1
2-4	1–	2
4 -6	1	3
6-10	2–	4
10-16	2	5
16-24	2+	6
24-39	3	7
39-61	3+	8
61-96	4	9
>96	5	10
>200	8	
>500	9	

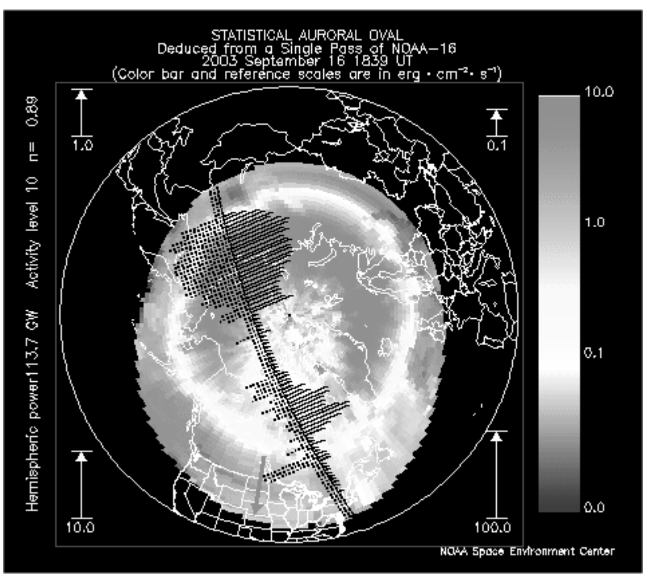


Fig 1-28 — North Polar view generated by the POES satellite, during magnetically upset period (K=5-6). Very dark areas inside the auroral donut are areas of high ionization, while the lighter tones outside the donut show much less ionization. The width of the oval is generally smaller at the local noon meridian. The particular view is for 1839 UTC on September 16, the day after the view in Fig 1-27A. Note that the auroral zone is at its widest around local midnight (across the USA) while the activity is minimal around local noon. The arrow near the bottom (in the mid USA) shows the local noon meridian.

3.2.7. Monitoring the Solar Wind

Early into the winter 2008-2009 season (still in the dip of the sunspot cycle, sunspot count nearly zero) the Top Band addicts witnessed unusually frequent and good quality propagation days over paths going through the auroral zones (Europe to Hawaii, Europe to Alaska, etc). It appears that these conditions coincide with the solar wind being at a very low speed (less than half of what we measure on average). The SOHO spacecraft (Solar and Heliospheric Observatory) continuously measures the solar wind data, and displays the speed (and other data) on its Web page (**sohowww.nascom.nasa.gov**). During the periods of extraordinarily good polar path propagation (on 160 meters), the auroral belt was barely visible, as could be witnessed on the NOAA Web site reporting the POES auroral activity (**www.swpc.noaa.gov/pmap/AnimateN.html**). It

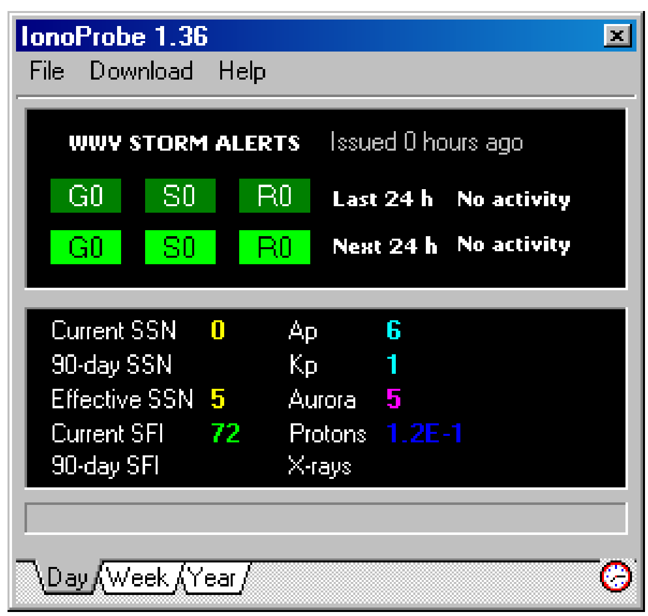


Fig 1-29 — *IonoProbe*, which you can have running permanently on your computer, looks for continuous updates on the Internet and shows you the very latest geomagnetic data.

appears that maybe, by monitoring the solar wind speed (in addition to all other relevant parameters), we can better assess the propagation conditions in the first instance on 160 meters. But, don't forget, it's just one piece of the puzzle.

3.2.7.1. Getting Geomagnetic Data and Using Them

The A-Index and K-Index for Boulder, Colorado, are broadcast by WWV on 2.5, 5, 10, 15 and 20 MHz every hour at 18 minutes past the hour. All DX spotting networks provide the latest WWV numbers. The *IonoProbe* program from VE3NEA also monitors the electromagnetic data relevant to HF radio. The list of parameters monitored includes A_p/K_p indices and the NOAA POES Aurora Activity parameter on a scale of 1 to 10 (**Fig 1-29** and Fig 1-2).

You can view a Solar Terrestrial Activity Report, which shows a chart of the solar flux, the sunspot number and the planetary A_p index, on **www.dxlc.com/solar.** See Fig 1-4.

How do we use this data? Low A- and K- numbers acquired at stations near the polar regions for a sustained period are prerequisites for good conditions on paths that go near or through these polar regions. It is the K-Index that is the most important one, since it gives you a more differentiated status than the A-Index. "Near the poles" means that the Boulder figures are not the most suitable ones! K indices obtained from the observatories in Inuvik, Baker Lake and Cambridge Bay in Canada are ideal because they are located within the auroral belt, when it is active.

We have to realize that K and A indices are measurements derived from what *has already* happened. If these indices have been zero for eight hours (or longer) and provided there is no abrupt change, the chances are real that the low bands will be in fair-to-possibly-good shape on polar paths.

Because of the sun's 27-day rotation cycle, low geomagnetic activity may be recurrent, especially during the declining and minimum phases of the solar cycle. During the ascending and maximum phases, the recurrent trend often becomes very unreliable (see also Section 2.2). It is a good idea for the serious low-band DXer to make a continuous log of broadcast A- and K-values. Such logged A indices are particularly interesting to predict the level of magnetic activity in another 27 days. W4ZV keeps a piece of paper marking the distance for 27.5 and 55 days and he uses this "ruler" to quickly calibrate the graph at **www.dxlc.com/solar** (Fig 1-4).

I guess most Top Banders have come to grips with the fact that K and A indices are there to confirm what they have already witnessed, good or bad conditions. N6TR, a well-known Top Band DXer from the Pacific Northwest (Oregon), complained: "I am very skeptical that any of the numbers mean much. I have had good openings with high K numbers, and no openings with longstanding low numbers. About the only thing I can count on is that interesting things seem to happen just as the K starts to rise."

Since auroral absorption is often initiated by CMEs or coronal holes on the sun, we should be able to predict auroras two to four days before they hit us (at least for the CME-induced auroras). Before satellite technology was available, we had no detailed information on CMEs and forecasting was based only on the use of recurrence tendencies, extrapolating conditions only from log data of A and K values from 27 and 54 days earlier.

You can also subscribe to *Sky & Telescope* magazine's AstroAlert service (actually written by Cary Oler of STD). This gives 24-48 hour notice by e-mail alerts of major CME events. See **www.skyandtelescope.com/resources/proamcollab/AstroAlert.html**.

3.2.8. Getting More Information

Today we have a number of satellites that keep a constant eye on the sun and send a continuous flow data to the Earth, data that is being converted into "readable" reports that are available in abundance on various Web sites on the Internet.

Solar Terrestrial Dispatch has a Web site (**www.spacew.com**) where you can find all sorts of information related to radio propagation. You can subscribe to a very useful daily summary of auroral activity. This is sent to you by e-mail from **www.spacew.com**. These reports forecast magnetic storms based on sun-surface and solar-wind observations, done from satellites. Such reports, together with viewing the NOAA-generated images themselves, are helping low-band DXers to better understand what makes it all tick and to better plan their activities.

SWIM (Space Weather Information Monitor) is a state-of-the-art, professional program created by Solar Terrestrial Dispatch. *SWIM* can monitor, display, animate or print to your printer over 200 space weather related Internet resources. See **Fig 1-31**. You can expand and manage thousands of additional Internet resources quickly and easily. You simply cut and paste Internet URLs for resources you find interesting and *SWIM* will immediately begin managing those resources for you. It tracks near-real-time geomagnetic A and K indices from as many as 26 global magnetic observatories world-wide. For further info, see **solar.spacew.com/swim**.

3.2.9. Correlating Geomagnetic Data with Conditions

Statistical analysis has been done on a representative group of long-haul DX QSOs from the US West Coast on 160 meters for a two month period. The occurrences were

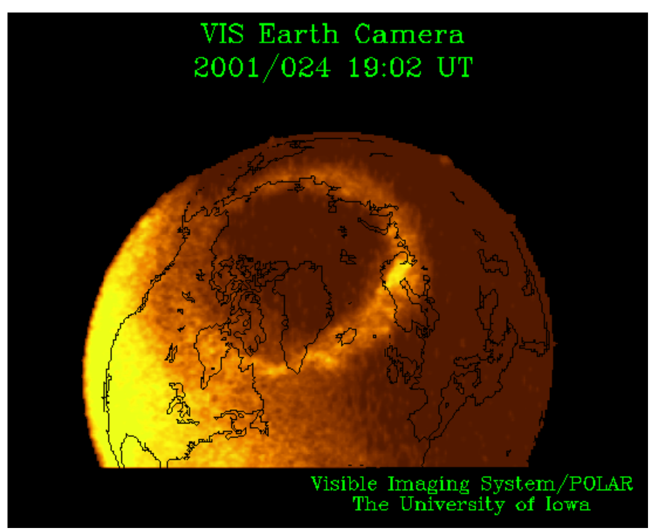


Fig 1-30 — Visible-light image of the Earth. The auroral belt is obviously only visible on the dark side of the Earth. Note the terminator moving across North America. (Photo courtesy University of Iowa.)

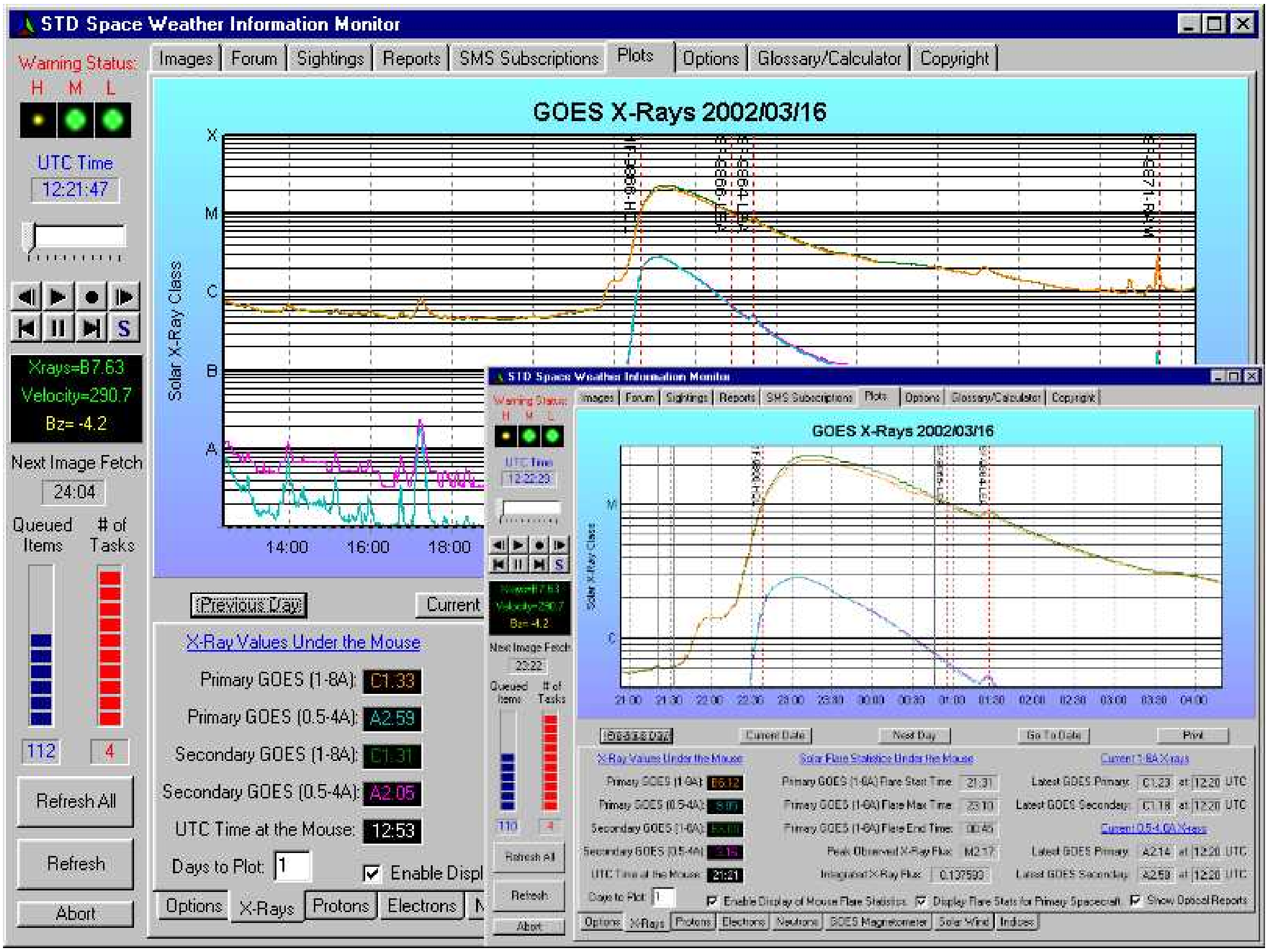


Fig 1-31 — One of the many SWIM screens, which show all imaginable data related to space weather. Graphs and pictures are automatically updated in "near real-time," so long as your computer is connected to the Internet. (Courtesy STD.)

checked against the K-Index:

- 62 percent of all QSOs were made on days with a K-Index of zero
- 30 percent with an index of 1
- Not one QSO with a K-Index above 3.

Studies like this always show periods of low K and A indices but with no propagation, indicating that other parameters are involved in low band propagation, especially on 160 meters. Thus it is obvious that there are other mechanisms that enter into the picture on 160 meters and that determine the overall attenuation on a given path. These mechanisms are still largely unknown or, to a large part open to speculation.

In a study, the *Top Band Monitor* did a survey and tried to correlate A-Index figures with days of good conditions on Top Band during the 1993/1994 winter. The author tried to link upward swings in A-Index with good Top Band conditions, and downward swings with bad conditions. My conclusion from studying the data was that only 10% of the good opening on 160 meters were correlated to a downward swing of the A-Index (Ref 173).

KBØMPL, who has a PhD in statistics, did a study on A indices and 160 meter propagation (Ref 174). She concluded: "Boulder A-Index changes, by themselves, apparently are not related to good or bad propagation days." She continued adding, "This does not mean that there is no relationship between good propagation and the A-Index."

Along the same lines Tom Rauch, W8JI, wrote on the Top Band Reflector: "I've given up totally on watching the A and K indices to estimate how the band is. What I find is generally when 10 through 20 meters is good, 160 is poor." That sounds like a simple and sensible guideline.

But we should not forget that magnetic activity is far from the only mechanism that rules conditions on the low bands, and more specifically on 160 meters. There are still many unknown mechanisms that make the residual attenuation on 160 meters vary significantly, even when the geomagnetic activity is low.

Sometimes we have weeks of really good conditions followed by weeks of fairly flat propagation. Strangely enough these good conditions on Top Band seem to happen frequently in October and November, what we would consider "early in the season." During both periods there are upswings and downswings of geomagnetic activity. The low bands and more particularly Top Band are still areas where many things still have to be "discovered." That's what makes these bands so interesting and appealing to many!

Trying to assess or predict propagation conditions on the low bands (especially 160 meters) going only by the A and/or the K-Index definitely doesn't work. While it is true that high magnetic indices will almost always result in poor propagation for paths near or through the auroral donut, paths that don't transit these zones may not suffer at all, and they may even be enhanced. Path attenuation by mechanisms other than aurora is consistently there on Top Band, and so far scientists have not been able to unambiguously correlate these mechanisms to any measurable phenomena.

3.3. Local Atmospheric Noise

Most local atmospheric noise (also called *static* or *QRN*) is generated by lightning discharges in or near thunderstorms. We know that during the summer QRN is the major limiting factor in copying weak signals on the low bands, at least for those regions where thunderstorm activities are serious. To give you an idea of the frightening power involved, a thunderstorm has up to 50 times more potential energy than an atomic bomb. There are an estimated 1800 thunderstorms in progress over the Earth's surface at any given time throughout the year. The map in **Fig 1-32** shows the high degree of variation in frequency of thunderstorms in the US. Fig 1-16 gives the picture worldwide. On average there is a lightning strike somewhere on the Earth every 10 ms, generating a tremendous amount of radio frequency energy.

In the Northern Hemisphere above 35° latitude, QRN is almost nonexistent from November until March. In the middle of the summer, when an electrical storm is near, static crashes can produce signals up to 40 dB over S9, and make even local QSOs impossible (and dangerous). In equatorial zones, where electrical storms are very common all year long, QRN is the limiting factor in low-band DXing. This is why we cannot generally speak of an ideal season for DXing into the equatorial zones, since QRN is a good possibility all year long. If you live in the USA, check **www.lightningstorm.com**.

The use of highly directive receiving antennas, such as Beverage antennas can be helpful to reduce QRN from electrical storms by producing a null in the direction of the storm. Unless directly overhead, electrical storms in general exhibit a fairly sharp directivity pattern.

Rain, hail or snow are often electrically charged and can cause a continuous QRN hash when they come into contact with antennas. Some antennas are more susceptible to this *precipitation noise* than others. Vertical antennas seem to be worst in this respect. Closed-loop antennas generally behave better than open-ended antennas (such as dipoles), while Beverage receiving antennas are almost totally insensitive to precipitation noise.

In very quiet places it is not uncommon for atmospheric noise generated on the other side of the world (often on the other side of the equator) and propagated just like regular radio signals to be heard many thousand miles away. This often shows up as "waves" of noise at the peak time for gray line propagation between the areas concerned.

3.4. Effects Caused by the Electron Gyrofrequency on Top Band

Modern DXers are aware of some special mechanisms that determine propagation on Top Band. The theory concern-

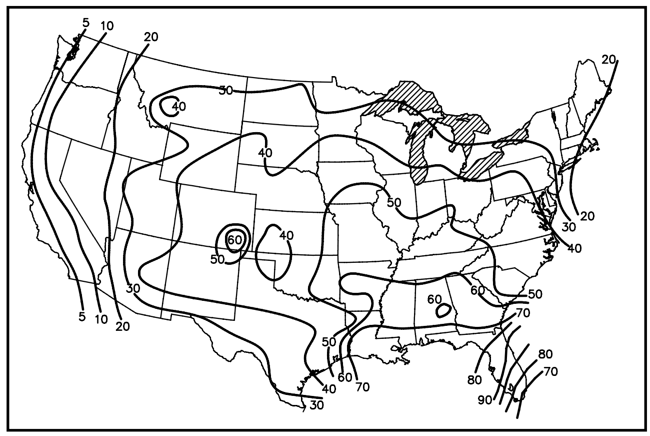


Fig 1-32 — This map shows the mean number of thunderstorm days in the US. The figure is related to both mountainous terrain and seasonal weather patterns.

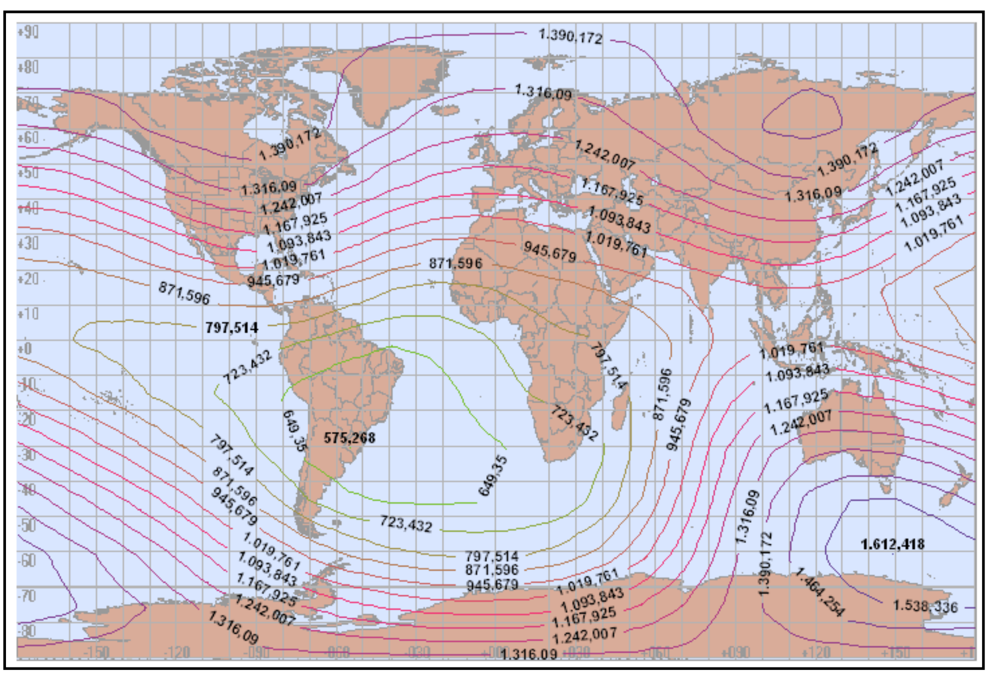


Fig 1-33 — This grid shows the worldwide distribution of the electron gyrofrequencies. These values are determined by the Earth's intrinsic magnetic field and are not influenced by the solar cycle. (Courtesy Cary Oler, STD.)

ing gyrofrequencies on 160 meters is covered in detail in the literature (Ref 142).

The electron gyrofrequency is a measure of the interaction between an electron in the Earth's atmosphere and the Earth's magnetic field. The closer a transmitted signal is to the gyrofrequency, the more energy is absorbed from the signal by the electron. This is particularly true for radio waves traveling *perpendicular* to the magnetic field. Gyrofrequencies are not influenced by the sun but change with location on the Earth (which is tied to intensity of the Earth's magnetic field). They vary between 700 and 1600 kHz around the world. A map of electron gyrofrequencies is shown in **Fig 1-33.**

You should remember that Top Band signals will be less strongly absorbed and behave more like a conventional signal is expected to behave the farther the frequency is removed from the electron gyrofrequency. Check the map in Fig 1-33 to determine the values of gyrofrequency your signals will encounter for a given path.

Absorption is higher along paths where the signal frequency is closer to the electron gyrofrequency, particularly on paths that are normal to the magnetic field. In other words, north-south paths are less affected than mainly east-west paths, such as from US East Coast to Europe or the US East Coast to Japan. Similar paths in other parts of the world may not be as sensitive because gyrofrequencies are lower.

If I had to quantify the impact of the gyrofrequency on Top Band propagation and compare it to the impact of the auroral oval, then I'd say that the auroral oval is the proverbial elephant, while the gyrofrequency is the mere mouse.

3.5. Polarization and Power Coupling on 160 Meters

Power coupling has to do with the way electromagnetic waves generated by the transmit antenna "couple" into the two characteristic waves that propagate through the ionosphere: the ordinary wave and the extraordinary wave. It appears that the polarization of the antenna plays an important role in achieving optimal coupling (minimum losses). Power coupling is generally greatest with the ordinary wave when the E field from an antenna is parallel to the geomagnetic field and the least when the two are perpendicular to each other.

In certain areas of the world vertical polarization will produce strongest signals, while in other areas horizontal polarization will. Fortunately, in the US as well as in Europe, vertical is the way to go. This may explain why even 0.5-λ high dipoles do not seem to work well from these regions on 160 meters, while they do fine on 80 meters. There are areas of the world, however, where horizontal polarization on Top Band is the more suitable polarization. This is true for large parts of Asia, Africa and parts of Australia. The geomagnetic latitude of the location is an important factor in this mechanism.

Fig 1-34 shows a Mercator map showing the geomagnetic latitude compared to the geographic latitude and longitude. W8JI, in a message on the Top Band reflector, put things in perspective: "Losses incurred by TOA (take-off angle) effects (a high-angle antenna vs a low-angle antenna) can be more than 10 dB. Losses incurred if you have very poor ground (in the far field) vs very good ground can be 4 dB. Losses due to improper magnetoionic power coupling can amount to approximately 1 dB. Most Top Banders use transmitting antennas with vertical polarization, which fortunately seems

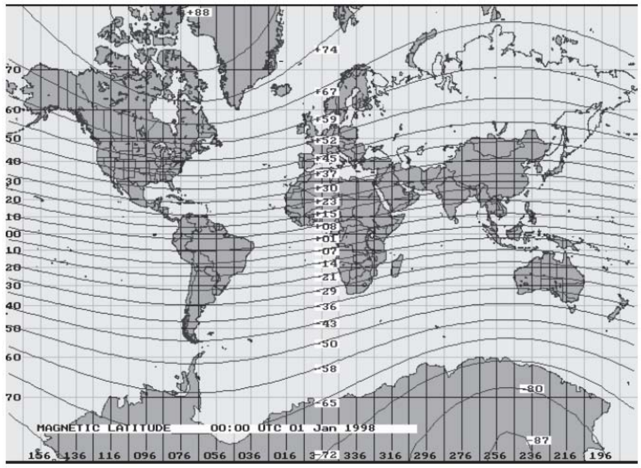


Fig 1-34 — Mercator-projection world map showing the geomagnetic latitudes. These are not the same as the geographical latitudes, since the magnetic North and South Poles do not coincide with the geographic ones. (Map generated by *Proplab-Pro*.)

to be the right choice from a power coupling point of view, at least if your QTH is at an average or higher-than-average latitude. It is only stations within 20° of the magnetic equator that may be concerned about power coupling. Even at these low latitudes it is better to have a vertically polarized antenna with a TOA of 25° than a horizontally polarized antenna with a TOA of 90°. This antenna would radiate 10 dB less signal at 25° (typical for a dipole less than ½-wave high). With this antenna you may win 1 dB in power coupling but lose 10 dB due to an inappropriate radiation angle!"

4. PROPAGATION PATHS

This section discusses the following items to help increase our understanding of low-band propagation paths:

- Great-circle short path
- Great-circle long path
- Particular non-great-circle paths

4.1. Great-Circle Short Path

Great circles are all circles obtained by cutting the globe with any plane going through the center of the Earth. All great circles are 40,000 km long. The equator is a particular great circle, the cutting plane being perpendicular to the Earth's axis. Meridians of longitude are other great circles, passing through both poles.

When we speak about a great-circle map we usually mean an *azimuthal-equidistant projection* map. This map, when covering the entire world, has the unique property of showing the great circles as straight lines, as well as showing distances to any point on the map from the center point. On such a projection, the antipode of the center location will be represented by the outer circle of the map. Great-circle maps are specific to a particular location. They are most commonly used for determining rotary beam headings for DX work. The advantage of a great-circle map is that headings from the center are straight lines, while the disadvantage is the extreme distortion near the antipode (the outer ring at 20,000 km).

Fig 1-35 shows great-circle maps using *DX Atlas* cen-

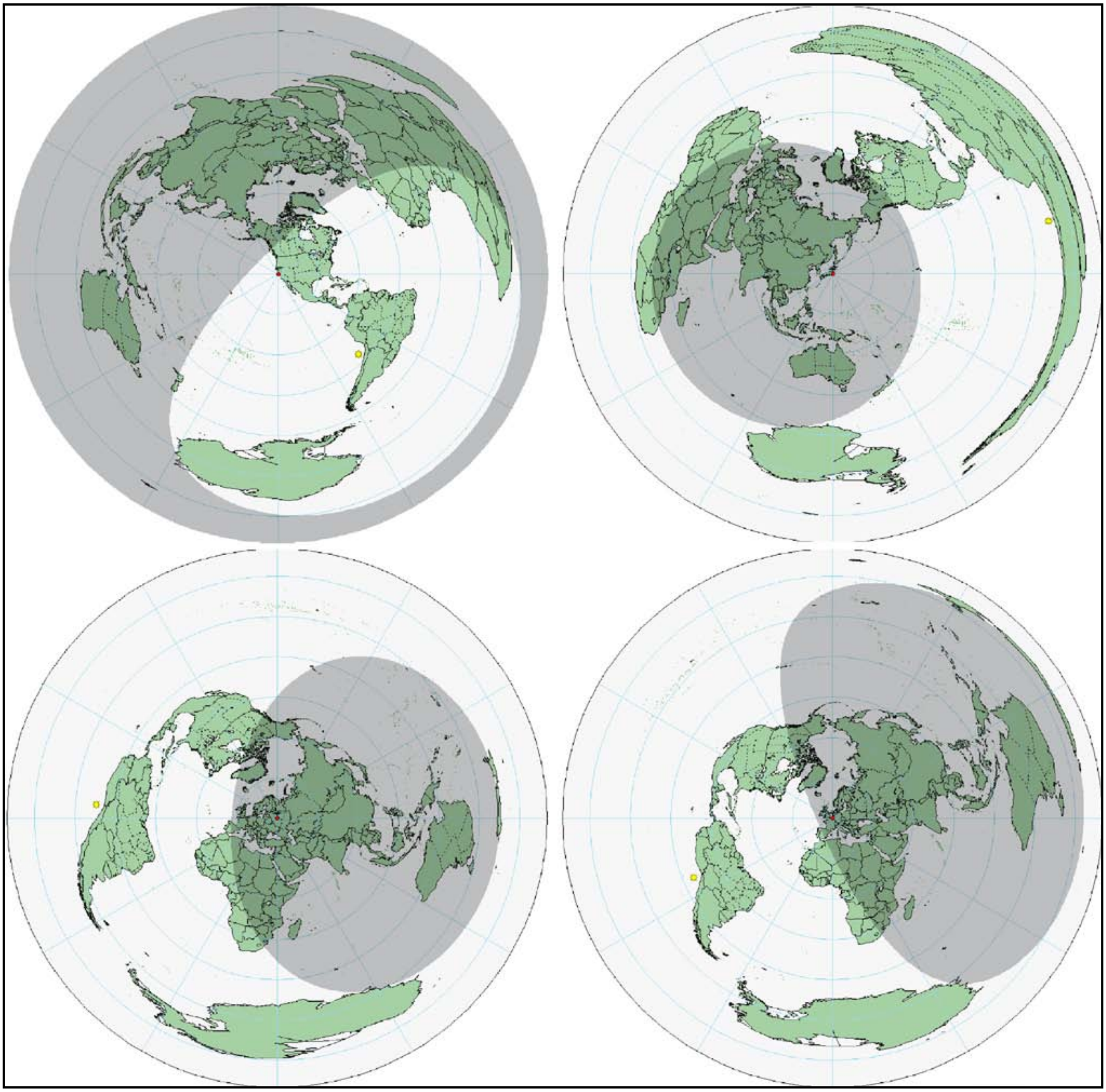


Fig 1-35 — Azimuthal (great circle) projections centered on San Francisco, Tokyo, Moscow and Belgium. These great circle maps generated by *DX Atlas* also show night and day. All maps made for January 1 at 0700 UTC. (Created with *DX Atlas* software.)

tered on San Francisco, Brussels, Moscow and Tokyo. There are various sources on the Internet where you can download great-circle maps or programs to make such maps.

4.2. Great-Circle Long Path

A *long-path* condition exists when the station at the eastern side of the path is having *sunset* at the same time as the station at the western end of the path is experiencing *sunrise*. A second, necessary condition is that the propagation occurs on a path that is 180° opposite to the short-path great-circle direction.

We will see further how "crooked-path" propagation can satisfy the first condition, but is not genuine long-path propagation. One example is the 80 meter propagation path from Western Europe to Japan at 0745 UTC in midwinter. This involves a crooked short path over northern Siberia (see also Fig 1-48 later in this chapter) and not across South America, as it would be if it were a true long path.

Bill, W4ZV, uses another definition of long and short path: signals are short path if they travel along the shorter great circle direction and any direction that is less than 90° off that direction. Similarly he defines long path as the path where signals travel along the longer great circle direction (the direction opposite to the geometrical short path direction) and any direction which is less than 90° off the longer great circle direction.

But, what's in a name? Let me stick to the more usual definition, as mentioned above.

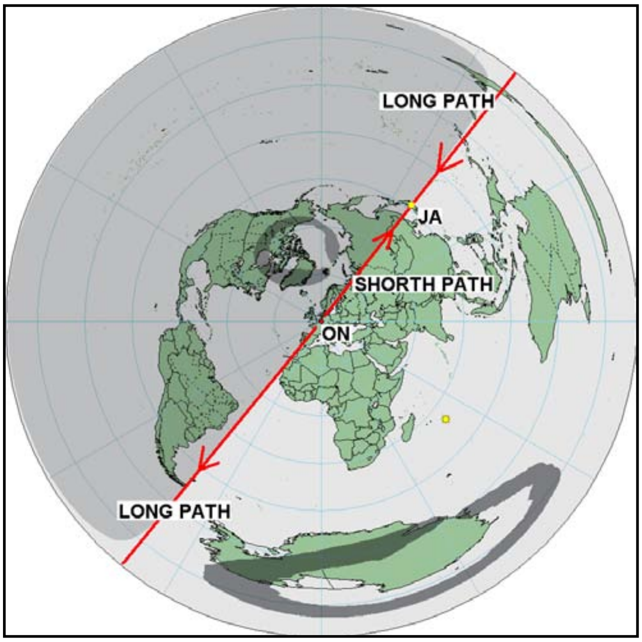


Fig 1-36 — The great-circle path for midwinter (0745 UTC) shows the short and the long path that exist simultaneously on 40 meters from Belgium to Japan. Both run along the terminator, a typical situation for the higher bands, but very uncommon for 80 meters and impossible on 160 meters. (Maps generated with *DX Atlas*, with additions by ON4UN.)

4.2.1. Long Path on 40 Meters

Long-path QSOs are quite common on 40 meters. From Europe we have a genuine long path to the US West Coast around 1500 to 1600 in midwinter. A very similar long path exists between Japan and Europe around sunrise time in Europe, especially around the equinox. In midwinter, when all the darkness is in the Northern Hemisphere, there still is some long path between Europe and Japan, but there is a generally much-stronger path that is a somewhat-crooked short path across northern Siberia. In general, the signal direction is about the same as the usual short path direction.

During midwinter, both long and crooked paths exist simultaneously for about 10 or 15 minutes around 0745 UTC (see **Fig 1-36**). This often makes copy very difficult because of multipath propagation due to the different time delays on each path. During the JA low-band contest in January 1998, I had to ask several JA stations to slow down their CW to allow me to copy through the multipath echoes.

4.2.2. Long Path on 80 Meters

Genuine long paths to areas near the antipodes are very common on 80 meters all through the sunspot cycle, provided there is a full-darkness path and that the long path coincides with areas of lowest attenuation (see also Section 4.3.1.2). Long paths on 80 meters are less common than on 40 meters, except to areas very close to the antipode. Very often paths which we call "long paths" are crooked or bent paths, somewhere between long and short paths (see Section 4.3.).

4.2.3. Long Path on 160 Meters

Genuine long-path QSOs on 160 meters are almost always to places near the antipodes. Such long-path QSOs are very rare during the sunspot-maxima years. During the low sunspot years I can hear G stations working ZLs long path on 160 meters approximately 30 minutes after my sunrise, but only on very rare occasions have I been able to work ZL on long path myself. Other near-antipode long-path QSOs have been made between VK6HD (Perth) and the US East Coast in midwinter (eg, the QSO between K1ZM and VK6HD at 2115Z on January 27, 1985).

Real long-path QSOs (long path that shows no path skewing) on 160 meters only seem to occur during a period centered on the one or two years at the minimum of the sunspot cycle. W4ZV, when he was WØZV in Colorado, remembers a few genuine long-path QSOs made from Colorado; for example, with UA9UCO and JJ1VKL/4S7. Another one that made history was between PY1RO and several JA stations at JA sunrise. Other long-path contacts were made between US East Coast stations and well-known calls, such as 9M2AX, VK6HD and VS6DO.

Many of the often called long-path 160 meter QSOs are really skewed long paths, and they happen at all stages of the sunspot cycle. Examples are the early 2003 QSOs between JT1CO and US East Coast stations. More details can be found on W4ZV's Web site **users.vnet.net/btippett/dx_aid_plots.htm.**

During the 1987-1988 winter, my first winter on 160 meters, I tried for weeks to make a long-path QSO with N7UA, but we never heard signals at either end. During December 1992, I ran a daily test with FK8CP on the long path (his sunset is within minutes of my sunrise), but we never made a QSO either (see also Section 4.3.1.4).

During December 1997, N7UA, with whom I had many long-path tests back in 1987/1988, made numerous so-called long path QSOs into eastern and northern Europe as well as into the UK around 1510 UTC. But were these genuine long-path QSOs? Let's have a look at non-great-circle paths, also called crooked or bent paths.

The only real long path QSOs I ever made on 160 meters are with a small number of ZL stations, including ZL7. In over 20 years of activity on Top Band I made exactly 14 such QSOs, all between 0735 and 0750 UTC in a period between the end of December and the end of January.

4.3. Crooked (Skewed) Paths

Most propagation paths over relatively short distances on 40, 80 and 160 meters are great-circle paths. We do know, however, that signals quite often come from anything but great-circle directions. So let's distinguish two categories for the path bending we often observe:

Bending caused by aurora: This case is the classic one. During periods of high geomagnetic activity (aurora), we identify signals coming from headings off great-circle directions, when the great-circle path would otherwise have to go through the auroral oval. We have all witnessed repeatedly how signals seem to be bending around the auroral belt. In Europe we work US West Coast stations beaming to central or South America under such circumstances.

Bending *not* caused by aurora: A second type of bent path on 80 meters (and especially on 160 meters) was witnessed by many operators on the US East Coast in January 1997. During the first few days of the VKØIR operation they remember well how the signals peaked right across Europe (60°) instead of on the direct path, which is about 110°. This could not have been a case of seemingly bending *away* from the auroral oval. Rather, it almost was like the signals were being *attracted* to it! During the VKØIR operation geomagnetic conditions were generally very quiet. The reason for this kind of bent path must be different, since there is no aurora involved.

Tom, W8JI, suggested on the Top Band reflector: "Skew paths are actually fairly common, and don't seem to be tied to anything unusual going on if the path is long. So it seems to me signals simply come from the direction of least absorption. And by the way, there always is the same skew on transmit as there is on receive." This lines up perfectly with what Thomas, KN4LF, wrote: "160 meter propagated signals are always going to travel along the path of least absorptive resistance." Sounds very logical, doesn't it?

We should not forget that we do not transmit with antennas exhibiting infinitely sharp directivity, and that hence our signals are not transmitted along a single narrow path that looks like a single thin line on a map. In reality we normally transmit in a broad direction, into a wide area where the signal will travel best in zones with least attenuation. And these zones may not be located along the straight line between transmitter and receiver! In other words, the path does not necessarily look like a thin straight line, but may be a bent or crooked one. If we accept this idea, then the question here is: "What causes such path bending?"

4.3.1. The Non-Heterogeneous Ionosphere

4.3.1.1. The Mechanism for Deviation from Great-Circle Paths

It is generally accepted that there are only three ways that signals propagate through the ionosphere:

- By *refraction* caused by ionization gradients
- By *reflection* caused by auroral ionization
- By *scattering* of signals by ionospheric or atmospheric irregularities, as well as irregular surfaces (ground or water).

The general mechanism that causes signals to deviate from the great-circle path is the presence of *steep horizontal ionization gradients* in the ionosphere. Signals traveling into a layer with a higher degree of ionization will be refracted or even reflected away from the gradient. Very steep gradients can be caused by aurora, for example. When there is low geomagnetic activity, however, scattering in the ionosphere itself (or at ground-reflection points) could be another mechanism that can cause path skewing.

The ionosphere is not a perfect mirror, but should rather be thought of as a cloudy and patchy region, with different areas of ionization. Tom, W8JI, stated: "There is more scattering and skewing going on than most of us ever know about, probably because it isn't a shiny smooth mirror up above."

We often visualize a radio wave as a single ray sent in a specific direction, refracted in the ionosphere (which we think of as a perfectly shiny mirror) and reflected from a perfectly flat reflecting surface on the Earth. HF energy, however, in most practical cases is being radiated in a range of azimuths (even if a Four Square antenna is used!) and over a range of elevation angles. Some signal is thus taking off in the "wrong" direction (that is, not in the great-circle direction toward our target) and may change course enroute by any of the mechanisms described above and yet still arrive at the target!

Part of our transmitted signal, of course, actually does take off in the "correct" great-circle direction, but it might encounter the auroral belt and be totally absorbed there. Even if it isn't completely absorbed, it may be reflected or scattered there and who knows what direction it may end up taking? When and if some signal does reach the destination, there is usually one path (straight, bent or whatever) where the received signal is substantially stronger than those received via other paths. Thus we are mainly aware of the most successful path.

Sometimes we hear signals coming from various directions at the same time. Tom, W8JI, wrote: "Many times the JAs are SW, and many days the JA signals arrive from multiple directions. When K1ZM and AA1K hear JAs from the NW, I'm hearing them better from the SW." This clearly demonstrates that signals from JAs don't travel on just one path. They are propagated in many directions and are received in different places from different directions. The mechanisms behind all of this are very complex ones, and aurora is but just one, but important, cause of path bending.

Reception from multiple propagation paths normally does not cause any problems, since the difference in propagation delays usually is quite small (1 to 10 ms). Sometimes, however, the delays are of an order of magnitude that cannot be explained by a *slightly* bent path. Tom, W8JI, wrote: "I can hear K9DX, when he is beaming NW, scattering in from the SW with ¼-second to ½-second delays on the echo. (Between John's TX antenna and my RX antenna there is probably a 60 dB null on the direct path.)" Such a long-delayed echo must obviously involve some other mechanism than minor path skewing.

The path direction may vary from day to day, and even on a given day may switch continuously and at a very rapid rate. Mike, VK6HD, observed in the Internet reflector: "Last night I had six QSOs with NA between 1134 and 1155Z. With the first one I thought there was very strong QSB, but then I checked my Beverages and I found that when the signal went down on the NE Beverage it came up on the SE one, and vice-versa. This switching was happening about every 20 seconds."

Clearly, many paths on the low bands are not simple great-circle paths. Testimonies in this respect are overwhelming. W8JI worded it as follows: "The only nearly 100% agreement you will see is people with directive antennas who do a lot of listening over long periods of time all agree that not much ever comes in through the magnetic pole areas, and that paths (for really long distance signals) on lower bands are not predictable."

As to the exact why and how, those questions remain largely unanswered. Cary Oler and Ted Cohen (Ref 142) point out that "Weak sporadic-E clouds, that might not affect the higher frequencies, can achieve a substantial impact on 160 meter signals by increasing absorption or refracting signals." Such sporadic-E clouds can induce a waveguide-like hop between the F layer and the sporadic-E cloud. They are also considered as a possible cause for skewed paths. There still is a lot of discussion ongoing in scientific circles about the mechanisms that trigger path skewing. Here are some regular, well-documented skew paths.

Carl, K9LA, did an in depth analysis of a typical bent path QSO between W4ZV and SM4CAN (March 10, 1999, 02:30 UTC, K=5-6) (Ref 159). According to the analysis by K9LA, from both sides the signals traveled along great circle directions which cross one another in a region west of the Canary Islands. Analyzing the ionospheric maps for that date and time, Carl found in that area sufficient steep ionization gradients that can explain the path bending. It appears that

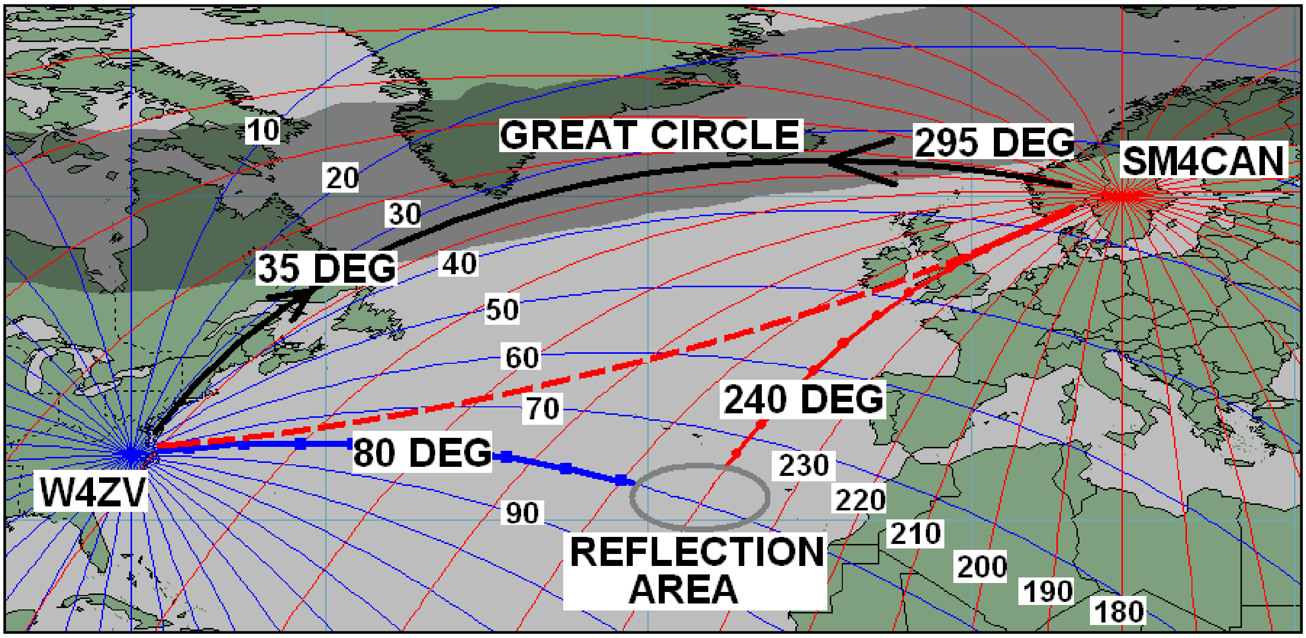


Fig 1-37 — Crooked path between W4ZV and SM4CNN: the more northerly smooth path is the great circle path crossing the auroral zone (the darker band). The broken line indicates the correct directions from which the signals are received (corresponding with the direction of the two bottom lines which shows the actual path followed by the signals). To be able to draw the actual path one must be able to localize the actual steep ionization gradient that makes the path "switch" from one great circle direction to another. As such analysis cannot be done in all cases, we represent the "crooked" paths with a smooth broken line, where only the azimuth angles are correct. (Map generated with *DX Atlas*, which indicates the great circle lines for the two locations.)

such steep ionization gradients are not only found near the auroral donut, but also at much lower latitudes. Even more interesting in Carl's study is that the mechanism involved was not one of refraction but rather reflection! Carl writes "When you think about it, this is kind of a nice thing for the auroral oval to do: provide us with another path when it shuts down the normal path due to increased absorption." Carl's analysis work is based on observations and scientific data, and all the pieces of the puzzle seem to fall into place, but he says himself "this is just a hypothesis."

In all the great circle maps that illustrate the crooked paths, I have symbolically represented the crooked path as a nicely curved dotted line (**Fig 1-37**). The only thing that is real is the path directions at both ends of the path; these are realities that have been observed by thousands of us on the low bands (especially 160 meters).

During the VP6DX (Ducie Island) expedition in February 2008, a number of 160 meter "long path" QSOs were reported on the VP6DX Web site. It is interesting to analyze these QSOs, to try to understand what made them possible.

On February 18, VP6DX worked A45XR at 1252 UTC. A45XR is within 300 km of the exact geographic antipode of VP6DX. For practical purposes, let's call it "the" antipode (see **Fig 1-38**). This means that all paths have the same length. On the great circle map centered on VP6, A45 is the outer circle. Fig 1-38 shows the daylight/darkness situation at that exact time. Sunrise on VP6 is around 1408, 67 minutes later. At that exact time it is also sunset in Oman. At 1252 UTC we are still 67 minutes before sunrise in A45 which means that the distance from A45 to the terminator is still approximately 2100 km. The report from VP6DX says that the signals were received coming from 305° which is a path right over Japan to A45. Notice that the propagation path is perfectly at right angle with the terminator near A45, which means that the travel distance of the signals through the gray zone is minimal. The path is far away from the auroral zones, which means there is no reason to assume a crooked path would be involved. It is amazing that 2100 km of the path was in daylight, all of it at the A45 end of the path. Being a perfect antipode path, we

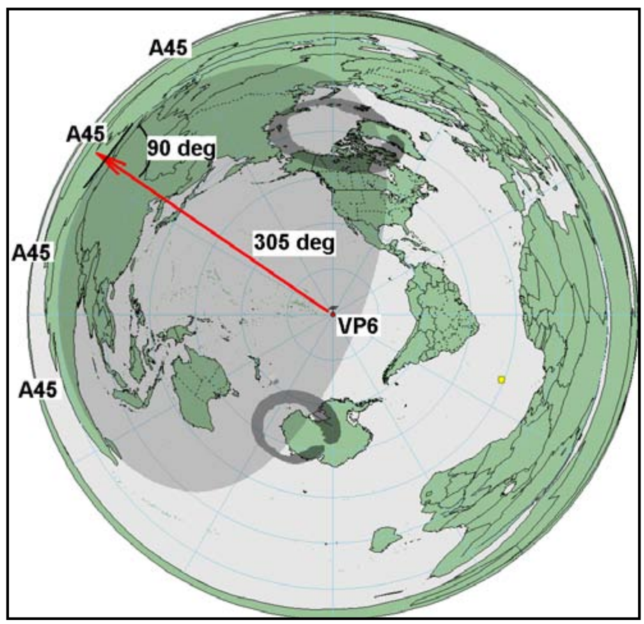


Fig 1-38 — The VP6DX path to A45XR for a genuine antipode QSO. See text for details. (Map generated with *DX Atlas*.)

can envisage the so-called effect of antipodal focusing (see Section 2.4.4.8). The distance being well over 11,000 km, it cannot have been straightforward multiple hop propagation, so ducting must have been involved. The only thing that is extremely remarkable is the exact time of the QSO. If the QSO had taken place at 1408 UTC, it would have been a perfect textbook example for a QSO right into the antipode. But the signals traveled 2100 km in daylight, which is quite a distance. In Belgium I had been copying VP6DX as late as 40 minutes after my sunrise, which I thought was quite exceptional, but this is even much more so.

Two days later the same DXpedition reported a number of long path 160 meter QSOs with southern Russia and Ukraine, all in the timeframe of 1345 to 1430 UTC (see **Fig 1-39**). All contacts were made with signals coming from the southwest (195° to 225°) as indicated by the arrow in Fig 1-39. If the signals would really have traveled all the distance at a heading of approximately 195° from Ducie, they would have traveled all the distance inside the gray zone, which seems very unlikely, because of the extra attenuation. Also the signals would have had to break through the southern auroral donut twice, which makes it even more unlikely.

The southwestern path from VP6 hits the southern auroral donut south of New Zealand (see **Fig 1-40**). Let us assume

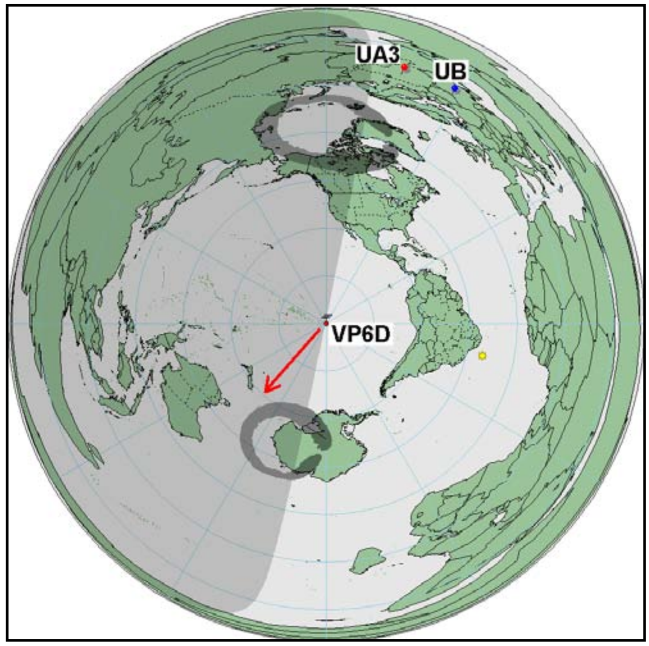


Fig 1-39 — Darkness/daylight great circle map centered on Ducie Island for February 21 at 1415 UTC. See text for details. (Map generated with *DX Atlas.*)

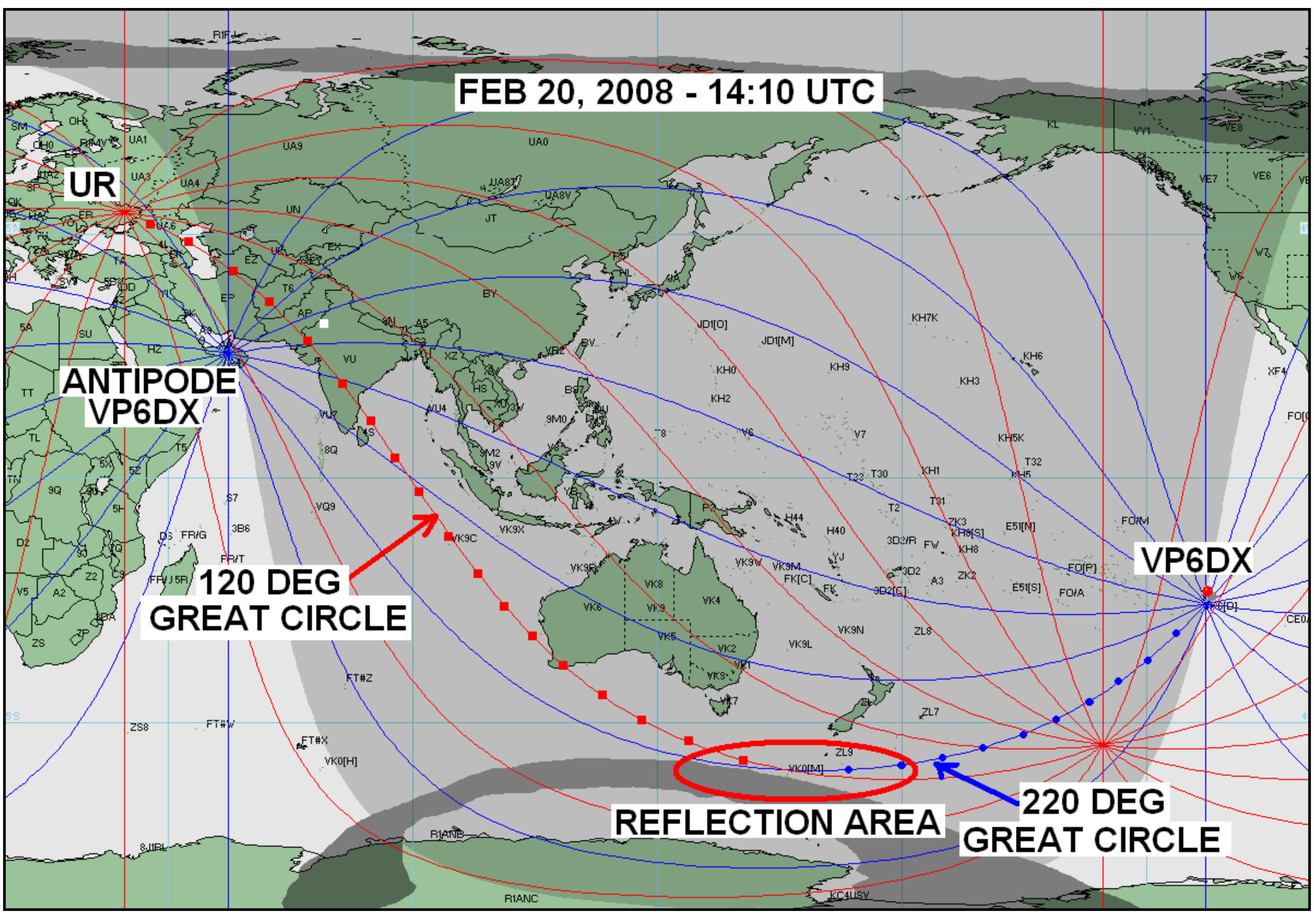


Fig 1-40 — Starting with the take-off angles as noticed by VP6DX, we have reconstructed a probable path for the QSOs between VP6DX and UR/UA at sunrise time on Ducie. This shows two sets of great circles lines (every 20 degrees), one centered on VP6, the other one on UR (Ukraine). See text for details. (Map generated with *DX Atlas.*)

there is sufficient ionization gradient in that area to refract the signal and send it back on a path south of Australia, across the eastern Indian Ocean and India toward UA/UB (this is the 120° path out of Ukraine identified by square markers on the map). The end of the path is still in daylight but relatively close to the terminator (800 km or 30 minutes). This crooked path is exactly 23,000 km long. I believe this is much more likely the path the signals followed rather than through the aurora and along the terminator.

During the same time period, a station in the Moscow City oblast was also contacted. The VP6DX Web site explained that for this QSO "the signal clearly arrived on the short path, crossing over Scandinavia." It seems extremely unlikely that a direct path was involved; the signals would have traveled all the way inside the gray zone and right through the auroral belt. **Fig 1-41** shows the situation at the time of that QSO. A path heading 330° to maximum 340° (or less) from Ducie Island is required to avoid the auroral belt. If sufficient ionization gradient was available in an area north of Mongolia in Central Asia, we can imagine that the signals would change heading and propagate on a path we can identify as a 60° path out of Moscow by the mechanism explained in Fig 1-37. **Fig 1-41** shows how both the path out of Ducie and the path out of Moscow avoid the auroral donut and meet somewhere across central Asia. Of course we are not sure that this is what really happened, but this is more likely than a direct path at a heading

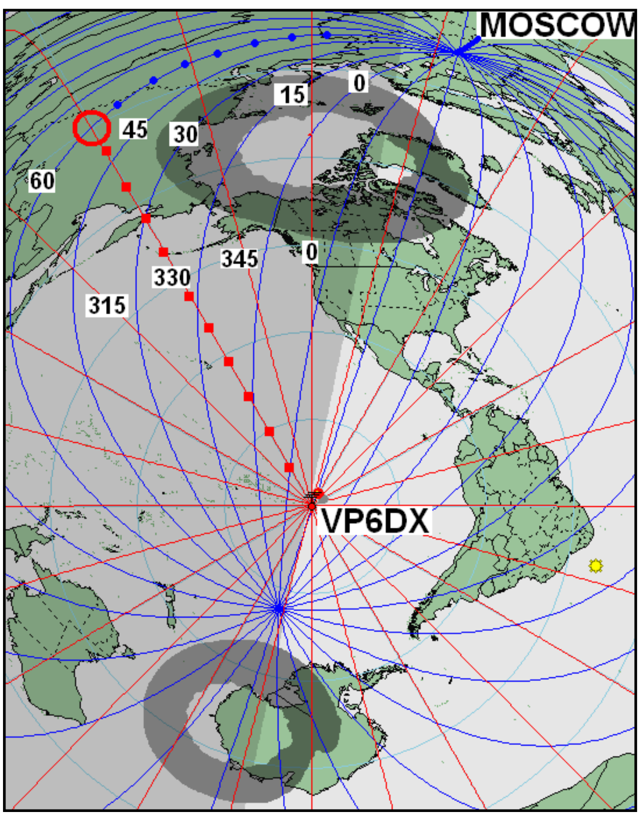


Fig 1-41 — This part of a great circle map centered on Ducie shows the great circle lines (every 15 degrees) coming out of VP6DX and out of the UA3 area, as well as the daylight/darkness situation around 1415 UTC on February 20. The circled area across northern Siberia is the area where path switching "could" have occurred. See text for details. (Map generated with *DX Atlas*.)

of 195° out of Ducie which is right along the terminator (or in daylight) and twice across the auroral belt.

I also think that even with Beverage receiving antennas it is not always possible to tell the signal heading with an accuracy of better than a few tens of degrees. All the time I copied VP6DX at my sunrise, the signals were equally strong on three Beverages: 270°, 295° and 315°.

What can we learn from these unique 160 meter experiences with VP6DX on Top Band?

That exceptionally "good" conditions happen *every now and then*. As we can be "certain" that these very long distance QSOs (16,500 km) could not have been made via a path going along the terminator nor via a path twice crossing the auroral donut, we can only conclude that path bending was involved. But for path bending to occur everything must be just right at the right place at the right time, and such a coincidence certainly does not happen every day!

During a two and a half week stay on Ducie this apparently happened only on one day. And the QSOs were made because during those two and a half weeks the operators were continuously active (transmitting, because listening is not enough on 160 — if everybody listens, nobody will hear anything even if the band is in the best of shape), from way before their sunset until way after their sunrise. Persistence and patience are unique virtues when it comes to working the really difficult stuff on Top Band.

But that by itself is not enough to make such exceptional QSOs. There are other variables involved. You need good antennas (both transmit and receive), good power and good ears. They had all of that on Ducie island. If we would have operations like that going to all the really rare countries (see Chapter 2, Section 16) we could "easily" work all countries on 160!

4.3.1.2. The Classic Skewed-Path Example: ZL Propagation from Europe

New Zealand is about 19,000 km on the short path from my QTH in Belgium, or about 21,000 km on the long path, very close to being the antipode (see Fig 1-33). From Belgium the short-path heading to New Zealand is 25° to 75° and the long-path heading is between 205° and 255°. When I work ZLs on 80 meters on long path during the Northern Hemisphere winter, signals almost always arrive via North America, at a heading of approximately 300°. This is 90° off the great-circle long-path direction. The path is not a great-circle path, but is inclined as if the signal were trying to leave the Southern Hemisphere as fast as possible (both the ZLs and the Europeans beam across North America in the winter).

As we move into spring, the optimum path between Western Europe and New Zealand moves from across North America to across Central America. Eventually, beaming across South America will yield the best signals later in the year (summer). Somewhere around the Spring Equinox all three paths produce equally good signals, when the signals are at their strongest.

Theory says that there is an indefinite number of great-circle paths to the antipode. Since low-band DX signals travel only over the dark side of the globe, however, the usable number of great-circle headings is limited to 180° (assuming there is no auroral activity screening off part of the aperture). This very seldom means that signals will arrive with equal strength over 180°, not to mention with the proper phase. The relatively short differences in path lengths cause time-delay differences too short to be able to noticed by ear. The strongest signals are received from the direction where the attenuation is least. This confirms W8JI's comment that "Signals simply come from the direction of least absorption."

These New Zealand-to-Western Europe QSOs are well-documented examples of gray line propagation, but none of these propagation paths ever coincide with the terminator itself. The actual path *happens* to be more-or-less *perpendicular* to the terminator at all times of the year (see also Section 2.4.4)!

To summarize, on 80 meters I have observed for nearly 50 years that long paths and paths to areas near the antipodes are skewed in such a way that the signals will apparently travel the longest possible distance in the hemisphere where it is winter. This is judging from the direction of arrival of these signals.

Similar long-path QSOs between Western Europe and ZL are possible on 160 meters during the bottom of the sunspot cycle. But here also, the signals do *not* arrive via the genuine long-path direction (205 to 255°) but from a direction at right angles with this heading (right over North America).

Fig 1-42 — Great-circle map centered on San Francisco, showing the 160 meter path for contacts into northern Scandinavia around 1500 UTC in midwinter (K=3) . The actual path is a crooked one. At both ends of the path, the direction is perpendicular to the nearby terminator. Propagation along the terminator in the twilight zones is impossible because of the auroral oval and additional D layer absorption. (Map generated by *DX Atlas*, with additions by ON4UN.)

4.3.1.3. South America Across North America in Northern-Hemisphere Winter

A similar path bending is also quite common on 80 meters over shorter paths. During the European winter, signals from Argentina and Chile regularly arrive in Europe at beam headings pointed directly at North America, up to 90° off the expected great-circle azimuth. The signals from South America appear to travel straight north in order to "escape" the summer conditions in the Southern Hemisphere, and are then propagated toward Europe. One striking example was when I worked 3Y1EE (Peter 1st) on 80 meters (January 28, 1987). The signals were totally inaudible from the great-circle direction (190°) but were solid Q5 from 310° (signals coming across North America). Similarly, when I worked CEØY/SMØAGD on 160 meters (October 1992), signals were only readable on a Beverage beaming 290°, while the great-circle direction to Easter Island is approximately 250°.

4.3.1.4. The Skewed Path between Europe and the US West Coast

Early on, some people believed that "long path" on 80 meters between the US West Coast and Europe followed the gray line terminator. We now know that what we commonly call a long-path QSO on the low bands actually involves signals transmitted in a direction different from that directly opposite to the short-path direction. The signals cannot actually follow the terminator, since absorption inside the gray line is much too high. We can safely say that low-band signals never actually propagate far *inside* the gray line. (See also Section 2.4.4.1). In addition, in this case the straight short path would go right through the auroral donut, which is very unlikely as well (it is possible on 40 meters though).

Nowadays, everyone acknowledges the existence of crooked (bent or skewed) paths. This means there is not only a short and a true long path, but any number of *alternative* paths that may be available for propagation. The so-called long-path on 80 meters between Scandinavia and Eastern Europe to the US West Coast is an example of such a crooked path. Looking at the darkness distribution on Earth for this in **Fig 1-42** and **Fig 1-43**, it is clear that a genuine (reciprocal) long path is out of the question, since signals would have to travel for more

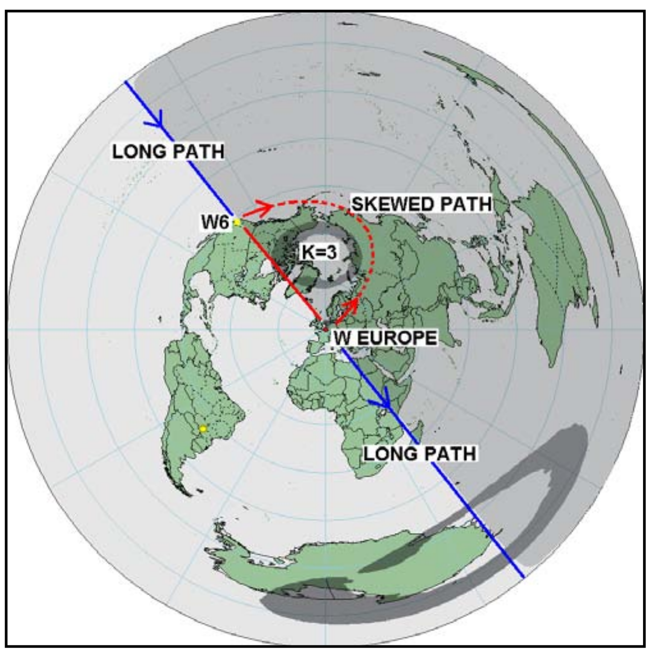


Fig 1-43 — The 80 meter long path between Europe and the US West Coast in midwinter around 1530 UTC (K=3) is neither a short path (the line running through the auroral belt), nor a genuine long path (>20,000 km in daylight). Instead, it is a crooked path. In Europe signals generally arrive at headings of 70° to 110° from True North. (Map generated by *DX Atlas*, with additions by ON4UN.)

than 20,000 km in daylight or in the gray zone.

Fig 1-42 also shows straight short path along the terminator, and going slightly through the auroral oval. If the K-Index has been very low for sustained periods, there may be such a propagation path on 80 meters, very similar to the path between Western Europe and Japan at European sunrise (see Section 4.3.3.1). The fact that the signals have to travel in the gray line zone causes additional D layer attenuation, which makes the path virtually impossible for 160 meters though.

In Europe the beam headings generally indicate an optimum azimuth angle of approximately 90° to 120°, which again is almost perpendicular to the terminator (see Section 2.4.4.). Along their way, the signals will be least attenuated in those areas of the ionosphere where the MUF is lowest, and thus they seem to travel along a crooked path, avoiding areas of higher absorption. OZ8BV reported a 90° to 100° heading when working the West Coast on 80 meter long path from southern Denmark. Ben was using a 3-element Yagi at 54 meters (180 feet) and is well placed to confirm this path (the genuine long path would be 150 to 160°).

D. Riggs, N7AM, using a rotary quad for 80 meters wrote: "We have learned that the 80 meter long path between the Pacific Northwest and Scandinavia is following the LUF (lowest usable frequency). I have always believed that the long path to Europe was not across the equator but leaves us at 240° and since the MUF is highest at the equator it cannot continue at 240° but it bends westerly going under the Hawaiian islands, across the Philippines under Japan and across the Asian continent to Scandinavia. The MUF charts prove this fact. The fact that the long path to Europe lies north of the equator is proven by the northern Europeans and after the West Coast peak."

So-called long-path QSOs have been made on 160 meters between the northern part of the West Coast of the USA (N7UA) and northern Europe (Scandinavia and the UK). Neil, G4DBN, was one of the lucky ones to have done that from the UK, and he wrote: "Bob, N7UA, and I had a QSO at around 1505 Z on 29 December 1997 and he was only audible on my northwest receiving antenna." That does not look like a typical azimuthal direction for long path, which should be approximately 150°. This QSO happened on a crooked path, almost but *not* following the gray line. It was outside the gray line at some distance from it because the additional D and E layer attenuation inside the gray line would make propagation over such distance impossible.

Similar QSOs from Scandinavia are probably easier than from anywhere else in Europe based on much 80 meter long-path experience, where it happens every day, all through the sunspot cycle, for quite a few months in wintertime. On 160 meters, long-path QSOs between the US West Coast and either OH or SM have not been really commonplace, but have occurred a number of times. These occur during both high and low sunspot cycle years.

During the 1997-1998 season a number of QSOs were made between the US West Coast and Scandinavian stations on 160 meters. SM4CAN, SM4HCM and SM3CVM all confirmed that the signals were coming from due east (90°), instead of the short path (335°) or long path (145°). N7UA notes that he was using his JA Beverage, since the SM stations were not audible on the over-the-pole European Beverage. SM4HCM called it "*a skewed path, somewhere between long path and short path.*" (See **Fig 1-44**). N7UA added that "…the lower the frequency, the more the long path moves toward north, away from the true reciprocal heading." These observations seem to be related to the principle that the lower the frequency, the more easily a horizontal ionization gradient can cause path skewing (see also Section 4.3.1.1. and Ref 147). This is supported by the fact that the amount of refraction incurred by an electromagnetic wave by a given gradient is inversely proportional to the square of the frequency.

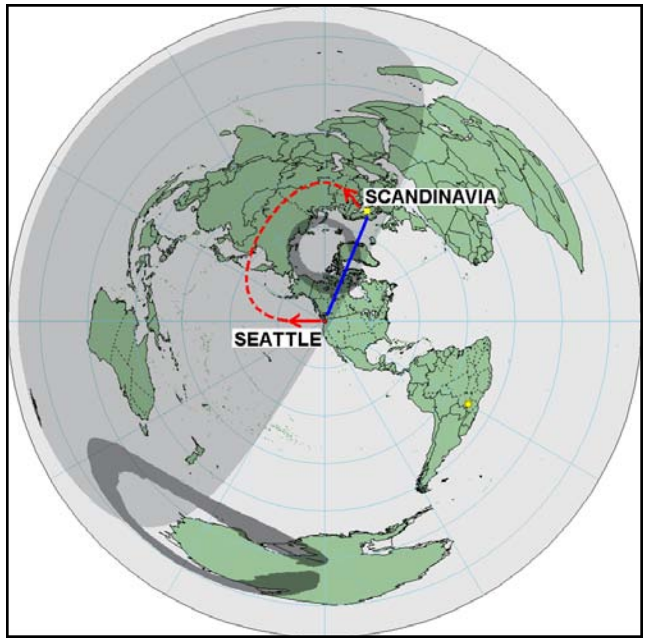


Fig 1-44 — Great-circle map centered on Seattle, showing the 160 meter path for contacts into northern Scandinavia around 1500 UTC in midwinter. The actual path is a crooked one. At both ends of the path, the direction is almost perpendicular to the nearby terminator. Propagation along the terminator in the twilight zones is impossible because of the additional D layer absorption. (Map generated by *DX Atlas* with additions by ON4UN.)

The historic QSO made during the winter of 1999 between N7UA and 5B4ADA is an interesting case. At the time of the QSO, 5B4ADA's sunset was 1436 UTC and N7UA's sunrise 1552 UTC (76 minutes of common darkness). Notice that the QSO (at 1510 UTC) was almost halfway between those times. It seems to be quite common for the path to peak near the mid-point between Europe/Asia sunset and North America sunrise, and it seems to confirm the observation that for best signal-launching conditions the path direction must be at right angles to the terminator. This keeps the signal as far as possible away from the lossy gray line zone (see also Sections 2.4.4 and 4.3.1.). Similar true long-path QSOs were also made between 4X4NJ and the US West Coast in that same period.

Bill, W4ZV, reports a number of what he calls long-path QSOs on 160 (with JT, UA9, S2, XU, XZ, 3W5, 4S7 and 9V1). All occurred with a common darkness path varying between 59 and 109 minutes (see Figs 2.23 and 2.24). Bill also remarks that QSOs over the short path or over near-polar regions seem to be best during low sunspot years, while the so-called long path seems to peak up in higher sunspot years. Scientists owe us an explanation for that remarkable and valuable observation.

While 80 meter long-path QSOs between the US West

Coast and Europe occur every winter on an almost-daily basis throughout the sunspot cycle, short-path QSOs (at USA sunrise) only occur when the following conditions are met:

- Near sunspot-minimum years when absorption is minimum.
- When the geomagnetic field is extremely quiet ($A_p < 5$).
- Centered on the period of maximum Winter Solstice darkness (December 21).
- Most common to stations located in the northern part of Europe.

It's clear that the short path suffers from the auroral oval, while the so-called long path stays clear of it!

4.3.1.5. The Skewed Path between Europe and Alaska

Besides the short path between Europe and KL7, we Europeans can often work Alaska on a non-genuine long path before their sunrise and just after our sunset on 80 meters. At that time (around 1600 UTC in midwinter) signals usually arrive in Europe from the east/northeast. This is clearly a bent path across Siberia to Alaska, thus avoiding the auroral belt (**Fig 1-45**), and is certainly *not* a true long path. If the path were a genuine long path, it would go right across the South Pole, which is in continuous daylight in the Northern Hemisphere winter. KL7Y noted the signals come in from the southwest, approximately 45° from the genuine long path.

But when geomagnetic conditions have been quiet for a long period, it sometimes *is* possible for signals to travel through the heart of the auroral zones. I remember an amazing QSO in the 1970s with KL7U on 80 meters at about 1600 UTC, hearing him only when listening at 350°, which is the direct short path right across the magnetic North Pole (the straight-line short path in Fig 1-45). Going only by the time of the contact, this would usually be called a long-path QSO; however, it was not, since the signals did not come in from the true long-path direction (approximately 160°) but almost from the regular short-path direction. It is obvious that this can happen only when there is no auroral absorption at all, since this short path goes right across the magnetic North Pole.

4.3.1.6. The Heard Island Case

The VKØIR example has been covered in detail previously in Section 2.4.4.2. From the viewpoint of path skewing, the path is very similar to the one between Europe and New Zealand (Section 4.3.1.2.). For several days the signal appeared to be traveling through areas of low MUF (although we know that 160 meter propagation is not directly MUF related). Therefore the path from Heard Island to the US East Coast seems to travel as much as possible through the Northern Hemisphere, avoiding higher MUF areas in the south. The bent path also meets the most advantageous launching conditions where the path direction (at both ends of the path) is perpendicular to the terminator at those points (see Section 2.4.4.1.). This explains why these signals were received on the US East Coast via a bent path across Europe in January 1997 (see Figs 1-20 and 1-21).

4.3.2. Skewed Paths Avoiding the Auroral Zones

The second reason for path deviation is to avoid the auroral oval. When there is aurora in the Northern Hemisphere, signals on the low bands, if not totally attenuated, will often appear to arrive from a more southerly direction than you would expect from great-circle considerations. The path between North America and Europe is greatly affected by this phenomenon, because the magnetic North Pole lies right on that path. Between Japan and Europe there is much less influence.

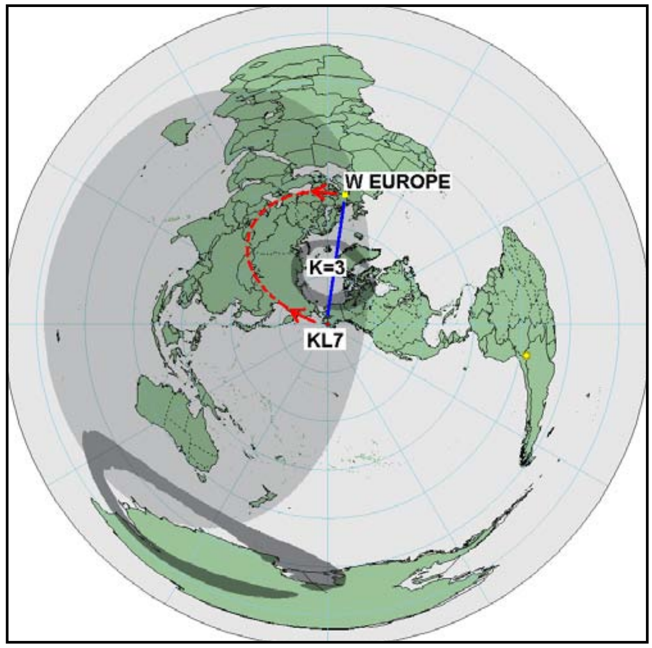


Fig 1-45 — The Alaska-to-Western-Europe 80 meter path around 1630 UTC in midwinter. The real long path is totally impossible because it travels for 25,000 km in daylight. The genuine short path travels straight through the auroral donut; hence there is a high degree of absorption at almost all times. The most likely path takes off from Alaska in a westerly direction and arrives in Europe east of the short-path bearing. Notice again that these two paths are almost perpendicular to the terminator. (Map generated by *DX Atlas* with additions by ON4UN.)

The aurora was described in detail in Section 3.2. Let us analyze a few paths that suffer frequently from the effects of aurora. To view these paths, get your globe, maps or switch on your mapping program that can show the auroral oval. With *DX Atlas* (see also Section 5.2.1) you can plot the great-circle paths, enter various values for the K-Index and then watch the evil effects of the auroral oval.

The short path between the West Coast of the US and Western Europe has always been a difficult path, because of the interference of the auroral oval with the great-circle direct path. For this reason, the short path is very rare between the West Coast and northern Scandinavia, which is inside the auroral oval.

Fig 1-46 shows the path between Western Europe and Seattle (in midwinter) at sunrise in Europe, for a K-Index of 0. The auroral oval has retracted to its minimum size and width. Although the great-circle path goes through the auroral oval, attenuation may be "limited", either because the signal underskirts the ionosphere at the narrow oval or because of slight bending around the oval. **Fig 1-47** shows the same for a K-Index of 9 (heavily disturbed magnetic conditions). Note the extreme width and major-axis size of the oval. Under such conditions, since Seattle actually lies on the border of the oval, there may be little escape from the aurora. As a rule, stations further south (eg, Southern California stations) may possibly make it into Europe, beaming across South America.

I remember one striking case like this where I made a good QSO with W6RJ on 80 meters. Bob was using a 3 element Yagi beaming to South America, while I was using my South American Beverage to copy him. The direct path, across the auroral oval, was totally dead at that time. Obviously the attenuation on this crooked path is greater than on a straight path when there is no auroral absorption, so only stations with good antennas and some power may regularly experience this path.

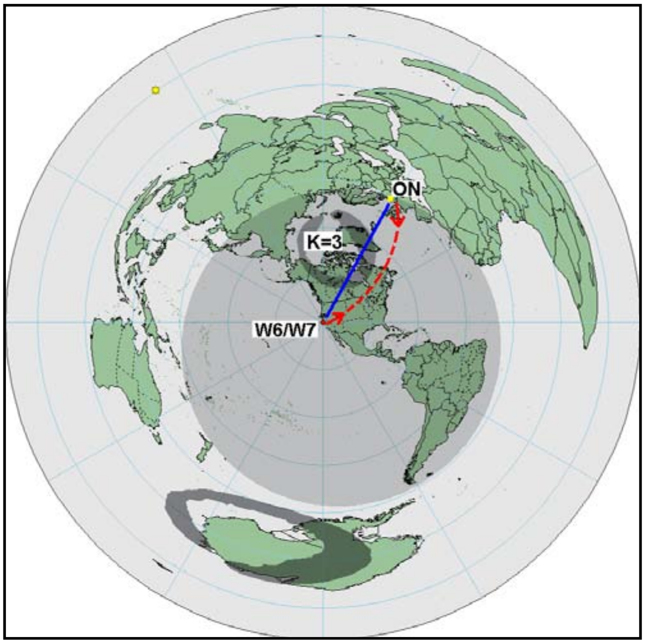


Fig 1-46 — The West-Coast-to-Europe "short" path on 80 meters in midwinter at 0730 UTC, for K=0 The signals either travel the genuine direct path, or maybe very slightly bent southward. (Map generated by *DX-AID*, with additions by ON4UN.)

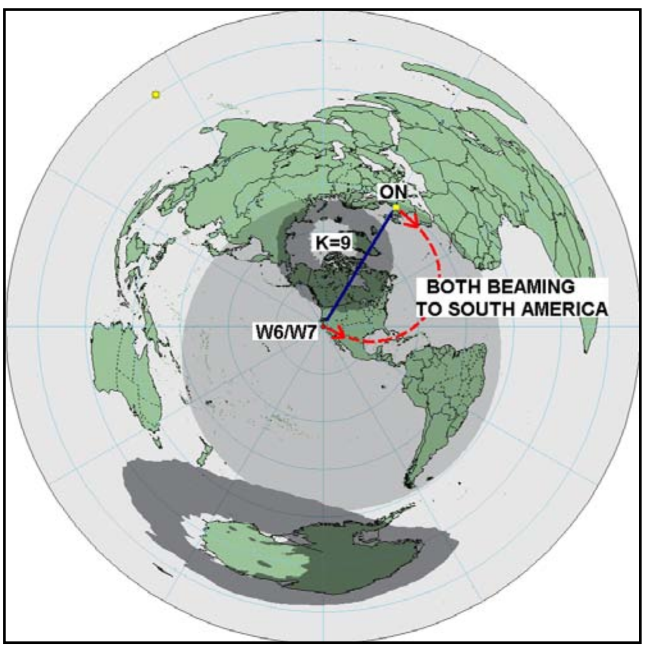


Fig 1-47 — Same midwinter situation is in Fig 1-46 but for K=9 (major disturbance). Note the extent of the auroral zone. The only way to make a QSO into Europe is for stations at both ends of the path to beam toward South America. (Map generated by *DX-AID*, with additions by ON4UN.)

Even when the auroral activity is very low, Bob finds that the short path between Northern California and Western Europe (a great-circle path of 25°) hardly ever peaks that far north, and he confirms that 75% of the time the path goes right *toward* South America, and the rest of the time peaks at around 40°.

An even more striking example of path skewing due to aurora is the path between Europe and Alaska on 80 meters. Looking at the map or globe, there is a great-circle path that is only about 7500-km long, but it beams right across the magnetic North Pole. The distance is similar to the distance between Western Europe and Florida. Straight short-path openings are rare exceptions, happening only a few times a year, when the K-Index has been at zero for some period of time. The main difference between the Seattle and the Alaska case is that from Seattle it takes much more path bending to go around the oval. Because of its more northerly location, Alaska actually lies *inside* the auroral oval. There is, of course, also the so-called long path between Europe and the US West Coast and Alaska. This was covered in detail in Section 4.3.1.5.

W4ZV, who operated for many years from Colorado as WØZV, points out that he never saw skew from Colorado to Japan, since the JA bearing (315° from Colorado) is nowhere close to Magnetic North (13° from True North). He also added that he worked Europe *much* more frequently from Colorado peaking on his 70° Beverage than his 40° Beverage, which was the true great-circle bearing from there. Under very severe geomagnetic conditions (K-Index = 6), signals would even peak on his 110° Beverage! W4ZV, now in North Carolina, confirms that a similar path exists from the US East Coast to the Far East. Signals from JA quite frequently skew to the south during geomagnetic disturbances (see also Ref 159).

Similar experiences are told by N5JA (Texas), who said that he found the best direction for UU4JMG changing from 40° at 0145 UTC, to 60° at 0215 UTC, to 90° at 0230 UTC and eventually to 120°. He reported shifts going the other way as well: "I've seen ON4UN in years past become first audible on the 90°, then be best on the 60°, and later best on the 40° or 20° Beverages." (What goes up must come down!)

KØHA, a top-notch Top Bander from the "black hole" of the USA (Nebraska), has had similar experiences copying European stations best on his 140° Beverage rather than on his 43° or 86° antenna.

Another striking example occurred on 160 meters during the first night of the CQ 160 Meter contest in January 1991: With the exception of VE1ZZ, not one North American station was heard until 0400 UTC. At that time North American stations started coming through rather faintly, but they were only audible when beaming to South America (240°). No signals from the *usual* 290° to 320° direction! Between 0400 and 0700 UTC, 80 W/VE stations were worked in 25 states/provinces. All of the signals came through across South America, including K6RK in California. On the North American Beverage only a few of those stations would have been worked.

The W4s in the southern part of the East Coast normally have a tough time piercing through the New England wall of signals trying to get into Europe, except when the aurora is on, says N4UK. That's the only time he can beat the W1s and W2s into Europe on a path skewed to the south. He thinks the skewed path is a great equalizer. It explains that when the aurora is hitting, he can still *bend around* it on his southerly skewed path, while the W1s and W2s are too close for skewing.

Steve, VE6WZ, who lives right below the auroral belt, says there is not much skewing around the belt on 160 meters from his place. "On 80 meters, I have noticed that in the last three years every 80 meter contact from VE6 into Europe has been skewed, peaking to the E-SE (not the direct path NE)." When he copied Wolf, DF2PY, on 160 meters he was definitely best on the SE Beverage (South America) and very poor on the N-NE Beverage (Europe).

Bill, W4ZV, remarks: "The longer the path, and/or the closer to crossing near the magnetic poles, the more often they skew. It's common here to hear JAs on two paths, one west or southwest and the normal northwest path. Often there is multipath as well, a fairly common phenomenon with JAs arriving from both 210 and 330 degrees."

It is clear that all these skewed paths are caused by magnetic disturbances, as they all happened during periods of fairly high geomagnetic activity. In general, you can say that path skewing on relatively short paths (<10,000 km) is quite common during disturbed conditions. Skewing can be anything from 30° to almost 90° further south of the normal great-circle direction (for stations living in the Northern Hemisphere).

For a station at high or mid-geomagnetic latitudes there are only two ways on 160 meters to make a QSO into an area that is completely hidden behind the auroral donut. Either you wait for the day when auroral absorption is very low or nonexistent (this happens a couple of days during each 11-year cycle) and work the station right across the magnetic pole. Or you find your way around the problem. That is, you work the station on a crooked path going around the auroral oval. Under such circumstances signals often appear to be coming long path (the direction the signals arrive from are close to opposite from the short-path great-circle direction) and likely travel a long distance on a crooked path. By traveling *around* the auroral donut they avoid auroral absorption, but by traveling a very much longer distance (than their direct great-circle path) they suffer considerable additional attenuation.

Along the same lines W4ZV wrote: "This can be done even in high flux years and even under relatively high geomagnetic activity. Otherwise, we must wait for the very bottom of the cycle and only on days with K_p of 0 or at most 1 to work through the auroral oval to the other side. Taken over an entire solar cycle, these days are rare indeed that contacts may be made by short path through the auroral oval."

On several occasions these crooked (long) paths have been reported to be quite unstable, and quite abruptly switch directions (over 90°), especially right at or after sunrise. Make sure to use all your Beverages and keep twisting that selector knob.

4.3.3. Crooked Polar Paths in Midwinter, or Pseudo Long Paths

4.3.3.1. Europe to Japan

In northern Europe we can work Japan in midwinter, just after our sunrise (0745 UTC) on 80 meters. The most common opening to Japan on 80 is at JA sunrise, around 2200 UTC. At first you might be tempted to call the 0745 UTC opening a long-path opening. Careful analysis using directive receiving antennas has shown that the signals come from a direction slightly east of north rather than their true long-path heading of 210°. The opening is rather short at my QTH (typically 15

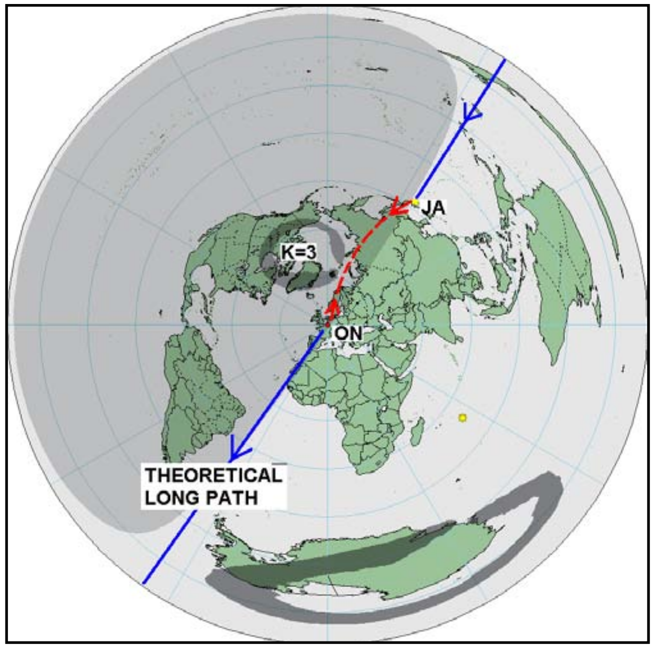


Fig 1-48 — The direct short path between Europe and Japan at Europe sunrise in midwinter is an example of quite exceptional propagation. It is clear that the theoretical long path is out of the question because it goes through daylight or gray zone for not less than 21,000 km. One might consider a crooked path bent "around" the auroral donut, but that path has never materialized. The signals (between ON and JA) always come in on a short path heading. (Map generated by *DX Atlas*, with additions by ON4UN.)

to a maximum of 30 minutes in midwinter during low sunspot cycle years and almost nonexistent during high sunspot cycle years). **Fig 1-48** shows the great-circle path *along* the terminator, along the gray line zone, and is *not* a typical example of gray line propagation, where low-band signals usually travel more or less perpendicular to the terminator.

These are certainly far from ideal conditions for low attenuation and the signals on the JA path are substantially weaker than the signals normally heard from Japan from the same heading, but at JA sunrise. Low-band propagation over long distances along the terminator is clearly not the rule, because of additional D layer absorption (see Section 2.4.4). If present, signals are weaker than normally expected over similar distances on paths that do not follow the terminator.

If we look at the darkness/daylight distribution across the world at that time (0745 UTC in midwinter), we see that we have indeed more than one path possibility: Paths ranging from true short path (30°) to alternative crooked paths bent slightly east or even west of the magnetic North Pole, all across areas in darkness. These alternative bent paths go right through the North Pole auroral zones, and hence will very seldom produce stronger signals than the path along the terminator.

A number of years ago K6UA, Dale Hoppe (SK) and others (Ref 108 and Ref 118) considered that low band gray line propagation went *along* the terminator, as is the case on the higher bands. It has, however, been proven over and over again that this is *not* the rule on the low bands (see Section 2.4.4.1), where the most spectacular propagation enhancements seem to occur when the propagation path is *perpendicular* to the terminator. The Europe-to-Japan short path at 0745 UTC in midwinter is a remarkable exception to the general rule.

Often such specific propagation paths (more or less along the terminator) are very area selective, probably because ducting phenomena are involved (see Section 2.4.4.3), occurring only when the necessary launching and exiting conditions exist. In several cases I was able to work JA stations with signals up to S9, while the same stations were reported to be undetectable in Germany only a few hundred kilometers away. In all cases the signals were loudest when they were coming in about 10° east of north. In Japan the openings seem to be very area-selective as well, as can be judged from the call areas worked. Northern Japan (JA7 and JA8) obviously leads the opening. Central Japan follows 30 minutes later, and southern Japan often is too late for this kind of opening at my QTH. In midwinter there is a 1½-hour spread in sunset time between northern and southern Japan.

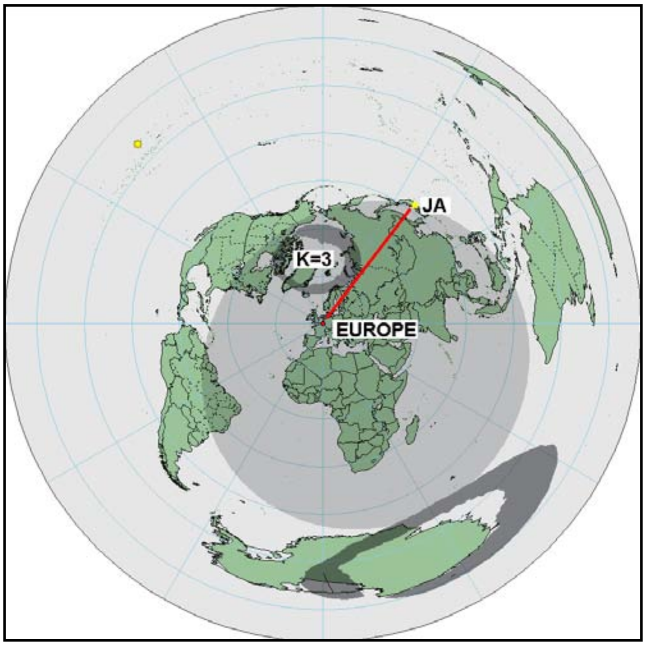


Fig 1-49 — The classic path between Western Europe and Japan for sunrise in Japan (midwinter, 2200 UTC, K=3). Note that the path direction is almost perpendicular to the terminator in Japan. (Map generated by *DX Atlas*, with additions by ON4UN.)

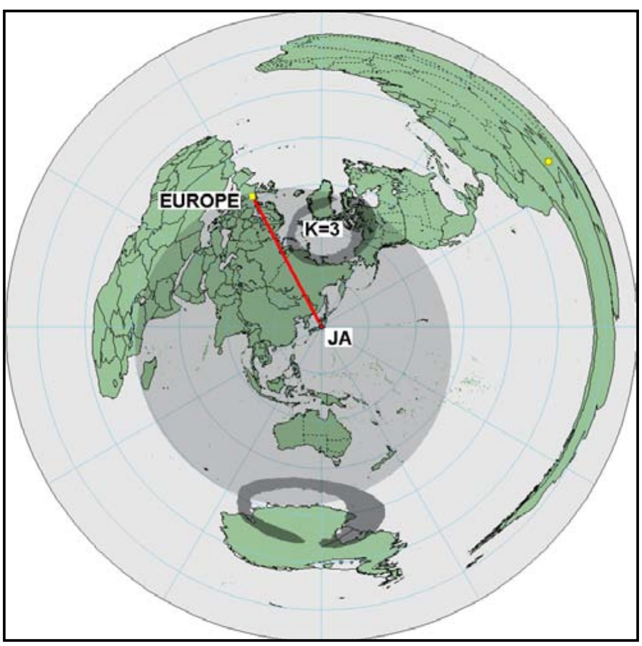


Fig 1-50 — The other path between Europe and Japan, at sunset in Europe (1600 UTC, midwinter, K=3). Here too the terminator is almost at right angle to the propagation path. (Map generated by *DX Atlas*, with additions by ON4UN.)

On 40 meters the situation is somewhat similar, and yet very different at the same time. At 0745 UTC in midwinter, JA 40-meter signals arrive from north or northeast (as on 80 meters), but when that path fades about 15 to 30 minutes later, it is immediately replaced by a genuine long path on 40 meters, where the signals now come in across South America, along the terminator as shown in **Fig 1-48**. I have never observed this genuine long path across South America on 80 meters, let alone 160 meters.

At JA sunrise time (2200 UTC) the short-path direction is almost at right angles to the terminator (see **Fig 1-49**). A similar good launching angle (with respect with the terminator) occurs at sunset time in Europe around 1600 UTC (**Fig 1-50**). From that point of view the 1600 UTC and the 2200 UTC openings are almost identical. In real life however, we find in Europe the 2200 UTC opening the better one. The reason is obvious: At 2200 UTC we have a sunrise peak, while a sunset peak is much less pronounced (if it occurs at all). Further, 1600 UTC is in the middle of the night in Japan.

4.3.3.2. New England to Japan

Brown, NM7M, analyzed a seemingly similar path (Ref 140) on 160 meters: W1 to JA. It is quite different however, since the direct path between New England and Japan goes through the auroral donut, even for a relatively low K-Index of 3. He also found two openings. One he calls a "gray line" path (the opening around 2140 UTC in midwinter) and another one that he calls a "black-line" path. I consider the former path as a most atypical gray line path for low frequencies, even though both ends are in twilight (a double gray line situation). NM7M claims that the few signals heard or worked on such occasions came out of the northeast direction, which is a crooked path across Europe. See **Fig 1-51**. K1ZM, however, who actually made these QSOs, said the path was from the southeast and not over Europe (both in December 1996 and January 2000). The moral of this story: Make sure you have receiving antennas covering *all* directions, and keep switching them.

The theoretical great-circle path between W1 and Japan at 2140 UTC (East Coast sunset in midwinter) goes parallel with the terminator, hence suffering severe attenuation due to the D layer already building up in that zone. The alternative crooked path travels across Europe (according to NM7M) or even further to the east and southeast, as witnessed by K1ZM. Whether beaming NE or SE, the signals paths are nearly perpendicular to the terminator and hence enjoy excellent launching conditions, since they spend little time in dusk/dawn where D layer absorption is enhanced.

The great-circle path from New England to Japan in midwinter at 1200 UTC makes an angle of approximately 50° with the terminator, and it is common knowledge that many of these

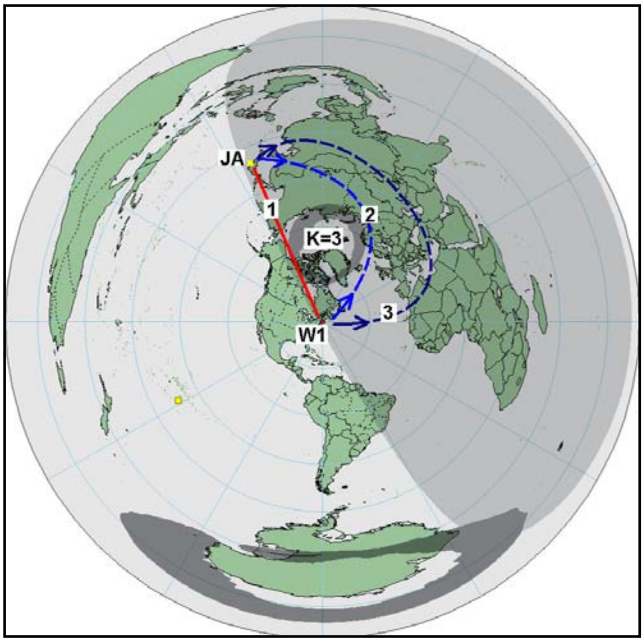


Fig 1-51 — Three propagation paths (for both 160 and 80 meters) from New England to Japan (at 2140 UTC, January 1). Path # 1 is the theoretical straight-line short path, which goes along the terminator and for a good distance through the aurora oval even with a K as low as 3. That makes it an impossible path. Path # 2 is the path NM7M quotes and which goes right over Europe (NE), while those who made the contacts received the signals from E-SE (path # 3). (Map generated by *DX Atlas*, with additions by ON4UN.)

paths are bent southward (many US East coast stations claim paths from 270° to 205°), making the angle with the terminator more like 90°. The 1200 UTC opening is a one-sided gray line signal enhancement (**Fig 1-52**), since it occurs at East Coast sunrise while Japan is still fully dark.

QSOs have, of course, been made much earlier (0930 UTC). At that time we have a most typical *black-line* path (see Section 2.4.2) occurring at midnight for the point halfway between the path ends (**Fig 1-53**). Here too, a slight skewing around the polar regions is very common. At this time of the day there is no signal enhancement by gray line phenomena, since both ends of the line are well into darkness.

4.3.3.3. Other Paths

Similar paths exist on 80 meters in midwinter between California and Central Asia (Mongolia) around 0030 UTC, between Eastern Europe (Moscow) and the northern Pacific (Wake Island) around 0615 UTC, and between the East Coast (and the northern part of the Midwest) of the US and northern Scandinavia around 1230 UTC. All of these polar-region paths are east of the North Pole and should not be influenced by aurora as much as paths going west of it. Use your globe, map or mapping program (eg *DX Atlas*) to visualize the paths that appear to avoid the auroral belt.

4.3.4. Selective Paths/Areas

I have often wondered how a DX station, or worse yet a DXpedition, can make low-band contacts while hoards of stations keep calling after a QSO has started. We've all seen cases where 80% of the pileup just keeps calling and calling (especially in Europe). For an observer located in the middle of all these callers, this is pure chaos.

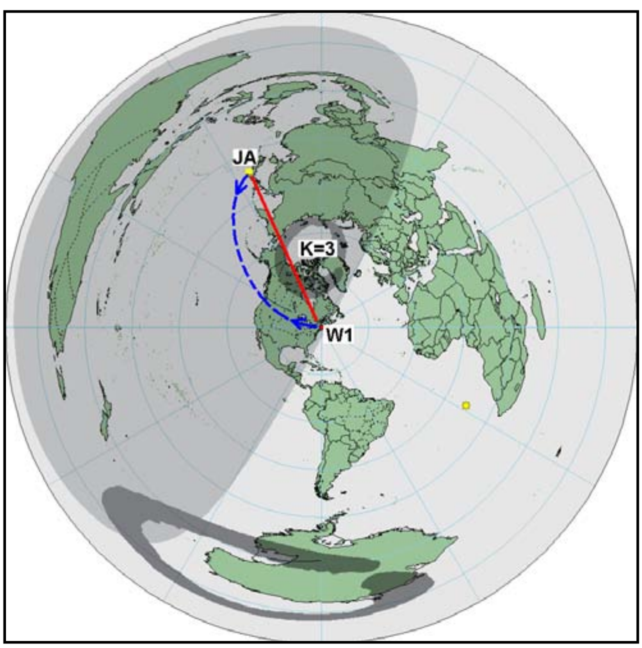


Fig 1-52 — The New England to Japan path (for 80 and 160 meters) at 1200 UTC (K=3). Actual signals arrive in New England from beam headings somewhat further west than the short path heading in order to skirt the auroral oval. (Map generated by *DX-AID*, with additions by ON4UN.)

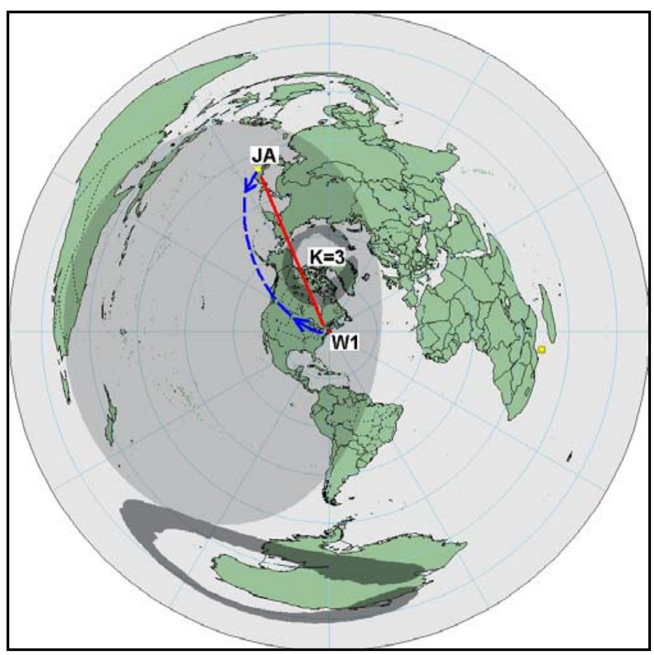


Fig 1-53 — The New-England to Japan "all night" (black line) path at 0930 UTC (K=3). At this time as well the beam heading is often bent further west than the direct path to avoid the auroral oval. (Map generated by *DX-AID*, with additions by ON4UN.)

I am convinced that *selective skip* is the reason why the DX station can work stations despite this chaos. ZS6EZ, an eminent lowbander and DXpeditioner stated: "On my DXpedition, it was very often like a searchlight, where only one very small area is audible at a time, and that area moves around with time. There is no regular movement either. You might have W1, W8, W9, WØ, W3, W4, W5, WØ, W7, WØ, W9, etc in a single opening. The pileup is never audible; just a few stations at a time." He adds: "This is why I contend that the argument of the 'East Coast Wall' doesn't hold in most sunrise openings to Africa. If the 'spotlight' is on the West Coast, the other coast is not even audible."

John, K4TO, commented to me on the same subject: "I have also observed that given propagation into a particular area, one signal might be 10 dB stronger than another, even though both stations were running the same power and comparable antennas. Given a few seconds time lapse, the situation may reverse. The station that was stronger earlier on is now the weaker of the two. Having observed this first hand, I find that my own anxiety level about working a particular DX station is now reduced to near zero. I feel that sooner or later, the propagation will come to favor me. I wish more DXers would have this same experience. I believe everyone would relax and enjoy the hobby more. Operating manners and techniques would improve greatly."

I think that this area selectivity may also be partly due to polarization rotation of signals on 160 meters. This can also explain the slow QSB that we often witness on Top Band. Sometimes we must wait a few minutes for polarization to "come back" in order to hear a marginal signal again. This is why on Top Band, where weak signals are involved, it is often essential to get a full call at a first try. Often a second try at a partially received call brings dead silence, until a minute or so later the same signal slowly comes out of the noise again.

Signal ducting between the E and F layers provides another possible explanation for the "searchlight effect." The landfall of the searchlight would relate to finding stations with appropriate launch angles, such that the RF could enter (or exit) the duct.

A most striking example was a QSO I made on March 13, 2003, with VP6DIA on 160 meters. During the 15 minutes before my local sunrise, his signals were solid Q5 and I worked him on a single call. I heard nobody else in Europe calling, and I was in touch with a number of East Coast stations on the Internet Top Band chat channel. They all confirmed not hearing a beep, while the signals remained solid Q5 all the time here.

The late Dan Robbins, KL7Y, came up with another theory that might help explain selective propagation. He concluded from his work with HF radars that the ionosphere can act as a "filter" for the angle of radiation: "Near the MUF, high angles are 'filtered' out. But below the MUF, low angles may also be filtered out. Once angles are filtered out they are gone, and the same narrow range of angles will propagate over and over again, even if the filtering conditions disappear on later hops. If the range of allowable angles is narrow, the propagation will occur in narrow distance bands from the transmitter, since distance is a function of angle of radiation. This is why one guy works the S9 DX and his buddy 300 miles away hears nothing. This is why the band appears dead, except for that loud 3B8 or whatever."

In other words, for any given path there will be an optimum frequency and angle of radiation that produces the maximum signal. If we are restricted to one frequency then there will be one range of optimum radiation angles for that path. According to Dan Robbins the optimum angle may be surprisingly narrow at times — narrow enough to account for almost all instances of selective propagation.

Fortunately, most of the antennas we use on the low bands have rather broad vertical lobes, nothing like the arrays used for OTH radar. This effect of "angle selectivity" is thus largely smoothed by the broad lobe of our transmit antennas.

4.3.5. Path Skewing: A Summary

On the low bands, and more specifically on 160 meters, we continuously witness bent propagation paths (crooked or skewed paths). At one end of the path (or more likely at both ends of the path), a signal arrives from a direction that is substantially different from the great-circle heading. Documentation of these facts is so overwhelming that there is no doubt about these bent paths.

Robert R. Brown, NM7M, wrote on the Top Band reflector: "In summary, path skewing results from horizontal gradients in electron density. For interpretation, the problem is to locate the gradient and identify its source. Without any gradient or source, reports of skewing are incomplete." I think he wanted to say that without an identified gradient or source having caused the gradient, scientists cannot explain the detailed mechanisms that intervened in that particular case. It goes too far to say that the report is not complete and to question the sincerity and accuracy of these reports.

I am a radio amateur, not a scientist, and so I take issue with NM7M's above statement. Most of my friends, serious and dedicated low band DXers, who, over the years, have reported skewed paths are serious hams, although not all of them are scientists. If only scientists are qualified for reporting skewed paths, we would have had very few reports over the years! By looking at the similarity of many reports hams can benefit from them, without necessarily knowing the source of the gradient(s) causing the bending. So, fellow low band DXers, keep on reporting such odd paths on the Top Band reflector.

For the low-band operator it is important to know that these bent paths are very common, both to be able to anticipate them and to make good use of them. Many of the observations are clearly linked with high geomagnetic activity. In these cases signals arrive from directions away from the auroral zones, which seems logical. In this case some sort of horizontal ionization gradients are causing reflection and bending of the path away from the auroral wall.

In an earlier edition of this book I wrote, "What is not so clear to me is how waves, bent away from the auroral donut (bent southward in the Northern Hemisphere when you are located south of the donut), are bent back north in order to *skirt around* the auroral belt. If that were not the case, a target that is *behind* the auroral donut would still remain unreachable. In my opinion, such crooked paths must involve two bending mechanisms, one bending away from the auroral zone, and one bending back in northerly direction beyond the auroral donut." We should all be grateful to Carl, K9LA, for digging into this mystery and writing about it (Ref 159). Carl's attitude versus ham's observations which have not so far been "explained" by scientists is a positive one, which all lowbanders should appreciate and applaud.

In addition to these aurora-related crooked paths, there are also many documented cases of signal-path bending on paths that do not go through high-latitude areas where aurora may be present. These are clearly not related to strong geomagnetic activity. A number of witnesses of both cases are listed in Section 2.4.

Many crooked paths are hard to explain by using the present state-of-the-art (ray tracing) computer models such as *Proplab-Pro* mainly because these programs use ionosphere models that do not include the necessary detailed information (such as details on the E layer) to create an accurate model. They are also monthly median models, and do not capture the day-to-day variability of the ionosphere that can cause crooked paths.

Our present understanding of wave propagation makes it impossible to *predict* propagation along the terminator for various reasons, such as instability and enhanced D layer absorption. Since these crooked paths happen in the absence of geomagnetic disturbances, reflection by steep ionization gradients does not seem to be involved all the time.

It appears to me that a great number of possible causes and mechanisms are involved on Top Band, making our signals skew and travel along paths that are not always along a great circle. Skewed paths, a phenomenon that we all observe very regularly (maybe we do not always realize it), in fact, day after day, are real. Explaining the mechanisms behind such skewed paths is the role of the scientists. Once they know all the mechanics behind it, they may even be able to write propagation-prediction software that works, and half of the fun of DXing on 160 meters will be gone! Not having all the answers yet, however, should not stop us from taking advantage of so-called anomalies and enjoying DXing on the low bands.

4.3.6. Path Skewing: Concluding Remarks

To reiterate an earlier comment, in many of the great-circle maps used in this section, the crooked paths are represented as a nicely curved lines. This is only a *symbolic* representation of the real paths (see Fig 1-37). What we observe is the azimuth takeoff angles at both ends of the path represented in these maps. The path the signals *actually* travel on their way between the terminal points is not along a nicely curved line. If we accept a scatter-reflection mechanism to explain these bent paths, the signal may actually travel in a straight line toward the scattering area (a region with a horizontal ionization gradient of some sort). Often there is a combination of various mechanisms that determine the actual path over its entire path. The real path will likely follow a jagged line rather than a nicely curved one (see Ref 159).

Since DX signals often come in from other directions than the great-circle direction, any serious low band DXer should have a choice of receiving antennas. You must switch them continuously, searching for the direction that produces the best received signal. Once this direction is established, you should transmit in that same direction as well.

5. TOOLS FOR SUCCESSFUL DXING ON THE LOW BANDS

5.1. Sunrise/Sunset Information

To have propagation on 160 meters, the entire propagation path must be in darkness. On 80 meters we can have some daylight (which means that we can live with some D layer build-up for up to one hour or even more) after sunrise or before sunset. On 40 meters this can be several hours. But on 160 meters the general rule is that the band goes dead very shortly after sunrise (or remains dead until very shortly before sunset). This means we need to know precise sunset and sunrise times. This is especially so if we want to take advantage of the propagation lift that happens around sunrise (and to a lesser extent around sunset).

For this purpose we must always use the *exact* sunrise time and never take into account twilight times (civil, nautical, astronomical). Watch out, some mapping programs also show a twilight line; make sure you use the real sunrise/sunset time (defined by the time that the center of the sun comes at the horizon).

More than 20 years ago I made available the "ON4UN Sunrise/Sunset Tables." I still use it now and then, as I often find it easier and giving a better overview than the sunset/sunrise data that are now available on the DX spotting networks, in any logging program and many other places, especially if you are looking for a certain date or for a certain period of time.

5.1.1. General Rules for Using Sunrise/Sunset Times

For all E-W, W-E, NW-SE and NE-SW paths you can expect normally two propagation peaks (for short path):

1) The first peak will occur around sunrise of the station at the eastern end of the path.
2) The second (generally less pronounced) peak occurs around sunset for the station at the western end of the path.

For N-S paths there are no pronounced peaks around either sunset or sunrise. Often the peak seems to occur near midnight (see Section 2.4.3). The use of the tables can best be explained with a few examples.

5.1.2. Example 1

What are the peak propagation times between Belgium and Japan on February 15? From the tables:

Belgium: 15 February: SRW = 0656 SSW = 1659
Japan: 15 February: SRE = 2130 SSE = 0824

where
SRE = sunrise, eastern end
SRW = sunrise, western end
SSE = sunset, eastern end
SSW = sunset, western end

The first peak is around sunrise in Japan or SRE = 2130 UTC. This is after sunset in Belgium (SSW = 1659), so the path is in darkness. Always check this. The second peak is around sunset in Belgium or SSW = 1659 UTC. This, too, is after sunset in Japan (0824 UTC) so the path is in darkness.

5.1.3. Example 2

Is there a possibility for a long-path opening on the lower bands? The definition of a long-path opening (see Section 4.2) says we must have sunset at the eastern end before sunrise at the western end of the path. In the example this is not true, because SRW at 0656 UTC is not earlier than SSE at 0824 UTC.

o , o , ON4UN Astronomical Applications Dept.
Location: E003 45, N51 00 Rise and Set for the Sun for 2003 U. S. Naval Observatory
Washington, DC 20392-5420

Universal Time

	Jan.		Feb.		Mar.		Apr.		May		June		July		Aug.		Sept.		Oct.		Nov.		Dec.	
Day	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set	Rise	Set
	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m	h m
01	0748	1549	0722	1635	0631	1725	0522	1817	0419	1906	0336	1950	0335	2003	0410	1932	0458	1832	0545	1724	0636	1620	0725	1542
02	0748	1550	0721	1637	0629	1727	0520	1819	0418	1908	0336	1951	0335	2002	0411	1930	0459	1829	0546	1722	0638	1618	0727	1541
03	0748	1551	0719	1639	0626	1729	0518	1820	0416	1909	0335	1952	0336	2002	0413	1929	0501	1827	0548	1720	0640	1617	0728	1541
04	0748	1552	0718	1641	0624	1730	0515	1822	0414	1911	0334	1953	0337	2002	0414	1927	0502	1825	0549	1717	0642	1615	0729	1540
05	0748	1553	0716	1642	0622	1732	0513	1824	0412	1912	0334	1954	0337	2001	0416	1925	0504	1823	0551	1715	0643	1613	0731	1540
06	0747	1554	0715	1644	0620	1734	0511	1825	0410	1914	0333	1955	0338	2001	0417	1924	0505	1821	0553	1713	0645	1612	0732	1540
07	0747	1556	0713	1646	0618	1735	0509	1827	0409	1915	0333	1955	0339	2000	0419	1922	0507	1818	0554	1711	0647	1610	0733	1539
08	0747	1557	0711	1648	0616	1737	0507	1828	0407	1917	0332	1956	0340	1959	0420	1920	0508	1816	0556	1709	0648	1608	0734	1539
09	0746	1558	0709	1650	0613	1739	0504	1830	0405	1919	0332	1957	0341	1959	0422	1918	0510	1814	0558	1706	0650	1607	0735	1539
10	0746	1600	0708	1651	0611	1741	0502	1832	0404	1920	0331	1958	0342	1958	0423	1916	0512	1812	0559	1704	0652	1605	0737	1539
11	0745	1601	0706	1653	0609	1742	0500	1833	0402	1922	0331	1959	0343	1957	0425	1915	0513	1809	0601	1702	0654	1604	0738	1538
12	0744	1602	0704	1655	0607	1744	0458	1835	0401	1923	0331	1959	0344	1957	0426	1913	0515	1807	0602	1700	0655	1602	0739	1538
13	0744	1604	0702	1657	0605	1746	0456	1837	0359	1925	0330	2000	0345	1956	0428	1911	0516	1805	0604	1658	0657	1601	0740	1538
14	0743	1605	0701	1659	0602	1747	0454	1838	0357	1926	0330	2000	0346	1955	0430	1909	0518	1803	0606	1656	0659	1559	0740	1538
15	0742	1607	0659	1700	0600	1749	0451	1840	0356	1928	0330	2001	0347	1954	0431	1907	0519	1800	0607	1654	0700	1558	0741	1539
16	0741	1608	0657	1702	0558	1751	0449	1842	0355	1929	0330	2001	0348	1953	0433	1905	0521	1758	0609	1651	0702	1557	0742	1539
17	0740	1610	0655	1704	0556	1752	0447	1843	0353	1931	0330	2002	0350	1952	0434	1903	0522	1756	0611	1649	0704	1556	0743	1539
18	0740	1612	0653	1706	0553	1754	0445	1845	0352	1932	0330	2002	0351	1951	0436	1901	0524	1754	0612	1647	0705	1554	0744	1539
19	0739	1613	0651	1708	0551	1756	0443	1846	0350	1933	0330	2003	0352	1950	0437	1859	0526	1751	0614	1645	0707	1553	0744	1539
20	0738	1615	0649	1709	0549	1757	0441	1848	0349	1935	0330	2003	0353	1949	0439	1857	0527	1749	0616	1643	0709	1552	0745	1540
21	0737	1616	0647	1711	0547	1759	0439	1850	0348	1936	0330	2003	0355	1947	0440	1855	0529	1747	0617	1641	0710	1551	0746	1540
22	0735	1618	0645	1713	0544	1801	0437	1851	0347	1938	0330	2003	0356	1946	0442	1853	0530	1744	0619	1639	0712	1550	0746	1541
23	0734	1620	0643	1715	0542	1802	0435	1853	0345	1939	0331	2003	0357	1945	0444	1851	0532	1742	0621	1637	0713	1549	0747	1541
24	0733	1621	0641	1716	0540	1804	0433	1855	0344	1940	0331	2003	0359	1944	0445	1849	0533	1740	0623	1635	0715	1548	0747	1542
25	0732	1623	0639	1718	0538	1805	0431	1856	0343	1941	0331	2004	0400	1942	0447	1847	0535	1738	0624	1633	0717	1547	0747	1543
26	0731	1625	0637	1720	0535	1807	0429	1858	0342	1943	0332	2004	0401	1941	0448	1845	0537	1735	0626	1631	0718	1546	0748	1543
27	0729	1627	0635	1722	0533	1809	0427	1859	0341	1944	0332	2003	0403	1939	0450	1842	0538	1733	0628	1629	0720	1545	0748	1544
28	0728	1628	0633	1723	0531	1810	0425	1901	0340	1945	0333	2003	0404	1938	0451	1840	0540	1731	0629	1627	0721	1544	0748	1545
29	0727	1630			0529	1812	0423	1903	0339	1946	0333	2003	0406	1936	0453	1838	0541	1729	0631	1626	0723	1543	0748	1546
30	0725	1632			0526	1814	0421	1904	0338	1947	0334	2003	0407	1935	0454	1836	0543	1726	0633	1624	0724	1543	0748	1547
31	0724	1634			0524	1815			0337	1949			0408	1933	0456	1834			0635	1622			0748	1548

Fig 1-54 — There are a number of places on the Internet where you can find a tool to print out sunrise/sunset times. One source is aa.usno.navy.mil/data/docs/RS_OneYear.html

5.1.4. Example 3

Is there a long-path opening from Japan to Belgium on January 1?

Belgium: 1 January: SRW = 0744 UTC SSW = 1549 UTC
Japan: 1 January: SRE = 2152 UTC SSE = 0740 UTC

Here, SRW at 0744 UTC is later than SSE at 0740 UTC. This is indeed a valid condition for a long-path opening. It will be of short duration and will be centered on 0746 UTC. Being a so-called long path does not mean that the direction of signal arrival is the opposite of the short path! In Section 4.3, I explained that this so-called long path to Japan is a typical example of a midwinter crooked path.

5.1.5. Pre-Sunset and Post-Sunrise QSOs

In practice, long-path openings are possible even when the paths are partially in daylight. Near the terminator we are in the so-called *gray line zone* and can take advantage of the enhanced propagation in these zones. The width of the gray line has been discussed earlier (Section 2.4.4.1.). A striking example of such a genuine long-path QSO was a contact made between Arie, VK2AVA, and me on March 19, 1976, at 0700 UTC on 80 meters. The long-path distance is 22,500 km. Note that the QSO was made almost right at equinox (March 21), and the path is a textbook example of a NE-SW path. On that day we had the following conditions:

Sunrise west (Belgium) = 0555 UTC
Sunset east (Sydney, Australia) = 0812 UTC

This means that the long path was in daylight for more than two hours. The QSO was made one hour after sunrise in Belgium and more than one hour before sunset in Australia.

Another similar example was a QSO with VKØGC from Macquarie Island (long-path distance 21,500 km). On January 21, 1985, a long-path contact was made on 80 meters that lasted from 0800 until 0830 UTC, with excellent signals. This was more than one hour before sunset on Macquarie (0950) and almost one hour after sunrise in Belgium (0731). Because the locations of these stations (VK2 and VKØ) are fairly close to the antipodes from Belgium, the long paths can safely be considered genuine long paths. Indeed there are no crooked paths that could provide an alternative to the genuine long paths. The gray line globe is a unique tool to help you visualize a particular path like this. Another example (Palmyra/Kingman Reef) was described in detail in Section 2.4.4.2.

Every now and then we hear about almost magical QSOs where on 160 meters contact was made well after sunrise (UUØJZ to KL7 four hours after sunrise), but these QSOs can certainly never be predicted (see Section 2.4.4.2)!

5.1.6. Calculating the Half-Way Local Midnight Peak

For east-west paths (±45°), in addition to the usual sunrise and sunset peaks there is often a so-called *mid-way midnight peak* (see Section 2.4.3.). To calculate the time of this peak use the sunset/sunset tables or a program to determine both sunrise and sunset times for both ends of the path. For example, a path between Denver, Colorado, and Belgium on January 15. The sunrise/sunset data are:

Colorado: sunset: 0001 UTC, sunrise: 1419 UTC
Belgium: sunset: 1606 UTC, sunrise: 0738 UTC

Midnight in Colorado: 0001 + (1419 – 0001)/2 = 0710 UTC
Midnight in Belgium: 1606 + (2400 + 0738 – 1606)/2 = 2352 UTC

The halfway midnight peak time is calculated as the mathematical average between the two midnight times:

Local half-way midnight is: (2400 – 2352 + 0710) / 2 = 0339 UTC

For north-south paths (±30°) there is a distinct propagation peak at local midnight at the half-way spot for both 80 and 160 meters. This peak is commonly called the *midnight peak*. How do we calculate the exact time of this midnight peak? Example: Path between New York and Paraguay on June 15.

New York sunset: 0028 UTC, sunrise 0925 UTC
Paraguay sunset: 2108 UTC, sunrise: 1034 UTC

Calculate the two local midnight (sun)times as follows:

Midnight-New York: (0925 – 0028) / 2 = 0429 UTC
Midnight Paraguay: (1034 + 2400 – 2108) / 2 = 0643 UTC
Halfway midnight time: (0429 + 0643) / 2 = 0536 UTC.

5.2. Mapping Programs

If you wish to know whether certain paths will be disturbed, you must visualize the path as well as the auroral oval. These will immediately reveal what's going on for that path for a given auroral intensity. Maps in various projections, as well as globes, can be used. You may have to fabricate your own auroral ovals to use with the map or globe (see Section 3.2.4.).

Nowadays, when every ham has at least one computer in the shack, computer programs do just what we want, with much less hassle. There are numerous propagation prediction programs around, but only a few address aurora directly. The following three programs all have their own merits and shortcomings, but at least they address geomagnetic conditions: *DX Atlas, W6ELProp*, and *Proplab-Pro*. These three programs were extensively used to create figures used in this chapter.

5.2.1. *DX Atlas*

DX Atlas (**www.dxatlas.com**) is an excellent software mapping program by VE3NEA. See **Fig 1-55**. *DX Atlas* displays world maps in rectangular, azimuthal and 3-D globe projections. Overlays are provided for amateur prefixes, CQ and ITU zones and grid squares. For any point selected by the user, latitude, longitude and grid square are displayed. The user can select a home location, from which the heading and distance are automatically calculated (both short and long path) to the mouse cursor on the map. Maps can be zoomed in and out, and the gray line can be added to any map display. The gray line automatically reflects the current time and date (as set on the computer), or a fixed time (past or future) can be entered for a specific purpose. Sunrise and sunset times for any point on the map are also shown.

A great variety of ionospheric maps can be called up in *DX Atlas*, such as MUF (3000) map; F2 layer critical frequency map; F2 layer height; E layer critical frequency; D layer peak

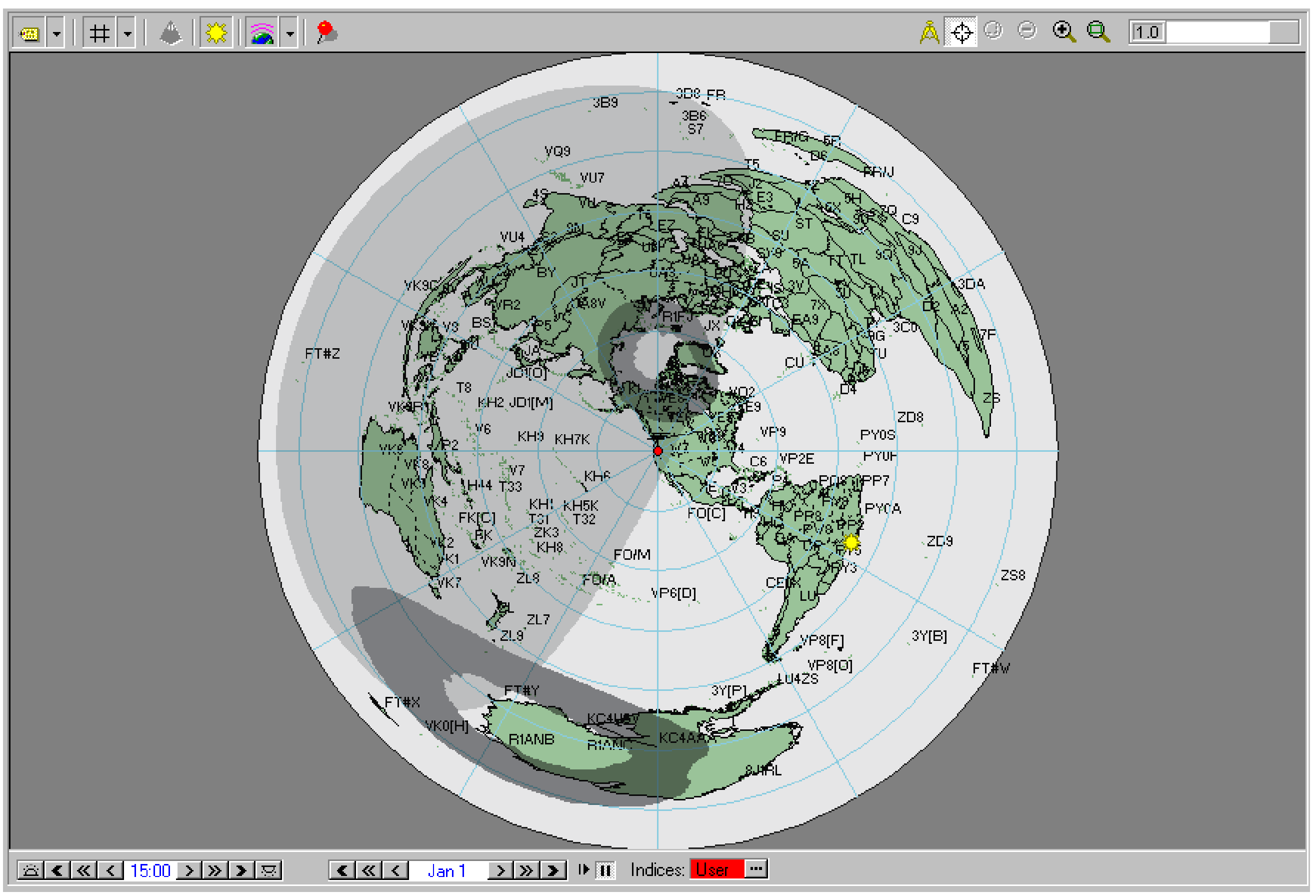


Fig 1-55 — Screen shot from the *DX Atlas* mapping program. Note the auroral ovals on this azimuthal projection. Various other map projections, and a large number of mapping facilities are available. (Courtesy *DX Atlas*.)

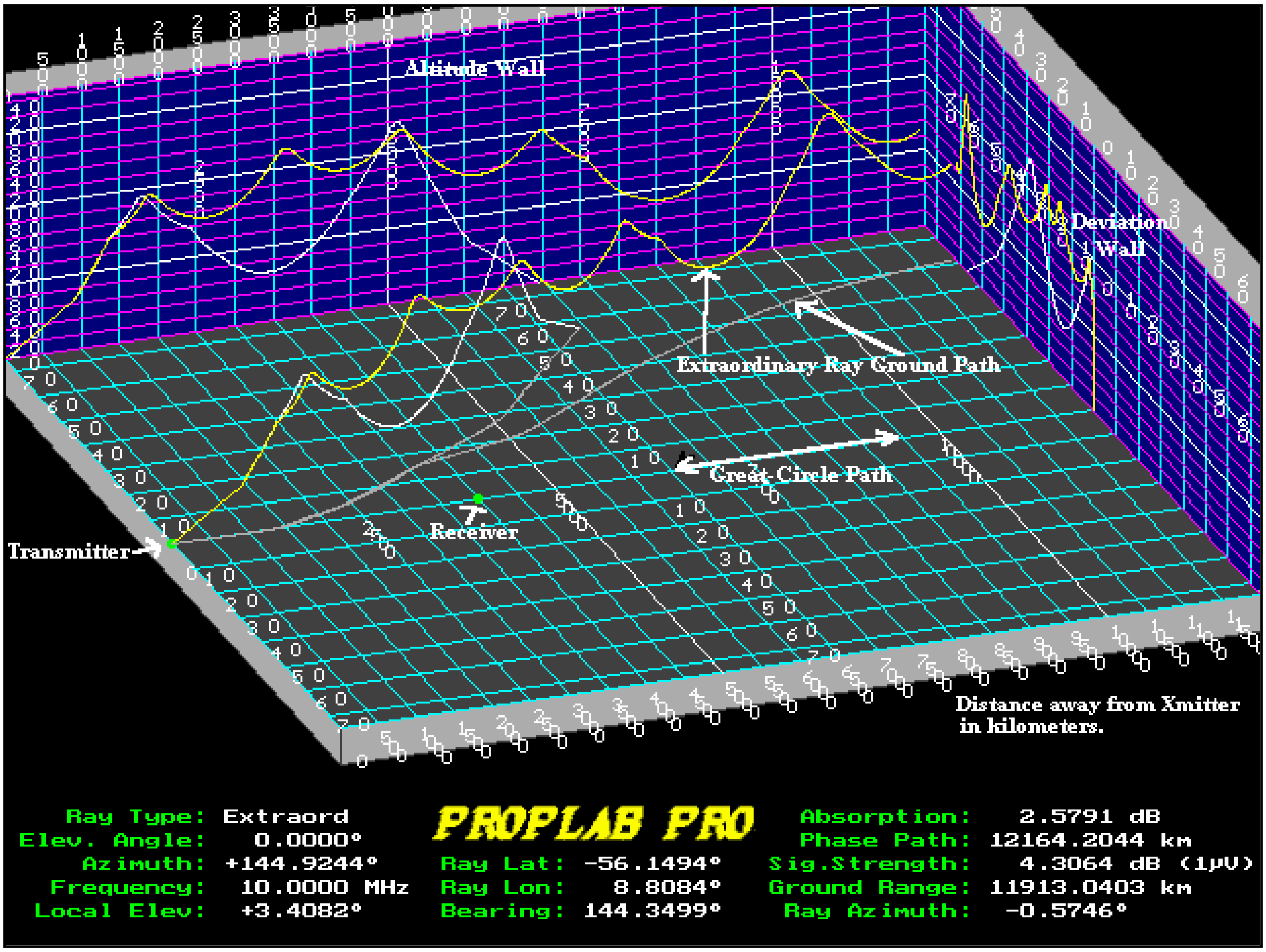


Fig 1-56 — 3-D ray tracing by *Proplab-Pro.* This software even addresses the subject of path skewing, although certainly not all the mechanisms causing this phenomenon are known well enough to be fully described in any present-day modeling program. (Courtesy STD.)

density; and auroral activity. The following geomagnetic maps can also be displayed: geomagnetic latitude; corrected geomagnetic latitude; magnetic dip; modified magnetic dip; magnetic dip latitude. The geomagnetic data input can be obtained automatically from the program *IonoProbe*, written by the same author (see Fig 1-1 and Section 3.2.7.1). You can download a trial version from **www.dxatlas.com**.

5.2.2. *Proplab-Pro*

Proplab-Pro is a professional ray-tracing program (See **Fig 1-56**). The author, Cary Oler of STD, calls it a "High-Frequency Propagation Laboratory." It is a full-fledged propagation-prediction program that also generates a range of maps. It is probably the most sophisticated and most advanced program available, but it is possibly too professional for the average ham (even for a dedicated lowbander) because of the very complex user interface. (Adding to the difficulty of use was the fact that the old version of *Proplab-Pro* was DOS-based — the new version is *Windows*-based, and runs on modern PCs.) But if you really want to study propagation in fine detail, I can highly recommend this program. It actually does three-dimensional ray tracing that includes the effect of the Earth's magnetic field, a rigorous calculation of absorption and it will predict and plot skewed paths. You can order and download the program from **www.spacew.com/proplab**. The program runs under *Windows* XP or Vista (see also Section 5.3.1).

5.2.3 *W6ELProp*

W6ELProp from W6EL (**www.qsl.net/w6elprop**) has a nice mapping program that shows the true short path, the true long path, the terminator, and the boundaries of the polar ionosphere. **Fig 1-57**. shows a typical *W6ELPr*op map (see also Section 5.3.1.4).

5.3. Propagation-Predicting Computer Programs

Earl, K6SE (SK), in a message on the Internet, wrote: "I gave up long ago on trying to predict DX conditions on 160 meters. One major observation I've made over the years is that on a night conditions are good to EU from here (northwest), conditions to JA (northeast) that same morning are not exceptionally good. And, if conditions are good to JA in the morning, in the evening conditions to EU are not good. The only 'prediction' I use now is: if conditions are exceptionally good in any particular direction on a given night (for example, to Europe on December 28, 1997 in the Stew Perry contest), I hope there will be a repeat one solar rotation later (27 days). Generally, I've come to the conclusion that 160 meter condi-

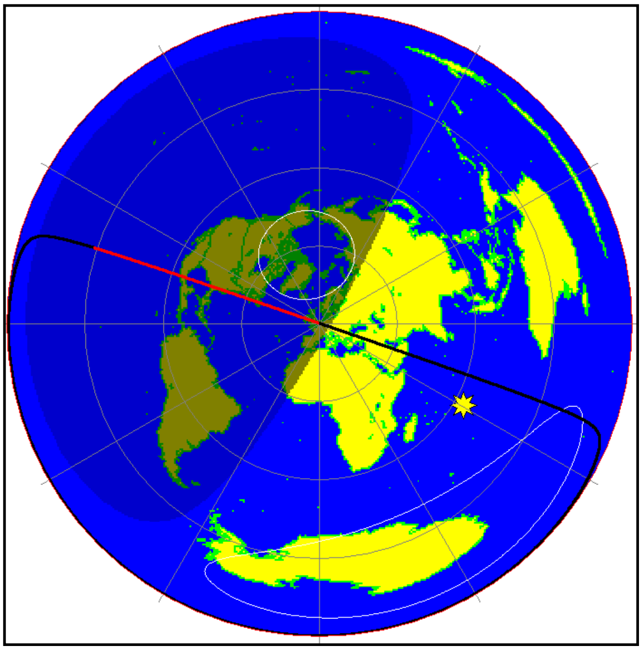


Fig 1-57 — Great circle map centered on Western Europe and showing the short and long path to VP6, daylight and darkness for mid December and the auroral zones (K=3).

tions are unpredictable, hi, so I just check the band every night to see what's happening."

Tom, N4KG, who says that he *lives* on the low bands, recently wrote: "I find very little correlation between solar flux and low band propagation, particularly on 160 and 80 meters. Each of the low bands has a distinct characteristic. Forty meters does follow solar flux in that the MUF can drop below 7 MHz when the flux levels are very low. During these times, 40 meters will close shortly after sunset on northern paths and there will be no European sunrise opening to the USA. With slightly higher solar activity, 40 meters will stay open all night. On 80 meters, solar flux is not so much of a factor. The *best* times are just before local sunset and after local sunrise. There is a significant enhancement at the terminator, which seems to be a daily event.

"There seems to be some validity to the theory that propagation is enhanced at the very start of a solar disturbance, but then the band goes flat for several days until the ionosphere stabilizes again. 160 meters does not correlate with *anything*! You may see nice enhancements at sunset and sunrise, or you may see nothing at sunset and sunrise but find good openings between 0200 and 0500 to Europe (from USA). On signals from the east, I have seen peaks at their sunrise and I have seen them peak a full hour before sunrise (before any daylight) and then vanish. Propagation prediction programs seem to be almost useless and often misguiding on the low bands because they fail to model the focusing effects at sunrise and sunset and they do *not* look at non-great circle paths. Most over the pole paths to the opposite side of the world come via skewed LP, at least here in eastern USA...."

I could not have said it better myself! This being said, current propagation prediction programs are useful for predicting 40 meter propagation, and some of them include very useful viewing and mapping facilities. For me, the final *acid test* that would make me a firm believer in propagation-forecasting programs is when such a program would successfully forecast the odd 80 and 160 meter openings *and paths* (directions), like the Europe to ZL path described in Section 4.3.1.2. I also would like to see a quantitative confirmation of bent paths that we so often experience. For anyone to accurately model the entire system that determines attenuation on 80 and 160 meters, we will have to know *all* possible mechanisms (and understand the day-to-day variability of the ionosphere — something we don't even understand on the HF bands). Today I have the impression that we see only the tip of the iceberg but the issue is still being addressed.

There may be light at the end of the tunnel. Rod Graves, VE7VV, himself an active low band DXer, has written a program that attempts some new approaches to predicting the more *odd* openings, the ones that other programs don't seem to recognize. The concept of the program is quite novel and employs a zone method developed by the author. It supports E-F region ducted paths and also directly addresses skewed paths. These are not skews due to magnetic disturbances, but skews due to the structure of the ionosphere at a given time. The program seems to predict very long-distance and long-path openings on 80 and 160 meters much better than any other. The limitation of the software is that it does not take into account auroral absorption. The software is still in a development phase, but may become available at a later date. VE7VV also admits that signals from locations near the antipode may travel paths other than the ones his program checks.

VE7VV comments: "Like most others, I do not use a program to tell me when the bands might be open to unknown locations. I just turn on the receiver to see what, if anything, is coming in from wherever and enjoy the discovery of the unexpected DX. However, when I want to make a sked with someone, or when there is a new DXpedition, I run the program to determine when the band might be open to that specific location and when the predicted optimum times are. I use this as my guide for when to be sure to be listening or for when to make the sked. Sometimes the program shows only what is common lore, or just provides another way of determining what could have been done with sunrise/sunset tables or gray line devices, but the program does it very conveniently for me. Sometimes, however, the program reveals times of openings, or times of peak signal strength, that are not obvious. This is especially true for paths over 10,000 km, particularly on the low bands (40 through 160 meters) where refraction and ducting phenomena are more prominent."

5.3.1. Propagation Software

An excellent in-depth review of many propagation related programs can be found on the Web site of ON4SKY (**www.astrosurf.com/luxorion/qsl-review-propagation-software.htm**). A short overview of the main programs follows. But first this: the programs based on MUF are only good from 3.5 to 30 MHz, for the good reason that the MUF has no influence on 160 meters (the MUF is always higher than 2 MHz).

5.3.1.1. Proplab-Pro

Proplab-Pro a professional and very comprehensive propagation analysis, propagation predicting and ray tracing program, can be used on all the low bands, including 160 meters. But it is more of an analysis program than a quick and easy to use prediction program. If you want to see what can create a duct, and visualize it, this is, as far as I know, the only program that will do that for you.

5.3.1.2. VOACAP

VOACAP is a free HF Ionospheric Communications Propagation Analysis and Prediction program from NTIA/ITS, originally developed for Voice of America (VOA). For a brief introduction go to **lipas.uwasa.fi/~jpe/voacap.** *VOACAP* by itself does not include input for planetary indices (A and K indices). This results in inaccuracies in disturbed conditions, especially at high latitudes as well as for Top Band predictions where geomagnetic and gyrofrequency effects are not taken into account. VOACAP is quite complex, and does not provide a really user-friendly interface for the more a casual user. *Ham CAP* (see below) is a shell written around *VOACAP* that makes it a very user-friendly application. Additionally, *VOACAP* was intended to only go down to 2.0 MHz, so its usefulness for 160 meter work is questionable. Yes, 1.8 MHz is close to 2 MHz, but validation data is likely to be lacking.

5.3.1.3. Ham CAP

If you look for a quick, easy (but not dirty!) and accurate propagation predicting program that works extremely well all the way down to 40 and even 80 meters, I can highly recommend *Ham CAP* (**www.dxatlas.com**) by VE3NEA, which is based on the *VOACAP* engine. *Ham CAP* (**Fig 1-58**) employs an empirical algorithm that takes the K_p value into account. This option can be toggled using the check box. K_p data from the Web are also available via the button. This makes for the program to be usable down to 80 meters. If you have *IonoProbe* (see Section 2.1) installed, the program will automatically tech the sunspot and K_p data (which you can manually override). If *DX Atlas* is installed on your computer, you can also see all the results on a much more detailed screen-sized map (**Fig 1-59**).

5.3.1.4. W6ELProp

This program, written by Sheldon Shallon, W6EL, is another user-friendly program that has some excellent mapping

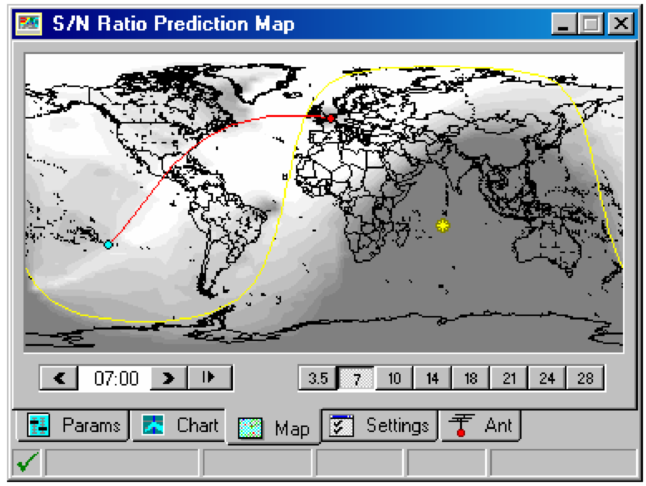


Fig 1-59 — You can also switch to a Mercator map showing the great circle direction between the two stations, the terminator. The different shades of gray represent the varying S/N ratios of the transmitted signal produced by the transmitted signal. Dark gray means negative S/N (no signal), while white indicated a very high S/N ratio. This example shows the 7 MHz path between Belgium and VP6 in mid February 2008 around 0700 UTC.

possibilities. (*W6ELProp* is a *Windows* version of *MiniProp Plus*.) It predicts ionospheric (sky-wave) propagation between any two locations on the Earth on frequencies between 3 and 30 MHz. The latest *Windows* version of the program is freeware and can be downloaded from **www.qsl.net/w6elprop**.The program also takes into account geomagnetic data (K-Index). See also Fig 1-57.

5.3.1.5. Conclusion

We should really not call these programs *propagation prediction* programs, a denomination which makes me think of black magic, Madame Soleil or Voodoo witchcraft.

These programs are propagation *modeling* programs. They make a model of the propagation on a given frequency between two points, based upon the data you (and the program, using statistical data) provide. In other words: based upon what we know (the mechanism, physics, mathematics), what we measure (SSN, K_p) and what we have experienced (the statistical data) these programs show you a *model* of what is to be expected.

To be perfectly honest, in my almost 50 years of DXing on 80, and in my 22 years on 160, I have almost never (successfully) used a propagation-prediction program on 160 and 80 meters for difficult-to-predict paths. None of the programs have told me why I once worked FK8CP, and why out of hundreds of days of attempts, only one day could I really hear his signals. None of these computer programs will tell you why it was possible just that one day. But then again, that's one of the charms of Top Band!

Now and then I do use a program such as VE3NEA's *Ham CAP* or *W6ELProp* to help me look for the best operating times on the higher bands (40, or even 80 meters, through 10 meters). On these bands propagation modeling programs perform very well, simply because the number of variables involved are limited and well known.

For the real low bands (80 and 160 meters), all you need

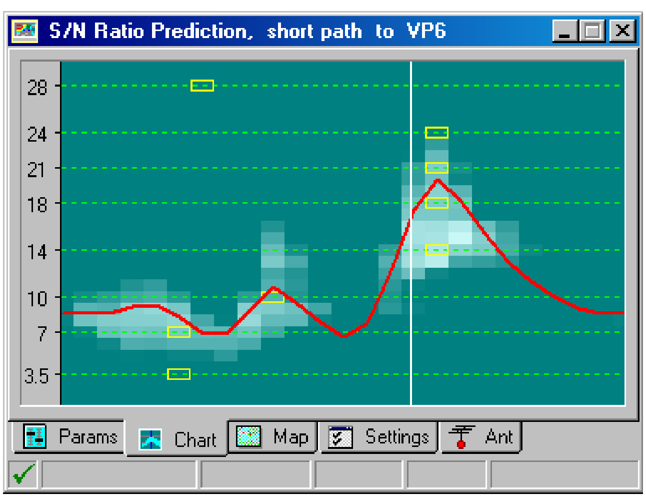


Fig 1-58 — *Ham CAP* is the author's favorite quick and easy propagation modeling program. It is very user-friendly and accurate, uses the powerful *VOACAP* engine but also introduces corrections for the planetary K-Index. This screen shows the MUF (February 2000) for a path between Belgium and VP6 (Ducie).

is a good mapping program (eg *DX Atlas*), good sunrise/sunset data and current WWV information about A and K indices. All you need in addition is the perseverance to be there every day, a good station and some luck which is exactly what makes 160 meters an adventure!

On 160 meters there is one huge problem when it comes to making 100% accurate models: we know only a limited number of all the variables and all the mechanisms that govern propagation on Top Band. In many cases we seem to be able to explain the unusual or odd behavior of the band *after the fact*. If predicting propagation were as simple as it is on the higher bands, would we be interested in Top Band?

Bottom line: I never use such programs for "telling me the future." I like to be *amazed* when the almost impossible happens. Amazement is what happened to me when I first saw Amateur Radio at an age of about 8. It made me become a ham and an engineer.

5.4. ON4UN Propagation Programs

When I was writing the original *Low-Band DXing* book, now more than 20 years ago, I developed a number of computer programs as tools for the active DXer. These MS-DOS programs are shareware and available on the CD that comes with this book.

The software runs perfectly under *Windows* XP and earlier versions in a DOS box. While the majority of my programs are technical programs related to antenna design (and are covered in the antenna chapters of this book), a group of programs deals with the propagation aspects of low-band DXing. The Propagation Software contains the following modules:

- Calculating sunrise/sunset times
- Listing sunrise/sunset times (based on a database of countries that you can maintain)
- Gray line program

This gray line program is quite unique in that it uses a specially developed algorithm that adapts the effective radio width of the gray line zone to the location and the time of the year. The width is also different for 160 and 80 meters. The gray line zone as calculated by the program is not the twilight zone (see Section 2.4.4.), which we see on most of the maps (eg *DX Atlas*). In addition, the user can specify a minimum distance under which he is not interested in gray line information. The printout (on screen or paper) lists the distance to the target QTH, the beginning and ending times of the gray line window, as well as the effective width of the gray line at the target QTH. **Fig 1-60** shows a screen dump of a gray line run for Belgium on February 27. Notice the short Chatham Island opening predicted between 0614 and 0623 UTC. I made the QSO on 160 meters at 0623, and copied ZL7DK until 0634 UTC.

```
                -W6-CAL-
USA                (SAN FRANCISCO)
==================================
37.78 DEG.N.        122.41 DEG.W.

DATE           SUNRISE     SUNSET
----           -------     ------
JAN 1          15.25       01.01
JAN 16         15.24       01.15
FEB 1          15.14       01.33
FEB 16         14.58       01.49
MAR 1          14.41       02.03
MAR 16         14.19       02.17
APR 1          13.55       02.32
APR 16         13.33       02.46
MAY 1          13.14       03.00
MAY 16         12.59       03.13
JUN 1          12.49       03.26
JUN 16         12.47       03.33
JUL 1          12.51       03.36
JUL 16         13.00       03.31
AUG 1          13.13       03.19
AUG 16         13.25       03.03
SEP 1          13.39       02.40
SEP 16         13.52       02.40
OCT 1          14.05       01.54
OCT 16          14.19      01.32
NOV 1          14.35       01.12
NOV 16         14.51       00.58
DEC 1          15.06       00.51
DEC 16         15.19       00.52
```

Fig 1-60 — Example printout for one of the more than 500 locations in ON4UN's Sunrise/Sunset tables. Times are given for half-month increments.

6. DIFFFERENCES BETWEEN THE 40, 80 AND 160 METER BANDS

6.1. 40 Meters

- 40 meters is like an HF band that works at nighttime (it's almost like VHF for a Top Bander!).
- Propagation prediction can be done by classic MUF-based programs.
- Gray line propagation also happens *along* the terminator, as on the higher HF bands.
- Gray line zones can be very wide (many hours even at medium latitudes).
- Skewed paths are not as common as on 80 and more so 160 meters.
- 40 meters allows you to work any distance, if properly equipped.

6.2. 80 Meters

- With well-equipped stations at both ends, any distance and path can be covered at the right time of the year.
- During low sunspot years, 80 meter propagation may be influenced by MUF.
- Gray line enhancement always occurs on paths perpendicular to the terminator.
- Working DX through the auroral belt is not uncommon, even in high sunspot years.

6.3. 160 Meters

- Propagation is not at all dictated by MUF, and only marginally by the solar cycle.
- Besides the auroral phenomenon we still do *not* know what makes a good DX night or a bad one. Mystery is still a big part of Top Band!
- Auroral absorption is most pronounced on 160 meters.
- Skewed paths most frequently occur on Top Band.
- Gray line enhancement appears to occur most frequently on paths perpendicular to the terminator.

- 160 meters has a distinct area in which working DX is more or less like a piece of cake — anything in a circle of approximately 5000 km around your own QTH. For instance, Western Europeans can work the East Coast of the USA almost daily. The "light-gray" zone is W5, W8 and W9 land. WØ land is "dark gray" and for anything beyond that, conditions must be well above average. This is quite different from 80 meters, where longer distances are possible every day and where the transition between "easy" and "difficult" seems to be much more vague.
- Real long path (that is, without path bending) on 160 meters is rather exceptional, except for stations very near the antipodes, and as a rule only occurs during low sunspot cycle years.
- If 80 meters is swinging, there is no guarantee that 160 meters will be any good. When 80 is bad, though, 160 will likely be bad as well. So don't extrapolate from the higher band to the lower band. This very often does not work.
- Very typical for 160 meters is a slow and deep QSB, especially on marginal paths. It's advisable to get a call right the first time. There may not be a second time or it may be minutes later. I have seldom seen this on 80 meters. Patience is important also. You may have to wait for propagation to peak to you.
- During low sunspot cycle years, 160 meters usually has very pronounced peaks at sunrise (sometimes also at half-way midnight), especially for really long-haul paths, and when both ends of the path are located on the terminator. Very often we notice quite a sharp peak in signals within minutes of sunrise. You can almost set your watch by it. The sunset peak on 160 meters is also much less pronounced than the sunrise peak. There seems to be a broad "peak" within an hour or so after sunset.
- On 160 meters skip is often very area selective (for various reasons).
- Working DX through the auroral donut is very difficult (impossible) on 160 meters.
- The thrill of working a new one on 160 meters is ten times the thrill of doing it on 80 meters!

7. THE 160 METER MYSTERY

Understanding and predicting propagation on 40 meters is pretty straightforward and 80 meters is well understood as well. With the right equipment and knowledge on both ends, you could probably work 300 countries in a few years on 80 meters.

The 160 meter band is a totally different ball game. The more I have been active on 160 meters, the more I am convinced of how little we know about propagation on that band. True, we know a few of the parameters that influence propagation, but far from all. For a long time I have kept daily records of the K and A indices, sunspot numbers, etc, together with my own observations of conditions on 160, to find a correlation between the data and actual propagation. But I have found very little or none; only negative correlations. We know more or less when it definitely *will not* work, but not for sure when it *will* work.

Of course, we must realize that on Top Band we are in a gray area, where things are sometimes possible but often not. There are dozens of parameters that make things happen, or not happen. They all seem to influence a delicate mechanism that makes really long-haul propagation on 160 meters work every now and then. Understanding all of the parameters and being able to quantify them and feed them into a computer that will tell exactly when we can work that elusive DX station halfway around the globe will probably be an illusion forever.

Top experts in the field of radio propagation like Ted Cohen (N4XX) and Cary Oler (of Solar Terrestrial Dispatch), wrote: "Top Band is one of the last frontiers for radio propagation enthusiasts. It involves regions of the Earth's environment that are very difficult to explore and are poorly understood. These factors have led to our failure to predict propagation conditions with any level of accuracy. They also account for our inability to explain some of the puzzling mixtures of conditions that make this one of the most interesting and volatile bands available to the Amateur service. Top Band may be the lowest band in the amateur spectrum, but it has one of the most promising and exciting futures possible!" I could not agree more.

There is no interest from the broadcasters in understanding and especially predicting long haul propagation on 160 meters. Broadcasters and utility traffic operators are interested in knowing the operating frequency that will give them best propagation. They are not interested in studying the subject of "marginal propagation," just on the edge of what is possible. Therefore, long-haul DXing on 160 meters will probably always remain a real hunting game, where limited understanding, feeling, expertise and luck will be determining factors for success. Don't forget your hunting weapons — your antennas and your equipment.

Top Band is the band where these weapons are your most important assets. A good location (eg, at the edge of saltwater), decent transmit antennas (low angle, with a few dB gain if possible) and some power easily add up to 10 dB or even 20 dB or more over a "mediocre" setup and this is a lot. Good receiving antennas can make up to 10 dB better signal to noise ratio. Ten dB may not be much if the signals are S9. If, using the mediocre setup, the signals are 5 dB below the noise level, then an extra 10 dB makes the difference between a contact and … just noise. In other words, there is little we can do about propagation, but a lot we can do about our station. That's why 90% of this book covers what we have in hand: our antennas, our station and our operating techniques.

I am writing this 5th edition of this book at the bottom of my third sunspot cycle on 160 meters (1987, 1997 and 2008). Have these dips in the cycle been similar? Not in my opinion. I always thought 1987 was the best one, but that is likely to be a subjective statement as in Belgium we were only allowed to operate 160 meters starting in 1987. Everything was new in those days. I remember working W6RR my first day on the air and KH6AT in my first month on 160 meters.

Today I would say that the 2008-2009 winter was the best season ever. Early as September, it was possible to work the US West coast almost every day, and contacts with KH6 (again KH6AT and others in Hawaii) and KL7 happened almost every week! What makes the winter 2008-2009 season such a good Top Band season? It is true that the sun remained in a prolonged winter sleep this time. Fact is that we, year after year use better equipment and better antennas, and that more and more hams have found their way to 160 meters and found it to be the place for unequalled challenge and excitement. Maybe some of the factors influencing propagation over the very long paths are linked to the physics that more and more influence weather globally.

I find it very remarkable that over the past few seasons we have seen more and more pronounced QSB on these long haul signals. QSB takes signals *up* as well as down, and makes available an extra layer of signals that, in my opinion, were previously unavailable for a contact.

What made the 2008-2009 winter exceptionally good was the total absence of any aurora activity for weeks on end, which made it possible to work across the poles with a certain consistency. That has been very rare. On the negative side, it appears that the deep sleeping sun often left the ionosphere with too little ionization (gradients) to set up solid and reliable signal ducts which are required for really long haul propagation (greater than approximately 13,000 km).

Contents

CHAPTER 2

DXing on the Low Bands

I asked Bill Tippett, W4ZV, to be my critic, guide, counselor and support for this chapter because he provided such good help with the Fourth Edition and was lightning fast as well. Bill needs no introduction to the active low band DXers, but let me just introduce W4ZV to the newcomers to this playground of our hobby. Bill is one of the very few hams in the world with more than 300 countries confirmed on 160 meters, and he was the first to reach that number, in 1998. As of mid-2010, Bill is leading the 160 meter DXCC count with 327 countries confirmed! That says enough. You don't achieve this unless you have a profound know-how. And Bill was willing to use his know-how and expertise to help me with this chapter. It is indeed an honor for me to have Bill review this chapter and help me with his suggestions.

Bill Tippett, W4ZV, shown here operating VY2ZM, was the first ham to work 300 countries on 160 meters. He is also moderator of the Top Band reflector on the Internet.

Bill was first licensed at age 12 in 1957 as KN4RID. The DX bug bit hard as soon as he made his first DX contact on 15 meters, and he went on to become the first US Novice class licensee to achieve DXCC. Bill made the DXCC Honor Roll seven years later in 1964 while he was still a teenager. School, work and marriage mostly curtailed his operating until he got the bug again in 1976 while he was working in Colorado when he became WØZV. In 1980, he moved to 40 acres in the country and began seriously chasing new ones on 80 meters. In October 1984, he put up a "temporary" 160 antenna for a few multipliers in the CQ WW SSB contest and became addicted to the band. Bill said, "When I first got on 160, some of the locals said I would do well to make WAC (Worked All Continents)." By April 1985, he was issued the first 160 meter DXCC for the WØ area, and was thoroughly addicted to Top Band. In 1993, Bill moved back to his home state of North Carolina and received the call W4ZV in 1996.

Thank you, Bill!

In 2006 my friend Mark, ON4WW wrote the document *Operating Practice* which was an overwhelming success, judging by the large number of positive reactions he received. The document is available in nearly 20 languages on his Web site (**www.on4ww.be/op.html**). Two years later Mark and I decided to expand this excellent document to cover nearly all aspects concerning ethics and operating procedures on the bands. I would like to thank Mark, ON4WW for having allowed me to do so, and for all his help.

Mark Demeuleneere, ON4WW

In June 2008 this new document *Ethics and Operating Procedures for the Radio Amateur* was accepted by the Administrative Council of the IARU as representing its point of view concerning the subject. A number of items from this new publication are included in this chapter. At this time there are translations of this document in more than 25 languages, and more are in preparation. These are available from various IARU member society Web sites. All the different languages are also available from one single Web site (**www.ham-operating-ethics.org**). In some countries the document has been made available as booklet.

Tree, N6TR, considers 160-meter DXing as a disease. He described the symptoms:

- Desire to be on the radio at sunrise.
- Desire to be on the radio at sunset.
- Desire to be on the radio at all times in between sunset and sunrise.
- Desire to struggle for months to work a single station in a new country.
- Never being satisfied with the antenna system and constantly trying new ones.
- Only comes down to see the family after working a new country (to gloat). During the rare fantastic opening, will come down after each new country and hold up fingers indicating how many new countries were worked so far. These events are rare and occur about once or twice in a century.
- Drinks lots of water before going to bed with the sole purpose of waking up in the wee hours of the morning to see if a new country can be found.
- Has problems getting to work on time during the winter months.
- Sends equipment and wire to people in countries not yet worked, hoping that the end result will be their QSL card on the wall, or rather in the safe!
- Spends thousand of dollars going to rare countries just so other people can work it.

And these are only some of the better-known symptoms. According to the late Rush Drake, W7RM, it's a painful disease: "To work DX on 160 you've got to love pain." The late Earl Cunningham, K6SE, changed that to: "You've got to love torture..." Who am I to disagree with the wise words of such eminent low-band DXers?

Our 160-meter band is usually referred to as Top Band, the band at the top of the wavelength spectrum, the band with top-notch operators, the band that's a top challenge and that gives you top excitement and satisfaction. Gary, NI6T, says: "One sixty? Not a band, but an obsession."

All kidding aside, low-band DXing is a highly competitive technical hobby. It is certainly not a communications sport for the appliance operator. It is one area of Amateur Radio where it really helps to be knowledgeable. This is not a "plug and play" hobby!

Fig 2-1 — Stew Perry, W1BB (SK), Mister 160 Meters, in his shack late in his career. Stew was the pioneer of 160 meter DXing, and the Stew Perry Top Band Distance Challenge, one of the most respected Top Band contests, was named after him.

1. MYTHS

Gerry, VE6LB, who is a successful low-band DXer from an urban QTH, from the middle of nowhere, right in the auroral doughnut, using simple antennas, summed up a few myths:

- There is no (or little) DX on the low bands!
- You need a big antenna and high power (it's only for the big guns) to work DX on the low bands!
- DX is so scarce that you need to spend many hours (mostly late at night) to find DX on the low bands!
- Any DX to be found on the low bands is on CW.
- There is no low-band DX during the summer.
- The low bands are too noisy to work DX.

2. REALITY

Let's look at some facts:

1) Over the years all DXCC entities have been available on 40 and 80 meters, with the exception of P5 (North Korea) on 80 meters. As of May 2010, more than 3000 DXCC certificates have been issued for 80 meters, and the 2009 DXCC Annual List included 109 stations with more than 300 countries confirmed on that band. On Top Band only two countries have, so far, never been available (as of May 2010): 7O (Yemen) and P5. As of May 2010, nearly 1900 stations have obtained 160-meter DXCC, and the 2009 DXCC Annual List showed 16 stations with more than 300 DXCC countries confirmed.

2) You will probably never win the CQ World Wide 160-Meter Contest from a suburban lot with a 50-foot antenna-height restriction. But you can work DXCC on the low bands, even with 100 W from a typical suburban lot. KH6DX/W6 worked over 100 countries on 160 meters from his mobile station! I have friends who have never run "power" (more than 100 W) and have over 100 countries on Top Band. It is true — of course — that the better the means, the more you'll be sitting in the front row when the show is on.

3) It is true that most of the DX on the low bands can be worked around sunset or sunrise, but that is a better arrangement than on 10 meters, where the DX shows up in the middle of the day when most of us are at work.

4) Too bad not all the low-band DX is on CW. (That's a personal note. I love CW so much more than phone!) Seriously, there are countries that are only available on phone and others only on CW. That's the name of the game. When it comes to Top Band though, CW is the name of the game! It's Top Band, and CW, that separate the players. That's why on Top Band more than 80% of the players use CW at least 90% of the time (see Section 20).

5) Ever considered that when it's summer here, it's winter on the other side of the equator? That means you can work good DX even on 160 meters during your local summertime.

6) Noise, whatever its origin, is one of the main challenges for the low-band DXer, but it certainly does not stop real hams from DXing. Even on VHF, noise is the limiting factor, be it a noise of a

Fig 2-2 — DX hall of fame member Jeff Briggs, K1ZM, also author of the ARRL published book *DXing on the Edge* (now out of print) built a 160-meter dream station (VY2ZM) on Prince Edward Island. He also was one of the first North American stations to have worked over 300 countries on Top Band.

Fig 2-3 — Jack Leahy, VE1ZZ, always first in and last out when it comes to Top Band openings into Europe. More recently this reputation is being challenged by VY2ZM (K1ZM's new station on Prince Edward Island)!

> very different origin. This is not a broadcast hobby, nor is it a communicator's hobby. In this low-band hobby we are driven to move the boundaries of what is possible. It's the challenge that attracts us.

Well, all of this does *not* mean that working DX on the low bands is just a piece of cake, a nice pastime for the appliance-type operator. But what makes so many love the low bands for chasing DX?

3. WHAT MAKES US CHASE DX ON THE LOW BANDS?

I included this question in every questionnaire that I sent out via the Internet in preparation of a new edition of this book. The answers always have remained the same. It is the *challenge*, the sense of fulfillment and of having done something difficult. (See Section 20.8.)

Low-band DXers are always near the edge of what is possible. The most successful low-band DXers are the pioneers who keep moving this edge. Improved understanding of propagation, together with better equipment, and most of all, better antennas, make it possible to dig deeper and deeper into the noise to catch the previously evasive layer of buried signals. Top Band DXers are those balancing themselves on the cutting edge of the DXing sword!

If you are looking for an easy pastime or if you just want to chat via ham radio, stay away from low-band DXing. Jeff, K1ZM, who now is one of the few US stations with over 300 countries on 160 meters, wrote on his survey reply: "160 is truly a *man's* as well as a *gentleman's* band. You want a challenge? Get on 160."

On 80 and 40 meters you do not need to have a genuine antenna farm to work DXCC, even within one year's time. Even on 160 meters, urban QTHs with small and low antennas regularly produce DXCC on Top Band. I have included in this book a short chapter on "Low Band DXing from a Small Garden" that shows what can be done from a small lot. There are many examples of rather modest stations on a small suburban lot that have done extremely well. My friend George, K2UO, worked over 250 countries on 160 from a ½-acre suburban lot. To be so successful from an average QTH requires a better-than-average knowledge of propagation, as well as a substantial dose of perseverance.

How many are we? I have analyzed the statistics of a number of major DXpeditions in the last years, expeditions that worked all the bands (and not just a short appearance on 160 the last day…). From these data we learn that the Top Band DXers going after a new country on 160 meters can number between 4000 and 5000. The 80-meter DXers seem to add up to approximately 10,000 and those of us who chase a new country on 40 meters are between 15,000 and 20,000. So, it is not just a few who like the challenge of the low bands, it is one big family.

4. THE FREQUENCIES

The frequencies used for DXing on the low bands are not the same in all countries. Therefore it is important that you know where to look for the DX. There are four levels we should look at:

1) What are the *International Telecommunication Union (ITU) allocations* in the three ITU regions?
2) What are *country allocations*? Each individual country can, within the ITU allocations, reserve specific frequency ranges. Are there mode-related subbands that are enforced?
3) Since such mode-related subbands do not exist in most countries, what is the *band plan* that radio amateurs have agreed upon in an international context? In others words, what does the International Amateur Radio Union (IARU) band plan say?
4) What is the *common-sense band planning* that low-band DXers apply, in case the IARU band planning does not meet our goals?

4.1. The ITU Allocations

The ITU has divided the world into three regions:

Region 1: Africa, Europe, former USSR countries, Middle East (excluding Iran) and Mongolia

Region 2: North and South America including Hawaii, Johnston and Midway Island

Region 3: The rest of Asia and Oceania

The allocations are described in the *Radio Regulations* published by the ITU. Article S5 describes the allocations in detail. The RR publication can be bought from **www.itu.int/publ/R-REG-RR/en**. Watch out: The ITU allocations are not the same as the IARU Band Plan!

4.1.1. 160 Meters

Region 1:

1800 to 1810 kHz: No Amateur Radio allocation (used for Radiolocation)

1810 to 1850 kHz: Primary allocation for Amateur Radio, however …

1810 to 1830 kHz: In 31 countries these frequencies are *not* available to Amateur Radio, at least this is what footnote 5.98 of ITU-RR5 says. Reality seems to be different in many of those countries however. Footnote 5.99 lists another 13 countries in which 1810 to 1830 kHz are open to other services (Fixed and Mobile) as primary users, in which case Amateur Radio is secondary. But, with the exception of the problems reported in Section 3.2.1, very few stations other than amateur stations are ever heard in this portion of the band in Europe.

1850 to 2000 kHz: Primary allocation to Fixed and Mobile services. The number of countries where this section, or the lower part of this section, is allocated on a secondary basis with a power limit of 10 W average has grown year after year.

Region 2:

1800 to 1850 kHz: Allocated exclusively to Amateur Radio

1850 to 2000 kHz: Shared between Amateur, Fixed, Mobile, Radiolocation and Radionavigation. In most South American countries this section is allocated to Fixed and Mobile services on a primary basis (which means Amateur Radio is secondary).

Region 3:

1800 to 2000 kHz: Shared between Amateur, Fixed, Mobile, Radionavigation and Radiolocation.

4.1.2. 80 Meters

Region 1:

3500 to 3800 kHz: Shared between Amateur, Fixed and Mobile services.

Region 2:

3500 to 3750 kHz: Exclusively Amateur Radio

3750 to 4000 kHz: Shared between Amateur Radio, Fixed and Mobile services. In LU, CP, CE, HC, ZP, OA and CX the Amateur Radio allocation is secondary. In VE and OX 3950-5400 can be used for Broadcasting as a primary service.

Region 3:

3500 to 3900 kHz: Shared between Amateur Radio, Fixed and Mobile services

3900 to 3950 kHz: Aeronautical Mobile and Broadcasting

3900 to 4000 kHz: Fixed services and Broadcasting

4.1.3. 40 Meters

Region 1:

One of the major achievements at WRC 2003 (the ITU World Radio Conference in Geneva) was the fact that a portion of the 40 meter band (7100 to 7200 kHz) that had previously been taken away from Amateur Radio was returned to the Amateur Radio service after March 29, 2009.

7000 to 7200 kHz: Exclusive Amateur (in some countries the Fixed service has a primary status and Amateur Radio secondary)

Region 2:

7000 to 7300 kHz: Exclusive Amateur Radio

Region 3:

7000 to 7200 kHz: Exclusive Amateur Radio (in some countries Fixed service can be used as primary service)

4.2. The IARU Band Plan

The IARU (International Amateur Radio Union) includes one national radio society from each member country (the most representative one) and sets out as one of its goals to establish and maintain a band plan that has been approved by all of the IARU radio societies. The IARU is organized in the same three regions as the ITU (see Section 4.1).

The band plan listed for Region 1 (R1) was adopted during the IARU R1 plenary meeting in November 2008. Since 2005 the band plan is based on *bandwidths* of transmitted signals, rather than on *modes*, which, for many of us, make this often difficult to understand and, in addition, open for various interpretations. The Region 2 band plan is based on bandwidths and also lists preferred modes. The data listed for Region 2 became effective January 1, 2008. Region 3 seems to stick to the good old mode parameter.

Table 2-1 (source: IARU Web page) gives an overview of the band plan regarding 160, 80 and 40 meters in the three

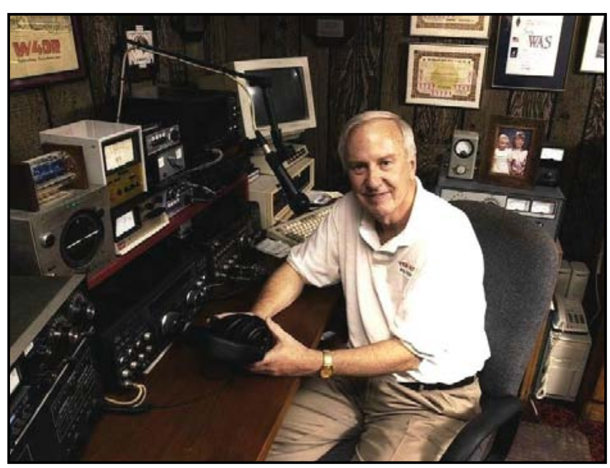


Fig 2-4 — Bob Eshleman, W4DR, DX Hall of Fame member, has 319 DXCC countries confirmed on 160 meters, 352 on 80 meters and 361 on 40 meters (as of June 2010).

Table 2-1
IARU Band Plans by Region

160 Meters

Region 1		*Region 2*		*Region 3*	
1810 – 1838	CW	1800 – 1810	Digimodes	1800 – 1830	CW
1838 – 1840	Narrowband modes	1810 – 1830	CW	1830 – 1834	RTTY, CW, DX
1840 – 1843	All modes, digimodes	1830 – 1840	CW, priority for intercontinental DX	1834 – 1840	CW
1843 – 2000	All modes	1840 – 1850	SSB, priority for intercontinental DX	1840 – 2000	Phone, CW
		1850 – 1999	All modes		
		1999 – 2000	Beacons		

80 Meters

Region 1		*Region 2*		*Region 3*	
3500 – 3510	Intercontinental DX, CW	3500 – 3510	Intercontinental DX, CW	3500 – 3510	DX, CW
3500 – 3560	CW, contest preferred	3510 – 3580	CW, contest preferred	3510 – 3535	CW
3560 – 3580	CW	3560 – 3580	CW	3535 – 3775	Phone, CW
3580 – 3590	Narrowband modes	3580 – 3590	Narrowband modes	3775 – 3800	DX Phone, CW
3590 – 3600	Narrowband modes incl unattended stations	3590 – 3600	Narrowband modes incl unattended stations	3800 – 3900	Phone, CW
3600 – 3620	Phone, CW, digimodes incl unattended stations	3600 – 3620	Phone, CW, digimodes incl unattended stations		
3600 – 3650	All modes, SSB contest preferred	3650 – 3700	All modes		
3650 – 3775	All modes, SSB contest preferred	3700 – 3775	All modes, SSB contest preferred		
3775 – 3800	All modes, priority for intercontinental DX	3775 – 3800	SSB (DX phone window), CW		
		3800 – 4000	All modes		

40 Meters

Region 1		*Region 2*		*Region 3*	
7000 – 7025	CW, contest preferred	7000 – 7025	CW, priority for intercontinental DX	7000 – 7025	CW
7025 – 7040	CW	7025 – 7035	CW	7025 – 7030	Narrowband, CW
7040 – 7047	Narrowband digimodes	7035 – 7038	Narrowband modes, digimodes	7030 – 7040	Narrowband, Phone, CW
7050 – 7053	All modes, wideband digimodes, unattended stations	7038 – 7040	Narrowband modes, digimodes incl unattended stations	7040 – 7100	Phone, CW
7053 – 7060	All modes, wideband digimodes	7040 – 7043	All modes, all digimodes incl unattended stations preferred	7100 – 7300	Phone, CW (Allocated on a secondary basis in Australia and New Zealand)
7060 – 7100	All modes, SSB contest preferred	7043 – 7100	All modes		
7100 – 7130	All modes, SSB	7100 – 7300	All modes		
7130 – 7175	All modes, SSB contest preferred				
7175 – 7200	All modes, SSB, priority for intercontinental DX				

Remarks:

The limiting frequencies indicated in the band plan are limits of the considered spectrum, which in many cases are not the same as the frequencies shown on the dial of your transmitter. Example: On 160 meters, the lowest acceptable dial setting for SSB (LSB) is 1843 kHz (the lower sideband will cover all the way down to 1840… and a little more).

This table does not mean that all countries in a given region are permitting operation in all of the segments mentioned in the table!

regions. The table shows the new 40-meter band plan that applies since March 30, 2009. Details about emergency centers of activity and QRP frequencies are left out. For complete band plan see **www.iaru.org/bandplans.html.**

4.3. In Practice

4.3.1. Where to Operate on 160 Meters

In view of the fact that most DXing on 160 meters is done on CW, a window reserved exclusively for CW is a must. So far the IARU band plans have such an exclusive CW subband. The same is not true for phone however. This probably stems from the historic days of Amateur Radio, but I cannot see any reason why the phone bands should not be as exclusive as the CW subbands!

New to the IARU 160-meter band plan is the inclusion of digital mode ("digimode") windows. In creating these new windows, consideration should be given for everyone already on the band before a new exclusive area is created, and this should be done through planning with existing CW and phone operators (rather than dictatorship) or the result will be a real mess and many hard feelings will be created. Before isolating

Fig 2-5 — Bob March, N7UA, the low band beacon from the Pacific Northwest, both on 80 and 160 meters. Together with Tree, N6TR, he is one of the low band beacons from that area.

several kHz from a prime area, the rule makers should make sure the operators already there (for many years) will have a suitable place to move and will be found willing to do so. As Tom, W8JI, wrote on the Top Band Reflector: "We need a long-term plan that does not displace primary users. The IARU needs to seek input from 160 operators before they mess things up for everyone, and cause a lot of hard feelings that last for many years." It is obvious that this message was *not* heard when in 2005 the band plans were changed. A band plan that is not accepted by the majority of the band users is a bad band plan, and should be changed.

Let's look at the ARRL-published 160-meter band plan, as downloaded from the ARRL Web site in May 2010 (**www.arrl.org/band-plan-1**). This is different from the IARU Region 2 band plan that came into effect on January 1, 2008! It is true that the IARU band plan lines up pretty well with the Region 1 band plan, which is positive. On the negative side we notice that, whereas in the Fourth Edition of this book I suggested not to squeeze the digital modes between CW and SSB, it appears that this was a loud and lonesome call in the desert. Note that this unfortunate decision in reality took away 2 kHz (1838-1840 kHz) of what is the DX window in Region 1.

ARRL Band Plan, 160 Meters

The data below comes from the ARRL Web site and does not line up with the IARU R2 band plan.

1800 to 2000 kHz:	CW
1800 to 1810 kHz:	Digital Modes
1810 kHz:	CW QRP
1843 to 2000 kHz:	SSB, SSTV and other wideband modes
1910 kHz:	SSB QRP
1995 to 2000 kHz:	Experimental
1999 to 2000 kHz:	Beacons

Five years ago I made a appeal to all IARU "rule makers" to think about the mode distinction existing between the CW band and the phone band, which is not realistic during major contest weekends (CQ 160 contests, CQ WW, ARRL 160, Stew Perry etc). It does not make sense to have a rule that nobody follows. Rules that a large majority of active hams ignore are bad rules, and must be changed.

Take for example Europe, where today only the frequencies below 1850 can be used with power higher than 10 W average (Remark S96 of the ITU frequency-allocation table), and that in whatever mode. It is clear that under such conditions it is not realistic to tell the contest operators they can only operate in SSB between 1843 (suppressed carrier frequency) and 1850. That makes two or at best three SSB channels for contesters running more than 10 W average.

It is my opinion that during the major contest weekends the 160 meter band plan should be set aside. Compare it to the following situation: In Europe most major roads have bike tracks alongside major highways. A few weekends every year, when major cycling events take place (the Tour de France, for example), bikers can use the entire width of the road. Let it be like that during a few of the major contests. Why does the band plan allow CW fanatics (and I am one of these) to transmit all over the band, while the poor phone guys, who actually need much more room, can only occupy a small portion (in Europe)? Can't CW fans relax during two or three contests every year and let the phone guys enjoy their contest?

And can't the phone operators relax a few weekends every year when the major CW contests are on? They could take the family out for a weekend. Why do some operators have to start QRMing QSOs under those circumstances? It saddens me to see that many people cannot appreciate that other people want to enjoy the hobby as well.

Let's use the entire width of the road when the Tour de France is on! I love the way Mike, N2MG, put it on the Contest Reflector: "A band plan, to me, is a lot like handicapped parking. Nothing is more frustrating than driving around a small parking lot over and over trying to find a place because I don't want to offend anyone by using one of several empty handicap spaces. When the lot is fairly empty, the dedicated spaces make sense — as do band plans. When at capacity, they do not. Blindly following band plans during a contest is like telling someone (those supposedly protected by the plan) that their transmissions are more sacred than the contesters."

In the Fourth Edition of this book I made an appeal to the responsible people in the many European countries to try to come to common regulations. Fortunately we can say that there has been a positive move in the past years. More countries in Europe now can operate above 1850 kHz, be it with only 10 W average power. Amateur Radio is, however, a secondary user in that part of the band, although I must say, from firsthand experience that I seldom, if ever, hear any other users of the entire spectrum below 2000 kHz. If we keep pushing, I am sure we can get the rest of that wonderful band with no power restrictions. Let's go for it! We need permission for more than 10 W to make this a hunting ground for DX.

Table 2-2 shows the Top Band European frequency/power allocations as of February, 2008. As compared to the table published five years ago, a large number of countries in Europe now have allocations above 1850 kHz. The data shown are for holders of the "full" license. In many countries holders of a beginner's license can operate with 10 W.

Today in Europe most countries allow high power on Top Band, at least at the bottom end of the band (1810 to 1850 kHz), where in the past only 10 W was allowed! Indeed, we should not forget that we have seen dramatic improve-

Table 2-2
European 160 Meter Allocations
As of February 2008

Country	*Allocation*	*Power*
Belarus	1810 - 1840	10 W CW
	1840	10 W SSB
Belgium	1810 - 1850	1000 W
	1850 - 2000	10 W average
Bulgaria	1810 - 1850	1500 W
	1850 - 1880	1500 W????
Croatia	1810 - 1900	1000 W
Cyprus	1810 - 2000	400 W
Czech Rep.	1810 - 1850	750 W
	1850 - 1900	75 W
	1890 - 2000	10 W
Denmark	1810 - 1850	1000 W
	1850 - 1900	10 W
	1930 - 2000	10 W
Estonia	1810 - 1850	1000 W (or 100 W for visiting foreign ham –T/R 61-01)
	1850 - 1955	1000 W (or 100 W for visiting foreign ham –T/R 61-01)
Finland	1810 - 1850	1000 W peak
	1850 - 1855	15 W
	1861 - 1906	15 W
	1912 - 2000	15 W
France	1810 - 1850	500 W
Germany	1810 - 1850	750 W
	1850 - 1890	75 W PEP
	1890 - 1950	10 W PEP (On special request)
Israel	1810 - 1850	1500 W
	1850 - 2000	40 W
Greece	1830 - 1850	500 W
Hungary	1810 - 1850	1500 W
	1850 - 2000	10 W
Iceland	1810 - 1850	1 KW
Ireland	1810 - 1850	1.5 kW
	1850 - 2000	10 W
Italy	1830 -1850	500 W
	1820 -2000	10 W
Lebanon	1810 - 2000	100 W
Lithuania	1750 - 1800	10 W
	1810 - 1850	1000 W
	1850 - 2000	10 W
Monaco	1810 - 1850	100 W input
	1850 - 2000	10 W
Moldavia	1810 - 2000	10 W
Montenegro	1800 - 2000	1500 W
Netherlands	1810 - 1880	400 W pep
Norway	1810 - 1850	1000 W
	1850 - 2000	10 W (1 KW during selected contests)
Poland	1810 - 1850	500 W
	1850 - 2000	10 W
Portugal, Azores, Madeira	1810 -1830	Max 200 W
	1830 -1850	Max 1500 W
RSA	1810 - 1860	400 W PEP
Russia	1810 - 2000	10 W
San Marino	1810 - 1900	1000 W
Slovakia	1810 - 1850	1500 W (3000 W in contest)
	1850 - 2000	10 W
Slovenia	1810 - 2000	300 W
Spain	1830 - 1850	200 W
Sweden	1810 - 1850	1000 W
	1930 - 2000	10 W
Turkey	1810 - 1850	30 W
United Kingdom	1810 - 1850	400 W erp
	1850 - 2000	32 W erp

ments in the regulatory scene in the past five years. There is reason to hope that further improvements can be realized in the next few years.

According to ITU the power limitation in Region 1 should be 10 W (average) above 1850 kHz. We nevertheless see a number of countries that do not follow that rule. As neighboring countries of those that do allow high power above 1850 kHz apparently do not suffer from it, we should have a good argument to make it like that for everybody. Our next goal should be to see the 10 W power limit above 1850 kHz officially abolished.

In Europe, 1840 kHz is the usual bottom end of the phone band. But it appears many operators are not aware that if they operate on a carrier frequency of 1840 kHz on LSB their sidebands spread 3 kHz down and that they are therefore taking out primary DX CW frequencies in Europe. Fortunately, the IARU band plan now clearly stipulates that 1840 kHz is the bottom end of the transmitted spectrum, which means no one should transmit (on LSB) below a carrier frequency of 1843 kHz.

Whereas in Regions 1 and 3 there is no provision for a band section reserved for DX work, 1830 to 1840 kHz has now been made such a DX window in Region 2. This seems to be the correct choice, as only 1830 to 1850 kHz has the primary and exclusive status in Region 2. Whereas in past years 1815 to 1825 kHz seemed to have developed as the *de facto* DX band, this is a very bad choice in Europe where 1810 to 1825 kHz is packed with signals of a radiolocation system, spreading from approximately 1810 to 1825 kHz.

Although in past years DXpeditions seemed to operate mostly between 1822 and 1828 kHz, it would be much better, given the QRM from radiolocation systems, for these stations to use frequencies between 1825 and 1835 kHz.

The IARU band plan decisions to use 1838 to 1840 kHz in Region 1 and 1800 to 1810 kHz in Region 2 for digital modes is a big mistake. It certainly ruins some choice DX frequencies on Top Band. It would not be a bad idea to consult with the Top Band community before making decisions like that…

About five years ago Tim Duffy, K3LR, asked European stations to tell him which were the "poor" frequencies on 160. It is clear that in many countries people suffer from harmonics of broadcast stations, spaced 9 kHz apart. Bad frequencies are 1809, 1818, 1827, 1836, 1945, 1854 etc (every 9 kHz). These are caused by harmonics or spurious emissions from broadcast transmitters. It appears that in Japan they use the same 9 kHz spacing as they put the same 9 kHz spaced frequencies on their black list. In the US the spacing between AM medium wave broadcast transmitters is 10 kHz, so 1810, 1820, 1830, 1840 etc are to be avoided. So, it is a good idea for DX stations to avoid using 1818, 1820, 1827, 1830, 1836, 1840, 1845, 1850 kHz etc.

To these we now need to add 1812.1, 1812.9, 1813.7, 1814.5 1815.4, 1816.2, 1817.1, 1817.9, 1818.7, 1819.5, 1820.3, 1821.1, 1821.9, 1822.7. 1823.5 and 1824.5 kHz. For stations in the Baltic Sea area and Europeans station beaming across that area, these frequencies are a total loss to work weak stations. Fortunately the signals are very clean and the space in between the modulation channels are usable even for weak signal work.

When you make a sked with a rare DX station, you will feel tempted to give him an "easy" frequency like 1825 or 1824 kHz. It's easy for everybody, so everybody chooses those easy frequencies for a sked. Use 1825.6 or 1824.4 kHz or something similar and unusual and there is less chance that

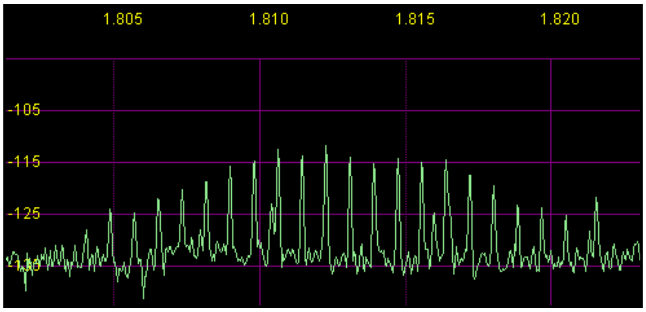


Fig 2-6 — Spectral display (created with *PowerSDR* and an Elecraft K3) showing the 20+ kHz wide non-amateur signal spreading from approximately 1800 to 1823 kHz.

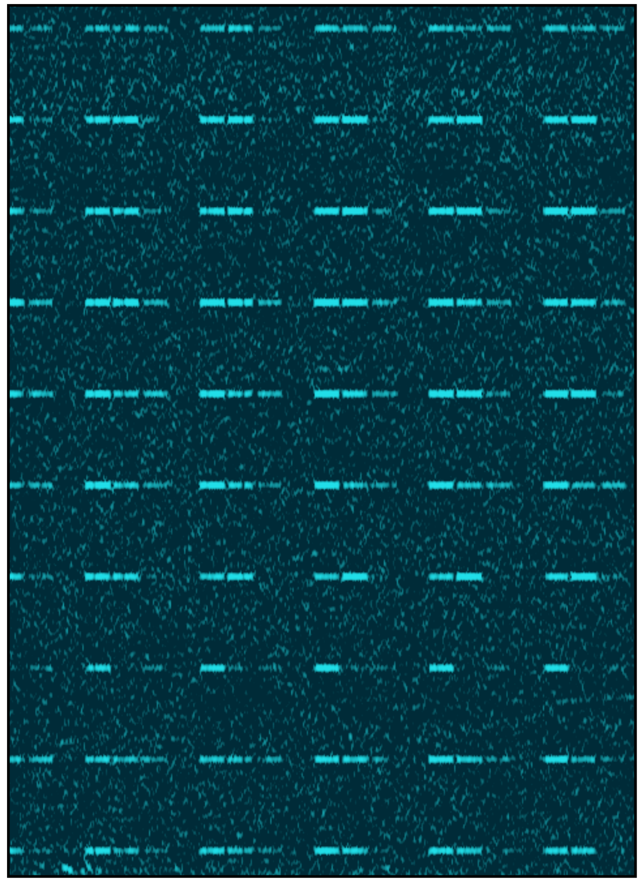


Fig 2-7 — Fast waterfall display (using *CW Skimmer*) showing nine of the approximately 24 modulation channels of the non-amateur signal shown in Fig 2-6. Each of these channels are separated ~825 Hz.

you will end up on a frequency already in use by someone else!

Whereas 1810 to 1850 kHz is an exclusive frequency range reserved for the Amateur Radio Service in Region 1 and 2, in Region 1 there are a number of footnotes telling that 1810-1850 kHz is a primary allocation for other services (Fixed and Mobile) in quite a few countries, the main ones being Belgium (nobody else *is* using it except Amateur Radio), Denmark, Spain, Greece, Italy, Netherlands, Russia, and Ukraine.

This situation of course opens the doors wide for these countries to use them and ruin this wonderful DX band which really should be an "exclusive" amateur band for all of us (between 1810 and 1850 kHz). The IARU should make a major effort to try to liberate this 40 kHz section of 160 meters from *all* other services.

Two kinds of non-Amateur Radio signals have been invading this section of 160 meters more and more over the past years. These signals have not been positively identified so far. One of them is thought to be coming from the Baltic Sea area (probably Russia) and is suspected to be coming from a radiolocation system (**alancordwell.co.uk/radionavigation/systems.html**). The signals consists of approximately 20 equally spaced CW-like signals (exactly 820 Hz spacing) that send three dashes (letter "O") at an equivalent CW speed of approximately 25 WPM. The third dash seems to have a signal strength that is down about 20 dB vs the first two dashes (see **Fig 2-6** and **Fig 2-7**). It needs no saying that this signal to a great extent ruins the JA DX band in Europe, and only JA stations that, by coincidence, operate half-way between two of the modulation channels are free of QRM.

Recordings made of these signals are on the CD that comes with this book (*1800-1820-CW-narrow.mp3, 1800-1820-SSB-3kHz-BW.mp3 and 1800-1820-AM-5kHzBW.mp3*). Anyone who can positively identify these signals, please contact the author.

Another group of "annoying" signals comes from what I think might be drift net or fishing buoy beacons. These are battery-powered, low-power transmitters mounted on a buoy and using a short vertical, usually transmitting with 10 W over a perfect reflector (the sea). In recent years these very bothersome gadgets seem to have multiplied here in Europe. One night in December 2008, I counted 37 such buoy signals, ranging between just audible and S9 between 1810 and 1850 kHz! These gadgets are used by fishing fleets to locate the buoys that hold their sometimes mile-long nets. They operate most frequently between 1.8 and 3.6 MHz (yes I have heard several in the 80 meter CW band as well!).

These beacons have become a substantial bother lately. They drift all over, not only on the sea, but also in frequency! You may start a DX QSO on a given clear frequency, and 10 minutes later one of those signals is slowly drifting onto your channel. In Europe all these buoy signals seem to come from the direction of the Baltic Sea or the Bering Sea, and some of them are even heard at noontime (by ground wave they must be coming from the North Sea?).

It appears that these transmitters are not licensed by any authority and are difficult to tackle. It is said that the most effective way to get rid of them is to operate on or very close to the beacon frequency. If the fishing fleet cannot hear the beacon reliably, they will likely change frequency. All of this to say that none of the frequencies in the 160-meter band are permanently "good" frequencies.

A few recordings made of these signals are also on the CD that comes with this book (*1820.mp3*, *Beacon-1812-0930z.mp3* and *Beacon-1819-0900z.mp3*). Anyone able to positively identify these signals, please contact the author.

Here are some other frequencies to remember:

Operating frequencies in South Africa

1810 to 1838 kHz: CW
1838 to 1840 kHz: digital modes
1840 to 1850 kHz: phone and CW

Operating frequencies in Australia

1800 to 1810 kHz: digital modes
1810 to 1840 kHz: CW only
1840 to 1875 kHz: phone

As far as I know all countries in the world can operate around 1810 to 1830 kHz. The only country left with a somewhat restricted playing ground is Japan, where 1810 to 1825 kHz (and 1907.5 to 1912.5 kHz) are the band limits. Over the past 20 years we have come a long way as far as attaining 160 meter band frequencies and standardizing them!

4.3.2. Where to Operate on 80 Meters

4.3.2.1 The Formal DX Windows

Although the 80-meter band is not allocated uniformly for all continents and countries, this isn't really a problem for the DXer. On CW all countries have an allocation starting at 3500 kHz. The DX window for CW is the same all over the world: 3500 to 3510 kHz. A secondary *de facto* window exists between 3525 and 3530 kHz, which is the lower limit for General and Advanced licensees in the US.

The 80-meter SSB DX window covers from 3775 to 3800 kHz. While the 3500 to 3510 kHz CW DX window has been internationally recognized by the IARU in both Regions 1 and 2 (see Section 4.1) for a long time, the phone DX window (3775 to 3800 kHz) only became a DX window in the Region 2 band plan starting in January 2008. Not that it really was a major problem, as it is common sense that reigns anyhow, since IARU band plans are voluntary in a great majority of countries. It really is a gentleman's agreement that we should all follow, at least if it makes common sense! If not, we should ask our national societies to change their band plans.

In the middle of the day the 80-meter DX segments can be used for local work, although we should be aware that local QSOs can cause great QRM to a DXer at a few hundred miles to, say, 1000 miles in the direction of the terminator. This station, which is already in the gray line zone, may be enjoying early gray line DX propagation conditions at his QTH. In Europe situations like this occur almost daily in the winter, when northern Scandinavian stations can work the Pacific and the West Coast of the US at 1300 to 1400 UTC, while Western Europe is in bright daylight and does not hear the DX at all. Western Europeans can hear the Scandinavians reasonably well, and consequently the Scandinavians can also hear Western Europe well enough to see the DX covered by QRM. The same is true for North America when local rag chews among stations to the east can interfere with DX for more westerly stations that are still in darkness. Hams must be aware of these situations so they don't interfere with DXers in other adjacent areas.

Fig 2-8 — Saka, JA1HQT, has one of the best signals on 160 and 80 meters from Japan. Saka uses a wire Four Square on 160 meters. On 80 meters, he uses a 3-element linear loaded Yagi at 36 meters height (good for 275+ countries).

Most countries in Europe, including Russia and the CIS (former USSR) countries can operate anywhere between 3500 and 3800 kHz, and in most countries there are no mode restricted subbands imposed by the government. The band plan for Russia and CIS countries has changed in the past and is now the same as in all Western European countries.

Operating frequencies in Australia:

Australia has a somewhat peculiar band plan. The most important change in recent years was the expansion of the SSB DX section from 3775 to 3800 kHz.

3500 to 3700 kHz: CW
3535 to 3620 kHz: SSB
3620 to 3640 kHz: Digital modes
3640 to 3700 kHz: SSB
3700 to 3776 kHz: No Amateur Radio
3776 to 3800 kHz: DX window
Note there is no provision for a formal CW DX window.

Operating frequencies in New Zealand:

3500 to 3550 kHz: CW
3550 to 3900 kHz: CW and phone
3620 to 3640 kHz: digital modes

Operating frequencies in Japan:

3500-3520 kHz: CW only
3520-3525 kHz: Digital modes and CW
3525-3575 kHz: CW and Phone
3599-3612 kHz: CW and Phone
3680-3687 kHz: CW and Phone
3702-3716 kHz: CW and Phone
3745-3770 kHz: CW and Phone
3791-3805 kHz: CW and Phone

Operating frequencies in South Africa:

3500 to 3510 kHz: only DX CW
3510 to 3580 kHz: CW
3580 to 3620 kHz: Digital modes and CW
3630 to 3800 kHz: SSB

The 80-meter band plan for the USA:

Since the FCC decided to expand SSB privileges in the US from 3800 kHz first to 3775 kHz, later to 3750 kHz and most recently to 3600 kHz for Amateur Extra Class amateurs (3700 kHz for Advanced and 3800 for General), the DX window has expanded from below 3750 to 3800 kHz during openings to the US. The top 25 kHz remains the focal area.

4.3.2.2. In Practice

Many amateurs are unaware that 80 meters is a shared band in many parts of the world. In the USA, 80 meters sounds like a quiet VHF band compared to what it sounds like in Europe. Because of the many commercial stations on the band in Europe, the 25-kHz DX window can often hold only a handful of QSOs in between the extremely strong commercial stations in the local evening hours. If you are fortunate enough to live in a region where 80 meters is either Amateur exclusive or not heavily used by commercial stations, please be aware of this and bear with those who must continuously fight the commercial QRM.

There now is a common (IARU) DX window in Regions 1 and 2, but, as it is the case with all band plans, these gentlemen's agreements are not respected by all. US stations complain bitterly about poor cooperation from rag chewers, who have another 200 kHz that could be used for their local contacts. In Europe we often see commercial stations popping up in the middle of the DX windows, probably because it often is relatively quiet in these windows.

The increased popularity of 80-meter DXing, together with the few DX channels available in the phone DX window, have created a problem where certain individuals would sit on a frequency in the DX window for hours (it seems like days) on end, without giving anyone else a chance. This problem is nonexistent on CW, where you have an abundance of DX channels in the DX window. This situation is also an excuse for creating DX nets (see section 12), where ethics are not always the highest.

4.3.3. Where to Operate on 40 Meters

When I write this paragraph it is March 29, 2009. The magic day. At 0246 on this magic day, Randy, K5ZD, a well known avid low-band DXer and contester, sent the following short but very meaningful message to the contest reflector "40 meters without broadcast QRM. Never thought I would hear that in my lifetime. Completely changes the nature of 40-meter SSB contesting."

Fig 2-9 — Most of us only chase DX. Bob Ferraro, W6RJ, chases real game as well. Bob has one of the stronger signals on 80 and 160 meters from the West Coast. He uses a 3-element Yagi (80 meters) and a wire Four Square (160 meters) from his mountaintop QTH about 30 miles east of the San Francisco Bay.

Checking 7100-7200 kHz in the following days revealed a situation that I thought might happen. The serious or well intentioned countries followed the ITU decision and moved their BC stations off these frequencies, and even during the evening hours, when that section of the band used to be a complete loss, hams can now work stations and even DX! Too bad some of those BC stations from countries such as North Korea, China and Ethiopia remain in place for the time being, but we can say that the new situation is a 99% improvement.

A little history explaining why this is a historic day: Before 1938, the entire 40 meter band, from 7000 to 7300 kHz was allocated exclusively to the Amateur service, worldwide. This allocation was obtained at the first "frequency conference" in history (Washington 1927) and successfully defended during the 1932 conference in Madrid. But in the late 1930s, some governments began lobbying to get part of our 40-meter band for broadcasting. By then it had become clear that the 7-MHz band was a prime frequency segment having excellent propagation characteristics for what they had in mind. And what did they have in mind? To have a top notch propaganda tool in the years leading to WWII and later during the Cold War years. In 1938 the Cairo frequency conference decided to give a major chunk of our best band to the broadcasters, provided they did not interfere with hams in Region 2 (North and South America). It is obvious that none of these brute force broadcasters ever took the non-interference issue seriously and hams in North and South America got stuck with terrible interference and in Europe, Africa, Asia and the Pacific 7100-7200 kHz became a total loss.

It took until WRC-03 that the "theft" of one of our major bands was partially rectified when the conference voted to evict the broadcasters from 7100 to 7200 kHz and return that portion of the band to hams in Regions 1 and 3. The broadcasters were given six years to evacuate this "sacred" terrain. After March 29, 2009, 7100 to 7200 kHz was to be free of broadcast stations and all the interference they generate. At the same time, in Region 2 the whole band 7000 to 7300 kHz remains exclusively reserved for the Amateur service. The broadcasting band in Regions 1 and 3 changed to 7200 to 7450 kHz effective March 29, 2009, while in Region 2, it remains 7300 to 7400 kHz.

As a result, the problem of "too little space" was solved

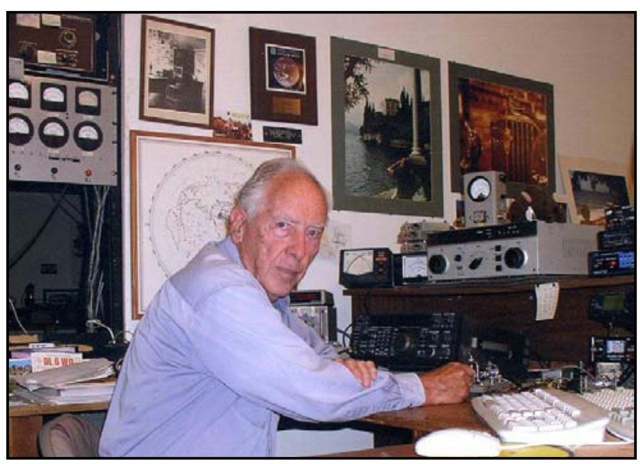


Fig 2-10 — Bob Leo, W7LR, alias Mr Montana on 160 and 80 meters. If we don't hear his signal in Europe, we know the band is dead to the "black hole."

to a very large extent. Since March 29, 2009, it is no longer necessary to operate split in order for stations in Region 2 to work stations in Regions 1 and 3.

Awaiting the magic day of March 29, 2009, the IARU societies in the three IARU Regions, and more in particular in Region 1 which was most affected by the broadcast invasion of 1938, prepared for the happy day and the following years by drafting a new and — of course — totally revised band plan. In addition the band plans in the three Regions were matched as much as possible (see Section 4.2)

I would suggest that all serious low banders adhere, as much as possible to this band plan. There is no longer an excuse to work phone in the CW subband, or to work CW way up in the phone band. Let's all of us stick to our own territory.

It will undoubtedly require some "getting used to it" but let's all do our best so that everyone feels that the new situation is a happy ending to a rather sad story.

5. SPLIT-FREQUENCY OPERATION

The split-frequency technique is a must for a rare DX station or DXpedition working the low bands. It should logically also be the way we work the much-in-demand DX on 160 meters. Signals on Top Band are often so weak that working split should really be the rule rather than the exception!

It is the most effective way of making as many QSOs as possible during the short low-band openings, because the weak signals often encountered are conducive to chaos if stations are calling the DX station on its frequency. It also gives a fair chance to the stations that have the best propagation to the DX station. With split operation the DXer with a good antenna and with good operating practice is bound to have a lead over the modest station. This is only fair. Why else would we build a station that performs better than the average?

In an answer to someone who said there never was good reason to operate split, Bill W4ZV, wrote: "The DXer realizes that his signal is weak compared to the hordes calling. In light of the increased numbers of callers brought by packet spots, this is even more understandable. He understands that *his* signal is apt to be covered up by those stations. Since many callers have adopted the "call until doomsday" technique, the DX is much less likely to complete a QSO within a reasonable period of time. Spreading the pile has only one goal: to make it more likely for the DX station to pull a call out of that mass of noise. The split frequency method attempts to make the pileup more efficient and to work the maximum number of stations in a given period of time. In some latitudes, that window of opportunity can slam shut very quickly. The DX op may grow weary of getting up well before dawn for mornings on end in order to be able to log just a few QSOs even though he hears a swarm of callers. What we need to do is listen to what the DX op wants. If it appears at odds with your personal operating ethics, don't call him. If you'd like to be in his logs, follow his instructions and observe what he is doing. What we really need is some restraint on the part of callers when the DX station comes back to someone rather than a continuation of this mindless calling, calling and calling."

Here are the "why" and "how" for working split on the low bands:

1) The "why": As the rare DX station stops its CQ, there are likely to be many stations calling, all at the same time. Though the DX station operator might pick out a good strong signal, others may continue calling him, and his reply to a particular station may be lost in the QRM. This will result in a slow QSO rate, even though the DX station hears the callers well.

2) The "how": By working split (listening on a frequency other than the one he is transmitting on), the callers will have more chance to hear the DX station reply to a particular station, as the pileup of callers is not on the same frequency where the DX station is transmitting. In this case the DX station should simply specify a single frequency where he will be listening.

3) The improved "how": If there are a large number of stations calling on the single split frequency specified by the DX station, chances are real that the DX station cannot discriminate among those calling simultaneously on the same spot. The only solution is that the DX station specifies a frequency *range* where he will be listening, in order to spread out the callers and make the layer less thick. He may specify "up 3 to 5" for example.

A few general rules apply for split-frequency operation:

1) If possible, the DX station should operate in a part of the band where the stations from the area he is working cannot operate (there are not that many left!), or in a section of the band that is generally considered the DX section.
2) The DX station should transmit both its call and its listening frequency after every QSO. It only takes a second to do so, and it goes a long way toward keeping order. It also helps to keep a constant rhythm in the operation, which is very important.
3) The listening frequency should be well *outside* the DX window. Excellent point! Too often I hear a DX station on 3503 kHz listening 5 up, ruining a major part of the DX window. There is no reason why he should not listen 10 or 20 up. The same applies to phone operation, where the DX station transmitting in the window should listen outside the DX window for replies.
4) If the DX station operator is working by numbers (call areas), he should exercise authority to reject those calling from areas other than those he specifies. He should not stay with a particular call area too long. At five stations from each area, at a rate of three QSOs a minute (that's fast!), it takes almost 20 minutes to get through the 10 numbers. He should always follow the normal order of the numbers — Ø, 1, 2 etc — so the crowd will be able to estimate how much longer they have to wait. Always go through an entire number cycle. *Better yet, avoid working by numbers if you can.*
5) A better method than working by numbers is to work by geographical areas. This also gives a better chance to remote regions of the world, where signals are often weak and openings shorter. If you use this technique because the pileup is too dense, rotate quickly between the continents or areas.
6) If there is a sudden drop in QSO rate, the DX station operator should check if his transmit frequency is still clear. If not, changing the transmit frequency a few kHz may bring relief. He should inform the crowd if he moves.
7) If the DX station's listening frequency is being jammed, he should specify a frequency range

Fig 2-11 — Wally Eckles, W8LRL, started DXing on 160 in 1972. As of mid-2010 he had nearly 320 DXCC countries confirmed on Top Band.

instead of a single listening frequency. On CW the split should be at least 2 kHz (preferably 5 kHz). For splits less than 2 kHz the pileup's key clicks are likely to spread onto the DX station's frequency. On phone it should be at least 5, preferably 10 kHz.

8) The DX station should always bear in mind that he should keep his listening range as narrow as possible; there are others stations that want to use the band as well. The quality of the operator at the DX end (or on a DXpedition) is often judged by how wide a listening range he needs to handle the pile-up!

5.1. Working Split on 160 Meters

Rare DX stations usually generate a pileup, which means they should as a rule operate split frequency. The generally accepted transmit window for DX stations is 1820 to 1830 kHz, with 1822 to 1828 kHz as the most popular range. It is good practice not to use frequencies that are a multiple of 1 kHz (eg 1824, 1826); but rather frequencies like 1823.3 1824.6 kHz etc (see also Section 4.1.). Do not use 1820, 1827 or 1830 (integers, multiples of 9 or 10) either (see also Section 8.2.).

The rare DX station, when operating split, should listen outside the *de facto* DX window of 1820 to 1830 kHz, and at least 2 kHz (and preferably 5 kHz) above its transmit frequency to avoid being disturbed by the key clicks of the calling stations.

5.2. Working Split on 80 Meters

On CW the main reason for the DX station to go split is when the pileup gets too big, in other words when the QSO rate drops. Another nice reason for the DX station operating in the CW DX window is to listen "up 25" to cover General and Advanced class stations in the US.

On SSB I can think of many good reasons to go split. In the first place, going split helps to not occupy the DX window more than necessary. Therefore the DX station should always indicate a listening frequency outside the DX window (and preferably below 3750 kHz). US stations wanting to work Europe should transmit above 3800 kHz and listen below 3750 kHz to keep congestion in the DX window as low as possible. The same holds true for JA stations working Europe. In a nutshell: Never occupy more than one channel in the DX window!

Middle East stations should transmit below 3750 kHz when working North America, and listen above 3800 kHz to avoid QRMing European stations. Stations in the Pacific working Europe should (if possible) transmit above 3800 kHz and listen below 3800 kHz or better yet below 3750 kHz to avoid US QRM.

It is not reasonable for a European to transmit inside the DX window (3780 kHz, for example) and listen on 3805 kHz. If this is done, two windows inside the US subband are occupied for one QSO, and the potential for QRM and confusion is increased. The inverse situation is equally undesirable.

If a European station wants to optimize the QSO rate (say, in a contest), it should transmit below 3750 kHz and listen above 3800 kHz. Every year I hear hordes of European stations trying to work USA stations in the ARRL DX phone contest in the *de facto* DX window (3750 kHz to 3800 kHz), where they must overcome local US and Caribbean-made QRM. I always enjoy doing the contest just below 3750, listening above 3800 kHz, and never have any such problem.

5.3. Working Split on 40 Meters

Prior to March 29, 2009, the nature of the frequency allocations in different regions made split-frequency operation a very common practice on 40-meter phone.

As a result of actions taken at the World Radio Conference in 2003, the 40 meter band in Region 1 has now grown from 100 kHz to 200 kHz. There no longer is any need for working split frequency on 40 meters during normal operation, and it should be avoided as much as possible, because split

Fig 2-12 — Rio, JA1JRK (highest 160 meter DXCC score in Japan with ~275 countries), and Toshi, JA1ELY. All Rio's equipment is home made (including all the antennas, 160 meters through 23 cm). He has one of the better JA signals on both 80 and 160 meters. Toshi, JA1ELY is the editor of the Japanese DX magazine. He also made the translation of the Second Edition of the *Low Band DXing* book. His 2-element quad on 80 meters also puts out a walloping signal.

operation inevitably takes much more space than working on a single frequency.

There will of course always be good reasons to go split, such as working a huge pileup from a rare place. Sound rules and good behavior will come about in the first "post March 29, 2009" years.

5.4. Working Split in a Contest

I don't think split frequency CW contest operation is done at all on 40 and 80 meters. Doing it on 160 meters has been the subject of discussion recently. So far it is seldom done, although it may be the only way to make contacts. On the other hand, our narrow 160-meter band is already fully congested during these contest periods, so we should be careful to not use two frequencies instead of one for making our contacts. This would inevitably make the congestion even worse.

Only for very rare stations operating during a contest on Top Band is there an excuse to do so. The way to do it is to listen 1 kHz up, and not over a wide range. If KH6xyz has propagation to Europe during the 160 meter contest, I bet he will make very few contacts, if any, if he won't work split. In such a case working split is probably the only way to make some QSOs, even in a contest.

5.5. DX Subbands in 160 Meter Contests?

Over the past years we've seen "rules" for DX subbands on 160 meters come and go (mainly in contest rules). This is especially a critical issue on Top Band, since local stations are extremely strong and DX stations are usually very weak, much more so than on 80 and 40 meters.

The classic scenario we've seen on Top Band is to reserve 1830 to 1835 kHz only for a DX station to call CQ Contest, after which others could reply to his CQ. The problem is "What is DX?" Usually DX means a station outside your continent. This means a P4Ø or a PJ2 in South America could sit in the DX window and works hordes of US stations in North America, while a KP2, FM or FG station cannot because they are also in North America.

And when the band is open between the US and Europe, who's DX? The Ws are DX to me and I am DX to the Ws, so who should be in the DX window? What should I do if I'm in the window and a European comes back to me? He may even be a new multiplier for me. Typically he is just a nice guy, who wants to give me some points. I could ignore him but he will probably keep on calling. So do I have to chase him off with a curt "QSY, I can't talk to you"? That's not a very nice thing to say to someone who's new to the game or who just is trying to do me a favor.

How about considering two windows, one where the USA can call CQ, and one for Europe. But where should the Africans and others go in this scenario? Well, then I guess we need four DX windows, one each for the USA, Europe, Africa and Asia (the Pacific can use the European window since opening times do not really overlap). But the 160-meter CW band is only 30-kHz wide. Let's see — we can reserve 10 kHz for Europe, another 10 kHz for the USA, and 5 kHz each for Asia and Africa. But that does not solve the problem: Where do the European stations go that want to work Europe? And how about the US stations that want to work US stations? They can do it in the US window, which means they will have 10 kHz, and they will all be sitting one on top of another in this crowded space. Impossible! All of that chaos, while the African and the Asiatic windows will be half empty with the small amount of activity from there.

Fig 2-13 — Jerry Rosalius, WB9Z, one of the leading 160 and 80 meter DXers from the Chicago area. Jerry was the first in the US Midwest to reach the DXCC 300 mark on 160 meters.

Well, maybe, we could give zero points for working your own continent. That would be the end of my contesting on Top Band. I don't want to spend 30 hours working only 150 DX stations outside of Europe. That would be really boring. Today we have a vibrant and exciting worldwide contest for everyone to participate in. If we make it a pure DX-to-DX contest it would be a dull, boring and insignificant contest.

So, let's forget about these DX windows and find technical solutions to the problems we face on Top Band. Let's clean up our transmitted signals and use better, more directive and more selective receiving antennas. Let's concentrate our energy on creating solutions rather than workarounds. The more I think about it, the less sense these DX windows make to me. Considering all of the above it looks like it was a wise decision for CQ World Wide 160 Meter Contest to abandon the DX window beginning in 2009.

6. ZERO BEAT AND RIT (THE CLARIFIER)

Two stations are said to be at zero beat when they are transmitting on exactly the same frequency. The term zero beat comes from the fact that if two stations transmit on exactly the same frequency, the resulting beat from mixing the two signals would have a frequency of zero Hz.

Unless working split frequency, it is common practice to zero beat, or better said, stations *should* be at zero beat, on phone as well as on CW. This does not mean this is always the case. A major advantage of a CW QSO is its narrow bandwidth (a few hundred Hz), provided both stations in a QSO transmit on the exact same frequency.

On CW we very often hear that stations in a QSO are not on the same frequency, but a few hundred Hz apart. For this there are two major reasons (often a combination of both):

- The first reason is that the operator does not apply the correct zero beat procedure.

• A second reason is the incorrect use of the RIT (receiver incremental tuning) or clarifier on the transceiver. Most modern transceivers have an RIT function which makes it possible to listen on a frequency which is (slightly) different from the transmit frequency.

On modern transceivers the frequency of the CW sidetone monitor (pitch) is adjustable (usually between 300 Hz and 1000 Hz), and automatically tracks the BFO frequency offset. In this case the zero beat procedure consists of making sure that the pitch of the sidetone (CW monitor signal) of the transmitter is at exactly the same frequency as the tone (pitch) of the station you are receiving. If you listen at 600 Hz and the sidetone pitch is set at 1000 Hz, you will transmit with an offset of 400 Hz with respect to the station you are working (the direction of the offset depends on whether you listen on LSB or on USB).

Older transceivers may have a fixed CW monitor frequency, often 800 Hz. The offset in those older transceivers was such that you would only transmit zero beat if you tuned in the station you were contacting so that the CW beat note would be the specific CW monitor frequency (800 Hz). Many experienced CW operators listen at a fairly low beat tone (300 to 500 Hz) instead of the more usual 600 to 800 Hz. For most people a lower pitch frequency is less tiring during long period of listening and in addition it allows for better discrimination between close spaced signals. As a result of this, those stations are transmitting off frequency by 350 Hz or so on CW. This is not a problem if the receiving station uses wide selectivity (500 Hz or more) but it could be a real problem with 100 to 300 Hz selectivity (I almost always use 100 Hz or 200 Hz on the low bands).

On phone I have noticed that a fair number of operators like to listen to SSB signals that sound very high-pitched, like Donald Duck or someone who's inhaled some helium. That means that they tune in too high on LSB. To the operator at the other end of the link their transmission will sound too low-pitched, because they are no longer zero beat. You will need to correct for that, or your audio will just sound like mumbling.

Both of these problems (a low CW pitch with old vintage transceivers and listening Donald Duck style on SSB) can be corrected by using the RIT (or XIT – transmitter incremental tuning) but care must be taken to do this correctly:

CW*:* Let us assume our transceiver is designed for working CW at an 800-Hz beat note, and it is only when listening at this note that the transmit frequency will be exactly the same as the receiving frequency. If you want to listen at a beat note of 500 Hz and still transmit at zero beat, you will have to set the RIT at –300 Hz if you receive in L-CW (CW on lower sideband) or at +300 Hz if you receive on U-CW (CW on upper sideband). This is where the RIT function of an older transceiver can come in handy and help you stay on frequency. Again, with modern transceivers the variable frequency CW pitch (beat note, tone) tracks with the BFO offset for demodulating the CW signal, and you need not worry as long as you use the same frequency for reception and sidetone.

SSB: If on LSB you tune in for a high pitch (say 100 Hz off frequency), you will have to correct this by using the RIT set at +100 Hz or the XIT at –100 Hz (the inverse for USB).

When trying to get through a single frequency (not split operation) pileup, it may be advantageous to transmit your CW 100 Hz (or even a few hundred Hz) higher or lower than the station you are working. In this case RIT or XIT, or better yet, a second VFO can help you do that.

When trying to get through a single frequency phone pileup, it can be advantageous to sound a little high in frequency. On LSB you should transmit a little lower in frequency than the station you are calling. If you use your RIT, adjust it for +100 or +200 Hz. Use –100 to –200 Hz for XIT.

Although you would expect modern equipment to transmit at exactly zero beat (on CW) when both the transmit and the received tone frequencies are identical, it appears this is not always the case. Bill, W4ZV, recommends checking your actual transmit frequency with a separate receiver, especially when you get a new transceiver, to make sure you are placing your transmit signal where you think you are! He added "When I had an FT-1000MP it had some quirk I never figured out which required adding 70 Hz to the transmit frequency to be zero beat."

It's a good idea to do some checks with a local station (or with a second receiver) to make sure you are truly "zero beat" on CW. This will save you lots of frustration. In contests I have often cleaned up my frequency and worked even the very weak signals, only to find out that there was a guy with an S9 signal calling me 400 Hz up. He was strong but I never heard him, while I easily worked stations that were 40 or 50 dB weaker than he was...

I, for one, never use the RIT or XIT. To me these "gadgets" can only lead to confusion and be the reason for being off frequency rather than be on frequency (zero beat). I actually see no reason why a transceiver having two VFOs would need to have RIT or XIT. Why would I want to use that tiny RIT or XIT knob when I can use one of the two the main tuning knobs to do anything I want? However, if your transceiver only has one VFO you will need to have an RIT or XIT to be able to work split frequency.

Slightly off frequency: The newer radios can achieve very narrow bandwidths without filter ringing. Using a narrower bandwidth means less total noise power, and in turn a better signal-to-noise ratio. But you may miss the stations calling you off frequency. It is not uncommon to find stations calling several hundred Hz off your transmit frequency, for no valid reason. So, if you use a very narrow bandwidth (almost a must

Fig 2-14 — Jim Wilson, N7JW, Mr "almost" West Coast. If Jim, located in the southeastern corner of Utah is S9 on 160, then maybe we can hear the West Coast here in Europe.

during a Top Band contest), make sure you listen around your transmit frequency to find the off frequency stations calling you.

If you call a DX station in a big pileup (not working split), it is a good idea *not* to call exactly on the DX station's frequency but maybe 50 or 100 Hz up or down. I have often been in a situation where I had a large pileup of US stations calling me on Top Band, all exactly on the same frequency (my transmit frequency), where I could not make out even a partial call. A much weaker station that called me 50 or 100 Hz up or down made it however (frequency diversity). Another solution to this problem is to throw in your call when all the other callers are done (time diversity).

7. TAIL-ENDING

What is *tail-ending*? A tail-ender tries to outrun the competition by being *faster than his shadow*. He is listening to the station being worked by the DX station, and a split second before that station turns it over to the DX station, he throws in his call, usually half on top of that station… He is literally *stepping on its tail.* Strictly speaking, when tail-ending you are intentionally transmitting on top of another station, and hence causing interference to that contact.

There are two sorts of tail-enders:

1) The tail-ender who does not wait for the invitation to transmit from the DX station. He feels better than the other operators, he is fast and strong, and he is convinced he will outrun and outscream the others. Such operating behavior is not very polite and rather aggressive. Please, don't be one of those.

2) The tail-ender who gives his call five or even ten times, because all he heard was the last station making a contact with the DX station. He has not really heard the DX station at all, and thinks that by screaming loud and long he will be heard. W8JI said on this subject: "The real problem is people who *can't* actually hear the DX station calling, not slipping a single call in at the *real* finish of a contact. I think the length of a call is directly proportional to how little the caller actually hears. If a person wants to make a DX contact, it only makes sense they at least have to *hear* enough of the DX station to at least know when the DX station is transmitting, and at least be able to pick letters and numbers out of the call sign. It is the callers who cannot do this minimal amount, and still insist on calling, who are the real problem."

7.1 Tail-ending in a Single Frequency Pileup

Tail-ending is especially harmful when practiced in a single frequency pileup. Often the ongoing contact is not really over yet when the tail-enders start calling. As Mike, W4EF, explained on the Top Band Reflector: "I think tail-ending is especially inappropriate when a DX station is operating simplex. If the DX station is down in the noise and I am pretty sure (but not 100% sure) that he responds to my call with 'W4EF 559' and I then respond with 'QSL de W4EF W4EF UR 559 559 PSE CFM PSE CFM BK.' All I hear when I turn it back is a bunch of tail-enders. How do I know for sure that my contact was a good one?"

7.2 Tail-ending in a Split Frequency Pileup

In split operation these tail-end transmissions will not harm the transmissions of the DX station, so the DX station can just ignore these arrogant callers and work the stations that behave as they should. That's one of the main reasons why a DX station working under marginal conditions (weak signals) should always work split frequency.

Fig 2-15 — Dietmar, DL3DXX, is certainly one of the best low-band DXpedition operators that the DX community could wish to see on any DXpedition. Remember the splendid show he put on from VP6DX on Top Band in February 2008?

7.3 How to Prevent Stations from Tail-ending

If the DX station has adopted a standard pattern for making QSOs, other stations will know when exactly to call. The DX station must always signal when he starts listening for new callers, and he should always keep the same rhythm, the same pattern.

I have to admit that I occasionally use tail-ending. Why? Only when and because the DX station does not show any fixed operating pattern at all. He does not signal when he's ready to listen for callers (for instance by saying "QRZ" or "up 10" or simply his call sign), or he works W1XYZ after having sent "W2A you're 59" or "QRZ W2." The DX station is not giving a good example nor is he imposing any authority on the frequency. In this case it is the DX station who is causing all the confusion.

8. BEING THE RARE DX, GOING ON DXPEDITIONS

You don't have to be on a DXpedition to be a rare DX station. There are still dozens of countries where the number of licensed radio amateurs can be counted on one hand. Operating as a resident or temporary resident from such a much-wanted country is very similar to working from a DXpedition. The required expertise to make low-band DXing a success is the same as required from top notch DXpeditioners.

8.1. DXpeditions and the Low Bands

Forty years ago it was rare to have a DXpedition show up on 80 meters, and 160 meters was out of the question. With a few exceptions Top Band was just the "Gentleman's Band" for daytime rag chews. Fortunately there has been positive change over the years and for most DXpeditions 80 or 160 meters has become just another band where they can make QSOs and make the DXers happy. During the lower years of the sunspot cycle they are definitely capable of bringing in a lot more DX than 21 or 28 MHz! The 5-Band DXCC, 5-Band WAS and 5-Band WAZ awards, as well as the single band DXCC awards have also greatly promoted low-band DXing. So have the single-band scores and record listings in DX contests.

Today there are still a few stubborn DXpeditioners who will only appear on the low bands in the last one or two days of operation, if at all. Staying on bands with the best QSO rates will not result in many Top Band QSOs. But logic tells you to tackle 80 meters and more especially 160 meters from the first day, as there may not be low-band openings every day.

WØCD writes in his survey reply: "DXpeditions going to new countries should give more time to 160 to be sure there is decent propagation. Not just a few hours the last night."

Joerg, YB1AQS, from the famous ZL7DK team said the same: "As we've found all the years — if possible, the 160-meter antenna has to be the first one up and the last one down."

A most striking example happened during the November 2009 DXpedition to Chesterfield Island (TX3A) by AA7JV and HA7RY. The first night on the island there was once-in-a-lifetime propagation on Top Band from Europe to the south central Pacific. TX3A was up to S7 (with fading) more than two hours before European sunset (that is a 16,000 km path!). They were knocking off Europeans like it was 20 meters, not 160 meters. The next day, signals were *many* S-units down. This phenomenal duct propagation was simply awesome. Besides this once-in-a-lifetime propagation on Top Band, TX3A was working into Western Europe what seemed like at least 90% of the days they were active from Chesterfield Island. What a great operation! If BS7H had treated 160 and 80 meters in a similar way (instead of being on the low bands just a few hours only on the very last day), more stations would have been able to work them. One thing is for sure, propagation on Top Band is *not* predictable, and exceptional propagation days do occur. If you want to make use of these, be there (on the low bands) *every day*!

A DXpedition should prepare well for the low bands. Best is to include an experienced low band DXpeditioner in the crew. If that cannot be done, they should ask an experienced low band DXer to determine band openings for the low bands. Nothing is more frustrating than to hear a Southeast Asian station working Europe on 80 or 160, during the 10-minute window that this path is open to the US East Coast.

Over the past few years we have seen an increasing number of DXpeditions that concentrate on low-band operating. Thank you guys who do that!

8.2. DXpedition Frequencies

A DXpedition should announce its operating frequencies well beforehand on its Web site. If there is a good reason to change one of the frequencies during the DXpedition, they should announce it as well. It's also a good idea to publish several *escape frequencies* in case of QRM, intentional or not. Stick to the published frequencies, otherwise your credibility may suffer. When leaving one band or another mode, always announce where you are going and repeat the information several times (not too fast on CW!). Don't just disappear; your supporters will be unhappy!

The following are the frequency ranges DXpeditions as well as operators from rare DX locations should use (and mostly do use) when working on the low bands.

1) *40 meter CW:* Transmit anywhere in the first 25 kHz of the band. To avoid making QRM to other DX stations in the lower portion of the bands they should listen as high as possible in the CW section, preferably between 7025 and 7035 kHz.

2) *40 meter phone:* At the time of writing (April 2009) there is not yet a standard way of doing things in the new 40 meter band configuration. Observe, learn and use your good sense is the best advice I can give at this time. And, of course, stick to the IARU band plan.

3) *80 meter CW:* DXpeditions and rare DX stations should operate in the DX window (3500 to 3510 kHz). Sometimes we hear fishermen on upper sideband right on 3.5 MHz. Staying above 3503 kHz seems to be logical. DXpeditions should make it a point always to listen well outside the DX window, at least 5 kHz above the transmit frequency (to avoid key click interference). They should not forget to listen now and then above 3525 kHz for the US General and Advanced class operators.

4) *80 meter phone:* The DX phone window is from 3775 kHz to 3800 kHz. The DX stations working split should always listen outside the DX window (either above 3800 or below 3775 kHz).

5) *160 meter CW:* The transmit frequency range used by DXpeditions and rare DX stations goes from 1822 to 1828 kHz, with split operation from 1830 to 1835 kHz for areas where these frequencies are available, but not forgetting the 1810-1825 kHz JA window. Better yet is to use the window between 1830 and 1835 kHz (now officially the DX window in Region 2). Also, in Region 1 this window is (at this time) still clear from commercial radio positioning system QRM (see Section 4.3.1.). It is advisable *not* to use 1827 (harmonic of 9 kHz, the BC station spacing in many countries) and stay away from "round" frequencies. Do not use 1830, 1832, 1833 but rather 1830.4 or 1831.7 etc.

DXpeditions and rare DX stations as a rule operate split frequency, both on CW and SSB. The advantage of split-frequency operation on the low bands is even more outstanding than on the higher bands. The openings are often much shorter and DX signals can be much weaker than on the higher bands, while signals from local stations can be very strong. Working split makes it easier for calling stations to hear the DX. Otherwise, the strong pileup of callers will inevitably cover up the DX station, resulting in a very low QSO rate.

Sometimes we hear DXpeditions spreading the pileups over too wide a portion of the band. This is not generally advantageous for the QSO rate, and most of all, is very inconsiderate to other users of the band. It is also common for two DXpeditions to be on at the same time, both listening in the same part of the band. The net result of this is maximum confusion and frustration for everyone involved. There will inevitably be many "not-in-log" QSOs, where people ended up in the wrong log.

Equally bad practice is not telling the callers you are working split or where you are listening. People will (understandably) start calling on the DX station's transmit frequency, which together with frequency cops will in no time ruin the show.

8.3. The DX Operator in Control of the Pileup

Sometimes, you hear a beautifully smooth pileup. A dream! A pleasure to listen to! Pure music! Sometimes, it's pure chaos. Let me be blunt: It is the DXpeditioner who's responsible for either situation.

CT1BOH once wrote: "There is a price to be paid when a DX operator runs a pileup. That price is QRM and is totally dependent on the DX pileup operator skill. The better the skills of the DX pileup operator, the cheaper will be the price he has to pay for his show. At the same time, the better the skills of the DX pileup operator the better the pileup will behave because everybody will try to mimic him in admiration of his skills."

Here are a few hints for the DX operator on how to realize a smooth running pileup:

1) When working split, the DX operator should mention it after each QSO. For example, in CW: UP 5, UP 5/10, QSX 1820, etc. In SSB: "listening 5 up," "listening 5 to 10 up," "listening on 3770" etc. In CW, they should listen at least 5 kHz away from their transmit frequency, to avoid interference from key clicks originated by calling stations. In SSB, the split should be at least 10 kHz. Some signals of calling stations can be very wide and cause a lot of splatter on your transmit frequency. In any case, the DXpedition station should always listen outside the IARU specified or the *de facto* DX window of the band.

2) The DX station should have a well-thought-out strategy and rhythm, always following a standard pattern in his transmissions. The way he ends his QSO and invites the pileup to call should remain the same all the time. This will inspire confidence in the public. If he keeps following that same pattern, the pileup will know that when he sends "P5DX 10 up," he is listening again for new callers. The DX station should maintain this same pattern, the same speed, the same rhythm. This way everyone will know exactly when to call. It should be like clockwork.

3) The DX station operator should realize that a large number of callers may have marginal copy on him. Therefore he should keep instructions simple and repeat them all the time. He should live by his instructions and never make "out of turn" QSOs.

4) Calling *by numbers* (call areas) seems to have become one of the standard approaches to handling a pileup that's become too big. If the DX operator wants to apply this method, he should stick to the following rules:

- Once started working by numbers, go through all numbers at least once. If he disappears in the middle of a sequence, or starts working random numbers all of a sudden in the middle of a numbering sequence, this is inevitably going to create commotion.
- He should never forget that when working by numbers, 90% of the stations are waiting, biting their fingernails!
- He should always start a sequence with 0 (or 1), and move up in numbers one by one. No frills. He should not specify numbers at random: first 0s, then 5s, then 8s, then 1s etc… This will drive the pileup mad. If he follows a logical sequence, the pileup can more or less predict when it will be their turn. A random system will make callers utterly nervous.
- On the low bands the DX stations should work a maximum of five stations of each number and make sure to work approximately the same total of stations per number. Even at five stations a minute, which is very fast for the low bands, it will still take time to complete the circle. This means some stations will have to wait and sit idle. Propagation conditions can change during the wait.
- This "by the numbers" method is no good for running a pileup on 160 meters since the propagation usually is quite area-selective anyway.
- By the way, this method is almost never used or needed on CW.

5) Calling *by continent or area* is a better method for handling large pileups on the low bands. The main reason for this is that long haul propagation on the low bands is always area sensitive. This method also gives a better chance to remote regions of the world, where signals are often weak and openings shorter. Here are a few recommendations for the DX operator who wants to use this method:

- Use this technique primarily to reach those areas of the world that have poor propagation or short openings to you.
- If you use this technique because the pileup is too dense, rotate quickly between the continents or areas. A good rule of thumb is that one should not stay with the same area for longer than 15 to 30 minutes.

6) Inform the pileup of your plans, tell them exactly how you will rotate between areas, and follow your planning. If possible revert back to working "all stations" as soon as possible.

7) Do not call by country. This inevitably leads to frustration. Why did he call for Holland and not for Belgium? Holland is only 20 km from here; why do they get a chance and not me?

8) On long haul paths on 160 meters skip is very often area selective and moving around. The secret to success on Top Band is to keep things simple. Simple instructions like, USA 5/10 UP or EU 7 UP are okay. More complicated instructions will inevitably lead to chaos on 160. However, do not just send UP. This will result in people calling less than 1 kHz from your frequency. Instead, specify QSX 5, or UP 5/10. If the pile is not too big, specify a single frequency (rather than a range) on which to listen.

9) On 160 meters always repeat the full call of the station you worked and confirm the contact, so he won't call you again for an insurance QSO. The same can be said of 80 meters if the signals are weak.

10) On 160 and 80 meters when paths are very marginal, send the call of the station you are replying to several times. After sending the report, send his call again and use a standard way of ending each QSO (TU, 73, your call etc). This is the signal for the crowd to start calling.

11) Once you return to a partial call with a report, stick to that station, and do not let him be overpowered by other callers. You're the boss on the frequency, show it. You decide who gets in the log, no one else. The pileup can be quite undisciplined, but often this is due to a lack of authority from the operator of the DX station. If the crowd notices that you stick to the original partial call, and that their out of turn calling is to no avail, they will eventually give up, and show more discipline.

12) The DX station operator should avoid frustrating his public. The only way to avoid frustrations is by meeting the expectations of the public (the DX chasers).

13) When working on 80 and especially 160, the operator should be armed with a good know how, know when and know

why (see Chapter 1). The DX station who works the really long haul stations on 80 and especially on 160 should continuously keep one eye on the gray line, and call especially for stations located near the eastern end of the path where the sun is about to rise. To distances greater than approximately 10,000 km the openings are most of the time limited to maximum 15 minutes right around sunrise with the station at the eastern end of the path. While the gray line is moving across Europe, the DX station in the Pacific should call CQ Europe only, and delay working US stations until after sunrise in Europe. Not doing so creates utter frustration in Europe. Unfortunately we see this happen again and again even with the best of DXpeditions.

14) The DX operator should never lose his temper. If the pileup remains undisciplined, he should not get too excited about it, but always stay cool. He should avoid sounding frustrated; to the contrary, he should inspire confidence.

15) It is important that the DX station operator emanates not temper but authority. All he must do is to firmly show the pileup that he is in charge, and that he sets the rules. If the situation does not improve, he can move to another mode or band, but should let the pileup know. A DX operator with a smooth pileup running has earned authority.

16) If the pileup grows too big, the DX station operator can eliminate those that copy you from those that "pretend" to copy you by suddenly changing the QSX frequency and then start working the ones who really are copying him! I saw BQ9P (October 2003) doing this with Europe on 160 and it was very effective and efficient.

17) Avoid copying half calls; this just slows down the QSO rate (especially on 160 where slow and deep fading is commonplace). On 160, when signals are weak, you need to copy the call in one single transmission.

18) From time to time ask if your frequency is still OK (especially if your rate suddenly drops). If your transmit frequency is being covered by QRM, ask your audience to look for a new transmit frequency for you.

19) If one particular station keeps calling out of turn, or gives his call endlessly just causing QRM, you may try to get rid of him by giving him an RS "00" report (phone) or an RST 009 (CW), by which you openly expose him as an *offender* (don't log the contact).

8.4. Pilot Stations and the Internet: Information Support for DXpeditions

After a first attempt during the AH1A expedition in 1993, the pilot station concept was introduced on a larger scale during the 3YØPI expedition in 1994. Mark, ON4WW, seems to have the honor of being the very first pilot station. (Pilot stations are operators not on the DXpedition who handle the information flow to and from the expedition.) Three years later the famous VKØIR expedition in January 1997 set the standard for how excellent logistics and a smoothly working pilot station can help a difficult DXpedition be a huge success. Both these expeditions were led by Bob, KK6EK.

In the past, DXpedition feedback and information had to be forwarded on-the-air during prime operation time. As a consequence information flow was minimal. Well-organized (and well-sponsored) expeditions nowadays use satellite telephone for communications with the home base, including the transfer of log data. The only limitation is that this kind of communication link is likely not to be continuously available for cost reasons.

The DXpedition pilot takes care of all the information flow to and from the DXpedition via one of these satellite links. He organizes himself to have a maximum of information from the "public" and to feed a maximum of information from the rare spot back to the public. He is the DXpedition's spokesman and public-relations man. His main tasks should be:

- Inform the world what the DXpedition operators hear on the low bands. What are the best propagation times? Are the announced frequencies OK, or have they been changed?
- Collect the desires, questions and remarks from DXers from all over, and convey a summary to the DXpedition operators.

These two tasks are there to optimize the results and to create confidence that all is being done to make the DXpedition successful. It also should prevent DX operators from asking questions themselves during valuable operating time or on the DX spotting networks.

In recent years, online logs have been made available by most DXpeditions. These logs are usually updated every day, so everyone can check whether he's OK in the log. This is important especially on the low bands, where due to noise, interference or fading, copy is not always perfect. This is important to keep stations from making a "backup" contact. I once missed a country (Malpelo) on 160 by not making a backup QSO, so I really cannot blame anyone for doing so if not 100% sure about the first try. Having the logs available on the Internet avoids this situation.

We should never forget that a DXpedition must be there for the DXers (the public) in the first place. To successfully add value to the DXpedition the pilots must be valued by the DXpedition leadership and must be fully integrated with the team. The pilots should be part of the decision-making process and not just the poor in-between guys who take a beating from both sides. Some examples of bad attitude and bad answers from a pilot are:

- "You are all complaining about the same thing." (In other words, leave me alone.)
- "I report what the operators tell me they will do." (Which means I'm just the in-between guy and I am far from sure that they will do what they say.)
- "We are making every effort to have a CW operator work from Eastern Europe across the Continent and across the United States on 80 CW at your sunrises." (This is an empty phrase with no message. A message with real content might be: "They will be on 3502, from 0300 to 0500Z on Feb 10.")
- "I have discussed the problem with the leader and while we are trying we cannot promise anything." (In other words, don't count on it.)
- "Alert: they will try 160 tonight. Time unknown." (There is no useful message here. We low banders, expect them on 160 and 80 every day anyhow!)
- "Golly, it's easier to criticize an operation than it is to put a rare one on the air." (A pilot must expect to receive criticism. That is part of his role. Criticism is only expressed when one thinks something is wrong. It's the role of the pilot to analyze the criticism, and to do something about it, to provide a solution, or at least an explanation.)

These are not fictitious situations. They were heard during a 2003 DXpedition that lead to great frustration for many low-band DXers.

Fig 2-16 — Tom Rauch, W8JI, probably has the best low-noise location in the southeastern US (south of Atlanta). Tom has been very helpful in the preparation of this new book (as well as with Fourth Edition). I consider Tom one of the most technically knowledgeable low-band addicts. His Web site, www.w8ji.com, contains a wealth of technical information, for which Tom kindly give me the authorization to use throughout this book.

9. CHASING AND WORKING THE RARE DX STATION

So far we have addressed the dos and the don'ts for the DXpedition operator and the operator of a rare DX station. But success in chasing the rare DX stations obviously also depends of the skills of the chaser.

CT1BOH not only wrote about the DX operator (see Section 8.3), but also about the DX hunter: "There is a price to be paid when a DXer tries to break through a pileup. That price is *time* and is totally dependent on two factors. The first one is his own skills as a hunter. The second one is the skills of the DX pileup operator. The not-so-skilled pileup operators *deserve* all the QRM they generate and get. It should be seen as an incentive to improve."

All of this is very true. Notice that when he talks about the DXer breaking through the pileup, he immediately brings the pileup operator (the operator of the DX station) back into the picture. What he says is true. In the first place it is always the DX operator's behavior and expertise that determines how smoothly and well-oiled the pileup is running.

But let us see what the DX hunter can and should do to break through the pileup and to keep the pileup running in a well-oiled and orderly fashion:

- Never call the DX station if you cannot copy him well enough. Well enough means you ought at least to be able to recognize your call if he comes back to you! Sometimes we see guys calling a DX station as soon as it's been announced on the DX spotting network, often without even having heard the station. Such a caller is just making a fool of himself.
- Make sure your station is properly adjusted before calling (is your transmitter set for split when calling in a split pileup?). Do not tune your transmitter on the frequency where the DX station is transmitting.
- Have you heard the instructions of the DX station? If not, wait and listen for instructions first! Make sure you know where he is listening. If you are not sure, do not call him on his transmit frequency!
- Never call before an ongoing QSO is completely finished. This means: no tail-ending (see Section 7).
- Correct timing is the key to success. In a pileup, do not start calling immediately, rather wait until most of the callers have stopped calling. This will increase your chances of getting through. This is not a competition where you need to be the first and fastest caller! What is important is to call at the right moment. Wait a number of seconds until the most excited callers have stopped calling and the QRM has died down somewhat before giving your call.
- If the DX station does not work split, you're in for a nerve-wracking session. Give your call two or three times, and then listen. If there are other stations still calling, don't start calling again immediately. Maybe the DX has already called you. Don't call endlessly! Throw in short calls every now and then. Stay relaxed, be patient and pray that eventually the DX station will go split.
- When the DX station is not working split and a large number of stations are calling, it may pay off to call just slightly higher or lower (maximum 100 Hz). See also Section 6. Do not forget however that timing is most important.
- If the DX station works split (thank heaven), first determine where he is listening. Listen in the pileup, and see what his operating strategy is. Does he stay on the same frequency? Does he move up or down a small amount, or is he really jumping around? Don't start calling him unless you know what he's doing. In such a pileup it's good to listen more to the calling "mob" than to the DX station!
- In a split pileup, give your call once, listen a second to see if the DX returns for anyone, if not, call again (one call) and keep repeating this cycle until the DX station replies to someone (hopefully you).
- On CW, do not call the weak DX station on 160 by sending DE YOURCALL YOURCALL K Maybe in QSB the DX station will just copy DE, and thinks it's a German station calling. Maybe he thinks the K is part of your call. Leave out the DE and the K, and send only what's essential: your call.
- When calling, always send your full call. No partial calls: It is not efficient and not a legal ID. Chances are that when the DX station comes back with ABC? you'll find there are a couple of "ABC" stations on the frequency. You are wasting everybody's time, including your own.
- If you hear the DX station coming back to you (you have heard all letters of your call), go back to him without repeating your own call, as this may make him think you are correcting the call he copied. Just say "QSL, you are 59" or in CW, CFM UR 599 or TU UR 559 or QSL UR 539. Avoid confusion.
- If someone repeatedly calls the DX station on his transmit frequency, tell him in a friendly way he's on the wrong VFO. Don't say or send "up stupid," or "up up up" or "idiot." We all make mistakes and don't like to be called idiots. Instead, say "XYZ up please" or "XYZ up" (assuming XYZ is the suffix of that station's call). You must identify the station; how else could he know your message is addressed to him? The low bands, and 160 meters in particular, are called "gentleman's bands." Act like a gentleman!

• If the DX station answers another station, keep quiet. The more stations that stop calling at this time, the faster the QSO will be over and the faster it will be your turn.
• If possible use full break-in (QSK) on CW. It will allow you to hear the DX station the second it comes back, even in the middle of your transmission.
• If the DX station returns with an error in your call sign, repeat a few times that part of your call where the error occurred. Ask the DX station to confirm your call.

10. CALLING CQ ON A SEEMINGLY DEAD BAND

We frequently hear that every wise DXer spends all his time listening, and only transmits when he's sure to make a contact. He never calls CQ DX; he just listens all the time and grabs the DX before someone else does. This rule for sure applies to the DXer, to the "hunter."

However, this rule does *not* apply to the DXpedition (the "hunted"). If the golden rule for a DXer is to "listen, listen, listen," then the golden rule for the DXpedition should be "call, call and keep calling CQ!" And please, don't give up after just a few minutes. DXpeditions should call CQ, even on a seemingly dead band, at the times they publish. You can be assured that there are hundreds of faithful low-band DXers digging for your signal.

And don't go away after just one or two contacts, even if there are no replies for a while. You probably will be announced on the DX spotting network, but it takes some time before the news gets out.

The ZL7DK guys said it so well: "During our stay we got at least one good opening in all possible directions, but on average not more than two per destination. The openings in the critical directions (mostly the polar paths) have to have absolute priority. The paths are open maybe 5 or 10 minutes a day, if they are open. If you are dedicated to working stations on these difficult bands and difficult paths, you must be there every day (to call CQ) to not miss any opening."

11. CW ON THE NOISY LOW BANDS

There's no doubt about it: CW is superior to phone when it comes to making a QSO under marginal conditions. CW can use a much narrower bandwidth, which means a better signal-to-noise ratio. I typically use a 100 to 200-Hz bandwidth on CW, versus 2.1 kHz on SSB, so the advantage is obvious. That's why Top Band DXers spend 92% of their time on CW. On 80 meters the figure is 78% (see Section 20.6).

What about PSK31? It is a fact that a well-trained CW operator can copy weaker signals in low-band noise much better than PSK31 does. This is because the decoder (the operator's brain) is vastly superior to the PSK demodulator/software. But I must admit, PSK31 comes close. There are of course other "very slow" digital modes such as WSJT (by K1JT) that can copy signals that are buried very deep in the noise, but these are not considered here. In the world of DXing we only consider modes that allow QSOs to be made at a "normal" rate.

One of the situations that makes copying signals very difficult is QRN. It appears there are two families of QRN: high-latitude QRN and tropical QRN. The difference is that crashes of tropical QRN generally last much longer than those generated by high-latitude QRN. With higher-latitude QRN the pauses between crashes usually last longer.

Fig 2-17 — Mark, ON4WW, operated from a lot of countries on the low bands during his career with the UN. With over 275 countries confirmed on 160 meters, Mark is now working his way up to the top ranks.

If you want your call to make it through high-latitude QRN, high-speed CW can sometimes be a solution. Dan, K8RN, who operated VK9LX on Top Band said: "...QRN was very bad even with Beverages for receiving. It seemed to me that if the stations calling sent their call fast, they had a better chance of making it through (between) the static crashes. If the speed was too low nothing made it through."

But high-speed CW is no good at all to pierce through tropical QRN. Rolf, SM5MX and XV7SW, recently commented on the Top Band reflector: "In this kind of tropical QRN, each QRN bang often lasts long enough to mask a call sign completely. From the DX end you may just understand that somebody is there and call QRZ?, but the same thing will happen again at the next bang, the next one — and the next and so on, if the speed is too high. So I found it tremendously helpful when people reduced the speed. Once you are able to pick out a letter here and there, you may be able to paste together a full call sign and eventually make it."

Referring to another issue regarding high-speed CW on the low bands, Tom, N4KG, commented "High-speed CW on the low bands by DX stations contributes to confusion and disorderly conduct in the pile-ups. Half of the callers can't copy anything but their own call signs, even with a good signal on a quiet band."

The DX station should determine the CW speed. His sending speed should be the speed he expects replying stations to use. Tom, N4KG, added: "DX stations sending above 30 WPM on the low bands actually reduce their rate and promote more broken calls. 25 to 28 WPM seems to work well for most cases. On long polar routes, with weak signals, QSB, and QRN, high speed is counterproductive. Sending a call twice at 25 WPM takes less time than three times at 30 WPM and is more readily copied."

Joerg, YB1AQS, formulated it as follows: "Even if you can hear everybody crystal clear — don't shoot at them in CW with 35 WPM! 22 WPM on 160 meters and 28 WPM on 80 meters are enough. Repeat their call sign two times before the report and at least once at the end."

Chris, ZS6EZ ,made an important remark along the

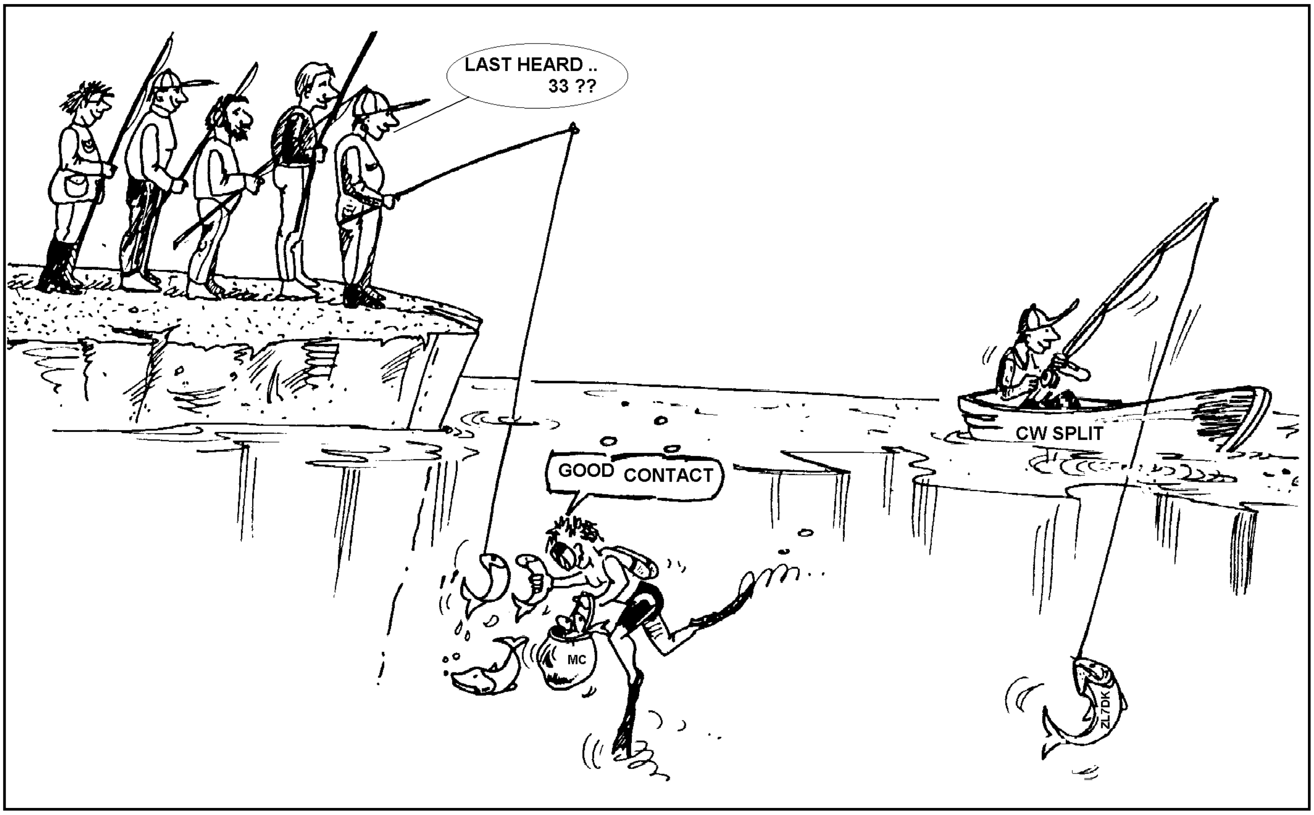


Fig 2-18 — List operations. Sigh…

same lines: "Never, never, *never* screw up the spacing in your call. If you use standard Morse spacing, the receiving station can often recover dits that are inaudible, by listening to the timing of the characters. For some reason, some people think the call is easier to copy if they leave exaggerated spaces between letters." Well it simply does not work that way. The rhythm is very important!

12. NETS AND LIST OPERATIONS

The use of lists, which occur daily on the HF bands, started with net operations on the HF bands in the 1960s. In most of these nets, a "master of ceremonies" (MC) will check in both the DX and the non-DX stations, usually by area. After completing the check-in procedure, the MC directs the non-DX stations, one at a time, to call and work the DX station. In most cases the non-DX station has indeed worked the DX station, but there was no competition, no challenge, no know-how involved. Some write the MC a letter, or send him an e-mail message or even call him on the telephone to get on his list!

What satisfaction can you derive from such a QSO? Yes, it gives the QRP operator a better chance to work the DX station, and the only thing you have to do is copy your report — and even that may be relayed to you. When it's your turn, the MC will call you and invite you to make a call. It's just like shooting fish in a barrel, in my opinion. Fortunately, lists have never made it on CW.

A list cannot be used if the DX station refuses to take part. Fortunately, we rarely see a DXpedition worthy of the name doing this. I remember hearing stations asking Carl, WB4ZNH, operating as 3C1BG on 80 meters, if they could run a list for him. Carl was insulted by the proposition.

If a DX station is involved in a list operation, it generally means he cannot cope with the situation. The ability to cope with a pileup is part of the game for rare DX stations.

In all cases, list operation can be avoided by working split frequency. Operating from lists always is a poor solution. There will always be a number of poor operators, as well as newcomers who think list operation is an alternative to working a pileup. None of the serious DXers or DX stations get involved in this list game.

13. ARRANGING SKEDS FOR THE LOW BANDS

Once you work your way up the DXCC ladder, you will inevitably come to a point where you will start asking stations on the higher bands for skeds on the low bands. You will often be asked to specify the best time for the schedule. Remember that you are asking for a favor, so try a time that is not in the middle of his night. Rather, get yourself up in the middle of the night! Also, don't go by a single schedule. Arrange a minimum of three skeds, or maybe a week's skeds, to hit the day with the right propagation. Tell the other party that the band may be okay only one day out of three or one day out of five. Find out how much power he runs on 160 and what antenna he is using, so that you know what signal to expect. Fortunately nowadays most DXers are quite knowledgeable about low band operating. We've come a long way in this respect since 35 years ago. I remember in 1974 I worked A51PN on 15 meters at 1300 UTC (the middle of my day) and asked him about 80 meters. He said he indeed had a dipole, and proposed to move to 80 right away… at my local high noon. I explained and the same night I did work my first A5 station on 80 meters.

Don't forget to have your sunrise and sunset information ready at all times. Most computer logging programs nowadays include it.

Tell your sked that you will call him. Don't give his full call; just give his suffix when you call, or even better, just send your call. Or just call CQ DX at the sked time exactly on the agreed frequency. Spot him after you make your QSO.

14. GETTING THE RARE ONES

Working the first 100 countries on 80 or 40 meters is fairly easy. Well-equipped stations have done it in one contest weekend. It's even been done in a single weekend on 160 meters! Anyone with a good station should be able to do it easily within a year and a growing number of stations have achieved DXCC on 160 meters. The major DX contests (CQ World Wide DX, ARRL International DX, Worked All Europe, All Asia, CQ World Wide 160 Meter, ARRL 160 Meter, Stew Perry Top Band Distance Challenge, etc) are excellent opportunities to increase low-band scores. A good time to look for semi-rare ones, by the way, is just before and after a contest, since that may actually be an easier time to work them due to less QRM.

14.1. Getting Information, Getting It All and Fast

Twenty years ago, if we wanted to get information on whatever subject, we had to consult books or periodicals. Maybe go to the library. We had to spend a lot of money for good technical books. Nowadays an abundance of good information on all kinds of subjects related to our hobby is available — for free — on the Internet. The Internet contains an almost unlimited wealth of valuable information, but unfortunately also a bunch of crap. The art is to be able to distinguish one from the other.

What we should not forget is that other people have made the information available, and recognize them for that. In this respect I would like to especially to pay tribute to our fellow hams who create Web sites full of up-to-date, very valuable information regarding all kind of aspects of Amateur Radio, including the art of DXing on the low bands. The site of my friend Tom Rauch, W8JI, is one such example (**www.w8ji.com**). Thank you guys!

14.2. DX Clusters (DX Spotting Networks)

In the old days we had lots of DX bulletins and local DX information nets, all over the world, to inform us.

Some 20 years ago, these information nets started to be replaced by local DX Clusters working on 2 meters or 70 centimeters. Information was much "fresher," and within a day or so a message sent from the US would arrive in Europe. In the mid 90s these DX Clusters started to be networked via Amateur Radio (UHF and microwave links), and later via the Internet. Today most DX spotting networks are accessed via the Internet (for a list with data for such DX spotting networks see **www.ng3k.com/Misc/cluster.html** or **ve9dx.weblink.nbtel.net/telnet/sites.html**).

The Internet now also links most if not all DX spotting networks worldwide. Our information systems are now global and all kinds of information is instantly available. DX spotting networks have completely replaced the DX information nets from yesteryear. Today almost every DXer is familiar with the DX Cluster. Wideband Internet has become commonplace in all industrialized countries, which means most of us can have instantly (within seconds) access to the DX information from all over the world.

It is true that the Internet and DX spotting networks have changed DXing in general. Some publications have pictured DX Clusters as the greatest evil in Amateur Radio. They are said to undermine the art of listening. Scott, W4PA, has a very strong view about this issue: "Shut off packet radio, and do it like a man." As a matter of fact, in just about all contests there is an "unassisted" category where one is supposed to have shut off the DX Cluster during the contest. We know also that unfortunately quite a bit of "cheating" is going on in this regard as some unassisted operators use DX Cluster spots.

Others advocate that we should be allowed to use all technical means (eg *CW Skimmer*, see Chapter 3, Section 1.7) at our disposal for DXing and contesting. These advocates consider all of this a superb driving force to further develop new technologies, new software and new gadgets in Amateur Radio. Technological aids to help the operator are one thing; computers taking over from the operator is another thing. Amateur Radio is a hobby for people. In my humble opinion people should be in the focus, not computers. QSOs made by computers, without a human being in the leading role, are not QSOs by the wildest imagination. Such achievements should never count as contacts for contests or in the framework of DX awards (DXCC etc).

Fact is that DX Clusters and a range of technological advancements are here to stay. We will all have to develop different skills that help us keep a technological advantage over competitive fellow DXers in this ever faster changing world. I am personally convinced that the DX Clusters and the reflectors and Web pages on the Internet are just a set of superb tools that have evolved in our wonderful hobby.

14.2.1. What Do You Do on a DX Cluster?

- You can use information from a DX spot to make DX contacts. A few guidelines: If a new DX station or new country is spotted, stay calm and do not start calling the DX station blindly. If you do not copy the station, just relax, maybe the skip will change in your favor. Make sure you copy the station well enough to verify if the spotted call sign is correct. Don't call if you cannot hear the DX station giving its own call. Listen for the DX station's instructions before calling (his listening frequency, is he working *everybody* or working by numbers or by geographical areas?).
- The spots come automatically on your screen in chronological order. But you can also retrieve old spots by band or by call sign or by combination of band and call sign.
- You can look up WWV information and Solar Flux Index.
- You can also find QSL information on some DX Clusters. If this function does not exist, retrieve the last 25 spots for that station, and chances are that one of the spots has the QSL info in the comment field.
- Who do you spot? Only rare DX stations that are of interest to DX chasers. Before spotting a station, first check if anyone else has just spotted that same call. Watch out for typos! Wrong calls can sometimes be found in logs because

the operator worked a station without even having heard a call sign, blindly having copied a busted (incorrect) call from the DX Cluster.

- Besides spotting you can also share valuable and interesting DX information with the DX community, using the ANNOUNCE/FULL (to all) function. This function should only be used to announce information that is useful for a great majority of DXers connected to the network. Example: You could announce that the DXpedition has just moved band, or frequency, or that they will be on such and such a frequency at such and such a time.

Great system, I bet you have worked new countries (band countries or mode countries) thanks to the DX Cluster system. Thanks to the guys that spotted the DX station for you. The value of the DX spotting networks would be nil if DXers did not spot interesting DX stations. It is a system of giving (spotting) and taking (using spots).

14.2.2. What Should You Not Do on a DX Cluster?

1) **Self spotting.** What's that? It's a personal advertisement to the whole world, saying: *Here I am, on this frequency, please call me*. If you want to make QSOs, call CQ or reply to stations calling CQ. Also, self spotting leads to disqualification in contests.

2) **Ego boosting**: A spot is not for telling the world how great you are. Don't spot a DX station (that's been spotted several times anyhow) with a remark: *I finally did it…*. In such a case you are not announcing the DX station, you're just bragging and telling the world how great you are.

3) **Disguised self spotting**: An example: You work a nice DX station that came back to your CQ. When you finish your QSO you spot the call of the DX station, which was there but went off the frequency after finishing the contact. This spot has zero added value for the DX community, as the DX station is gone, but at the same time you attract a bunch of DXers to your frequency, hoping that this will help you work some other DX stations.

4) **Spotting a friend**: A good friend of yours is calling CQ repeatedly, without reply. You want to give him a little push and you spot him, though he is not a DX station. Neither your friend nor you will gain respect in the eyes of the ham community by doing so.

5) **Asking a friend to spot you**: That is self spotting, using a cover up. Self spotting is not done, so do not ask your buddy to spot you.

6) **Playing cheerleader:** The cheerleaders are those who continuously spot their favorite contest station during a contest. It's like the supporters pushing bike racers during a race in the mountains. It isn't fair and it's unsportsmanlike.

7) **Sending a spot which really is a private message**: We need to realize that each spot, each message is sent to many thousand of hams all around the world. A spot from a European station saying *VK3IO on 1827*, with a comment *QRV???*, obviously is not a spot but a private message to VK3IO. What if everybody did things like this?

8) **Using the DX Cluster as a chat channel**:

- With the TALK function you can send individual messages to another ham on your local DX Cluster. Some DX Clusters have a similar function where you can chat privately with a user on another DX Cluster, of course provided these clusters are linked (by a radio link or internet). No problem if you want to use it; you are not disturbing any other users of the DX Cluster system.
- The ANNOUNCE/FULL (to all) function is a totally different story. Any message sent using this function will be sent to the users of all world wide linked clusters, and that may be many thousands of users at any given time. Be very careful when using this function. It seems however that most ANNOUNCE/FULL announcements are in fact intended for one particular person, where 9,999 others are forced to read a message which is of no value to them. Sending such private messages via the ANNOUNCE/FULL function is very selfish and considered cluster littering. Do not ever use this function as a (private) chat channel.

14.3. Internet Chat Channels (Chat Rooms, DX Boards)

While DX Clusters are not meant to be worldwide chat channels, such chat channels for the DXers are available nowadays and are the most real-time gadgets around. I learned about the ON4KST Chat Room when writing the Fourth Edition of this book, back in 2003. By now (2009) I had expected to see more such systems pop up in the following years, but that did not happen (or I am unaware of such systems).

But for those not acquainted with the system, what is a chat room? A chat room is a term used to describe any form of synchronous (real time) conferencing — in words, it is an online chat system. The ON4KST chat room can be accessed on his Web site (**www.on4kst.com/chat/**) and is designed for low band, VHF, UHF and microwave chats.

The screen of the ON4KST Chat Room for the low bands has three windows: the main window with chat text, a window showing the users that are logged on, and a third window showing you the latest spots for the low bands (40-80-160) collected from a large number of DX Clusters worldwide.

Many, if not most of the active low band DXers (mainly 160 meters) have been using the ON4KST chat room at one time or another. It is an interesting tool where you can talk to other low-band DXers provided they're checked in and available for a chat. The nice thing about it is that you certainly don't disturb the other party. By being online in the chat room, he has made himself available for a chat.

While such a chat room for low band DXing is undoubtedly interesting and useful, it could also be a dangerous tool. Even before such Internet chat channels came into existence we saw some would-be DXers using the DX Cluster to send messages like: "I am calling you on xyz kHz. Do you hear me?" Worse even and utterly unethical and unfair is "You are 339 did you copy my report?". Let's make sure we all use such chat channels correctly. Then we will have another technological tool with which we can responsibly enjoy our hobby.

It is clear in my mind that the use of a chat room such as the one run by ON4KST, should under no circumstance be used during a contest. Not even for saying hello and goodbye to a friend. In essence, using the chat room is the same as picking up the phone. And telephone contacts to talk to other stations are not allowed in contests, and if it's not in the rules, it should be considered a rule of ethics.

14.4. Internet Reflectors, DX Magazines and Other Sources

While many years ago, printed DX magazines served the noble purpose of informing the DXers of upcoming activities, this role is now taken over by DX Clusters and Internet reflectors.

Several special-interest groups on the Internet deal with low-band DXing. These interest groups use so-called *reflectors* to exchange information among their subscribers. Reflectors are semi-open mailboxes, to which anyone can subscribe, free of charge. Once subscribed, you will get copies of all the mail that is being sent to this reflector. By addressing mail to the reflector, you reach everyone who is currently subscribed to that particular reflector.

The Top Band Reflector (**lists.contesting.com/mailman/listinfo/topband**) is the place to be for all Top Band related information.

Other popular Internet sources for DX and contest information (related to any HF band) are:

- The DX Reflector (**www.njdxa.org/faq/faq-dx-news.php**).
- If you are into contesting, the Contest Reflector (see **www.contesting.com/FAQ/cq-contest**) is a very good source of information.
- Information on planned, current and past DXpeditions, can be found on NG3K's Web site at **www.ng3k.com**.

14.5. DX Bulletins

DX bulletins on the Internet have replaced paper DX bulletins. Most of these are weekly publications.

- Probably the most popular DX information sheet is the weekly *425 DX News* (**www.425dxn.org**), a no-charge Italian weekly DX bulletin with more than 11,000 subscribers worldwide. The editors of *425 DX News*, Mauro and Valeria Pregliasco (I1JQJ and IK1ADH), were inducted into the CQ DX Hall of Fame in 2007, as a recognition of the outstanding work they have done for the DX community over the year.
- Ted, KB8NW, edits the Internet edition of the *OP-DX* B*ulletin*, which is also a weekly DX bulletin: **www.papays.com/opdx.html**. It is also available on the Internet (free of charge).
- Bernie, W3UR (**www.dailydx.com**), publishes *The Daily DX* Monday through Friday. *The Daily DX* is available daily as an e-mail containing a collection of all the latest DX news.
- Carl, N4AA, publishes *QRX DX* in both paper and email formats (**www.dxpub.com**).
- The ARRL also publishes weekly the W1AW DX bulletin (**www.arrl.org**).

A comprehensive list of available DX newsletters can be found on **www.dxzone.com/catalog/DX_Resources/Newsletters/**.

15. ACHIEVEMENT AWARDS

There are a number of low-band-only DX awards. The IARU issues 160 and 80 meter WAC (Worked All Continents) awards. These are available through IARU societies including ARRL (225 Main Street, Newington, CT 06111, USA). ARRL also issues separate DXCC awards for 160, 80 and 40 meters. More information on these awards can be found at the ARRL Web site at: **www.arrl.org/awards.**

CQ magazine issues single-band WAZ (Worked All Zones) awards. See: **www.cq-amateur-radio.com/awards.html**. Rules, application forms and other useful information are available on the Web site.

In addition, there are very challenging 5-band awards: 5-Band WAS (Worked All States), 5-Band DXCC (worked 100 countries on each of 5 bands), both issued by ARRL, and 5-Band WAZ (worked all 40 CQ zones on each of the 5 bands, 10 through 80 meters), issued by *CQ* Magazine.

There are also the achievement awards issued by the sponsors of the major DX contests that have single-band categories which are also highly valued by low-band DX enthusiasts. The major contests of specific interest to low-band DXers are:

- Stew Perry Top Band Distance Challenge (last weekend of December, CW only): **jzap.com/k7rat/stew.html**
- CQ World Wide 160-Meter Contest (CW, usually last weekend of January and phone, usually the last weekend of February): **www.cq160.com**
- ARRL 160-Meter Contest (first weekend of December, CW only): **www.arrl.org/contests**
- CQ World Wide DX Contest (phone, last weekend of October and CW, last weekend of November): **www.cqww.com**
- ARRL International DX Contest (CW, third weekend of February and phone, first weekend of March): **www.arrl.org/contests**

Continental and world records are being broken regularly, depending on sunspots and improvements in antennas, operating techniques and participation.

Collecting awards is not necessarily an essential part of low-band DXing. However, collecting the QSL cards for new countries is essential, at least if you want to claim them. Unfortunately, there are too many bootleggers on the bands, and too many unconfirmed exchanges that optimists would like to count as QSOs. These factors have made written confirmation essential unless, of course, the operator never wishes to claim country or zone totals at all. Many other achievements can be the result of a goal that each of us sets out individually.

The ultimate low-band DXing achievement would be to work all countries on the low bands. This goal is not very difficult for a well equipped station on 40 meters providing all countries are available. It is possible on 80 (although P5 has never been activated here). However, it is almost impossible on 160 meters, although we see the 160-meter scores slowly climbing steadily and now reaching over the 320 mark.

16. STANDINGS ON TOP BAND

Ten years ago, when I wrote the Third Edition of this book, the big question was: "Who will be the first to get 300 countries on Top Band? In June 2010, 19 stations already have met or passed that magic number and Bill, W4ZV, now leads the pack with 327 countries! As of June 2010, the next four stations were K1ZM, 323; VE1ZZ, 321; W4DR, 319 and W8LRL, 318.

In November 1976, Wal, W8LRL, sent in 100 QSLs to DXCC and obtained award #3. W1BB and W1HGT beat him to #1, as they hand carried their cards to the ARRL! In 1986 Wal updated his score to 201 countries, and he made a third update in September, 2002, when he hand-carried 108 very valuable cards to ARRL HQ for 309 confirmed countries on 160. From there on Wal apparently started updating his score one by one...

As far as I know, today only two countries have never been available on 160 meters: 7O and P5. On 80 meters only P5 has never been made available so far.

You can check your DXCC ranking in the various DXCC listings available on the ARRL Web site (**www.arrl.org/dxcc**). These ranking are now updated daily.

Yuri, K3BU, makes available a listing of the major 160-meter contest records on: **www.k3bu.us/topband_records.htm.**

For many years a small number of North American stations (W4ZV, W8LRL, K1ZM, W4DR, VE1ZZ, K5UR and N4WW) have been leading the DXCC listings on 160 meters. In the past we have sometimes seen a few unfamiliar calls appear out of the blue, but they never remained in these top rankings after a thorough checking of their "credentials." I must say that the DXCC committee has been doing a difficult but very good job in keeping cheaters out of the DXCC.

In the Fourth Edition of this book, for the first time I did an analysis to find out why North American East Coast stations seem to be able to work 300 countries on 160 meters more easily than European stations, and the European stations more easily than Japanese stations. In preparation of this edition, I did the same analysis in 2008. In both cases I analyzed top 160-meter scores from the US East Coast (the combined countries worked by K1ZM, W4DR and W8LRL, a circle of approximately 250 km radius on the US East Coast), from Western Europe (my score) and from Japan (countries worked by all top 160-meter DXers combined — information obtained from JA7AO).

According to the late 2009 DXCC List (338 countries available) there are only 16 countries that have never been worked on Top Band from the US East Coast, 32 were never worked by ON4UN and just under 50 are missing from Japan.

I plotted these missing countries on three different great-circle maps, centered on Washington, Belgium and Tokyo, and showing the size of the auroral doughnut with low moderate aurora activity (k=3). See **Figs 2-19**, **2-20** and **2-21**.

For the US East Coast I found 34 DXCC countries hidden behind the auroral doughnut, of which they together need only 8 or 24% , in addition to 7 "easy" ones. "Easy ones" means countries that are relatively easy to contact from a propagation point of view.

In Europe I spotted 36 DXCC countries hidden behind the auroral wall, of which 25 or 70% are still needed. The difference between this case and the US East Coast case is very significant (70% vs 26%). Outside the Pacific black hole there remain only 7 "rather easy" to reach ones.

As for our JA friends, they have 51 countries hidden behind the auroral doughnut, of which they need 27 or 53% (five years earlier 37 or 73%) in addition to 23 "easy" ones (five years earlier, 29 "easy" ones). Note that a large number of the missing countries in the Caribbean area are right on the borderline between "behind" and "not behind" the aurora circle. For this exercise they were counted as being "behind" the aurora doughnut.

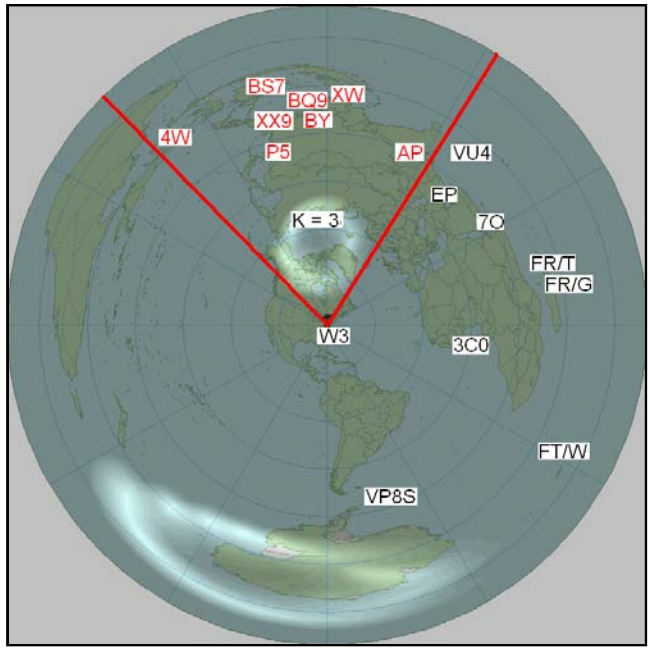


Fig 2-19 — Great-circle map (K=3) centered on Washington, DC, showing the countries needed on 160 meters by K1ZM, W4DR and W8LRL (as of February 2008). (Map generated by *DX Atlas*, with additions by ON4UN)

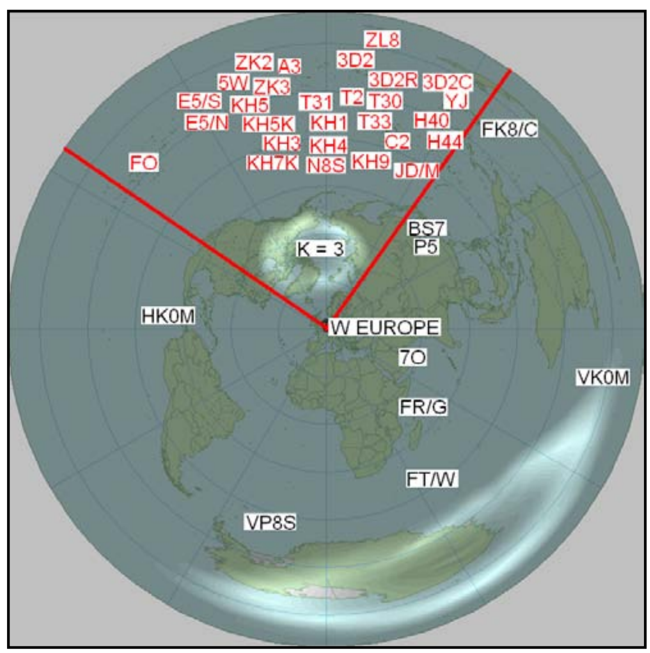


Fig 2-20 — Great-circle map (K=3) centered on Belgium, showing all the countries needed (October 2009) by ON4UN on 160 meters. Note that the large majority of needed countries are in the Pacific, behind the auroral oval (see Chapter 1, Section 3.2.4). (Map generated by *DX Atlas*, with additions by ON4UN)

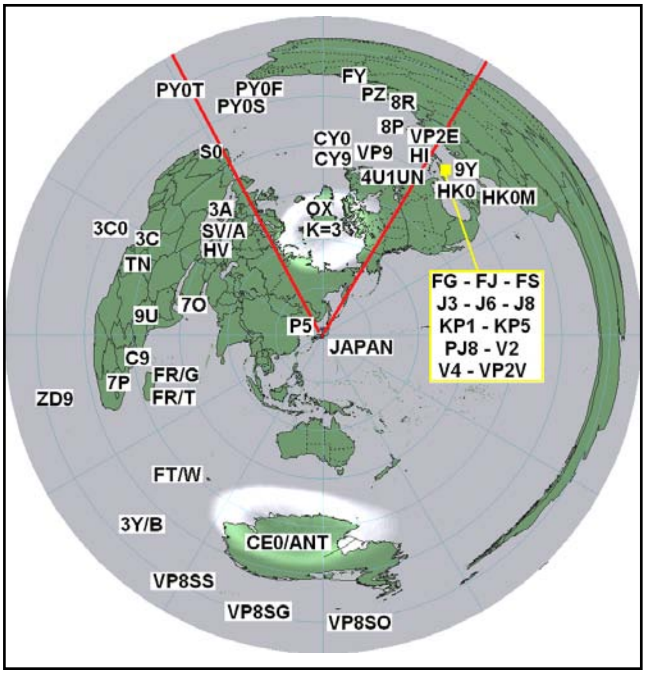


Fig 2-21 — Great-circle map (K=3) centered on Tokyo, showing the countries that have not been worked by any JA station on 160 meters (source JA7AO, May 2008). (Map generated by *DX Atlas*, with additions by ON4UN)

What can explain these differences in countries actually worked?

- The angle under which the aurora circle as seen from Belgium is approximately 100 degrees. In the USA East Coast case (centered on Washington DC) this angle is about 75° (Belgium is 51° N, same latitude as Calgary in Canada!).
- The European black hole (central Pacific) contains 12% more DXCC countries than the US black hole (East and South-East Asia).
- For Europe, the missing countries hidden behind the aurora oval are, without exception, small islands that are only activated now and then by DXpeditions (and very often at the wrong time of the year to make Top Band contacts possible).
- The countries in the European black hole are at distances of between 12,000 and 18,000 km (calculated average distance: 14,450 km). For the USA black hole this distance ranges between 11,000 and 16,000 km (average 13,200 km).
- For our JA friends a large majority of the missing countries behind the aurora oval are also at a distance of between 10,000 and 15,000 km, similar to the US East Coast case.
- JAs have had a handicap (until about 10 years ago) of a small allocation around 1900 kHz.
- The JA data were collected by JA7AO in early 2008, and combine countries from 26 JA stations (in order of countries worked at that time: JA1JRK, JA4LXY, JA3ONB, JA4LKB, JA7NI, JA2PJC, JA2FXK, JA1HQT, JA1GV, JH4UYB, JHØBBE, JA7OEM, JA3FYC, JA7AO, JA1CGM, JA4DND, JA8ISU, JAØDAI, JA3EMU, JA4CUU, JR7VHZ, JA6GCE, JA7FUJ, JA1EOD, JA3CJO, JA6LCJ, JA1DDH, JA6JPS and JA6GIG). More than 100 JA stations have worked DXCC on 160 meters, and more than 20 have over 200 countries!
- The Europe tally (my score) was set over a period of "only" 20 years.
- The US East Coast tally is from three stations spread over approximately 500 km of the coast, and one of them has been active during at least one more sunspot cycle than I have.
- Perhaps US Top Banders just spend more time on their radios, rather than writing books?

To me it seems that the slightly larger number of hidden countries and the nature of these countries (remote islands with little or no permanent ham population), together with a slightly greater distance to these targets, make the Western Europe case the toughest one! What is important is that we try to understand why our US East Coast friends do systematically better than what apparently can be done from Western Europe. Western Europe is a great place to work the US on 160, but not for working DXCC!

Interestingly enough, 160 meter WAZ is easier from Europe than USA or JA because of relatively high activity in Europe's toughest zones: Zone 1 (KL7) and Zone 31 (KH6). Japan's toughest is Zone 2 and the US East Coast's toughest is Zone 23 (JT and UAØY), both of which have relatively low activity and are directly behind the Magnetic North Pole.

We have seen in Chapter 1 that aurora is a major limiting factor for working DX on Top Band. The fact that we do work countries in the "black holes" beyond the aurora doughnut, and not only during the sunspot minima, merely means that sometimes we can get through, but these openings are rare indeed. Luck and patience will help you hit the right opening. This again emphasizes that DXpeditions should be on Top Band from the first day until the last day.

Sustained periods of propagation through the auroral zones are very exceptional on 160 meters. During the period mid-September 2008 to February 2009 we witnessed such extraordinary propagation. On many days we could work KH6 and KL7 from Western Europe (KL7 both in the morning and in the evening), and Scandinavian stations had a ball working the Pacific island stations right across the polar regions. But these are exceptions that we witness maybe once every 10 or 15 years! During that period we witnessed a fabulous opening between Europe and the West Coast on Sunday morning (GMT) during the 2009 CQ World Wide 160 Meter CW Contest, when large numbers of W6 and W7 stations were worked with fabulous signal strengths, up to one hour after sunrise in Europe.

Outside such enhanced periods, how do we work the rare countries behind the aurora wall? Under such "average" conditions, shooting through the aurora belt on short path is an almost impossible task most of the time, although this has become more routine in the recent solar minimum of 2007-2009. At best one can hope to shoot "around" the doughnut, but to be able to do this the ionosphere has to cooperate a great deal and create the right ionization gradients along the path (see Chapter 1, Section 4.3), which certainly does not happen frequently. From the US East Coast, countries such as JT, 9V, 9M2, BY, or UAØ are all right across the North Pole aurora doughnut (see **Fig 2-22**).

W4ZV reports that the best season for working such stations is from mid-December through mid-February. The opening from the US East Coast to places like 9M2 usually lasts only minutes. At that time propagation is coming from the southwest and travels all across the Pacific where it is dark.

Such a path was used in early February 2003 when a number of US East Coast stations worked JT1CO for their last zone on 160 meters. W4ZV and his friends VE1ZZ, W1JZ, K9HMB, K3UL, K9RJ, K1UO, K1ZM and W1FV had been trying for a long time, and they hit the right day. See **Fig 2-23**. For more detail on the mechanism behind such crooked paths, see Chapter 1, Section 4.3.1.1.

Bill, W4ZV, also remarked that these kinds of QSOs do not seem to happen frequently during sunspot minima, but rather during or near sunspot maxima. This may be an indication that a certain degree of ionization is critical to set up the required ionization gradient responsible for bending the signal path, which probably happens in a zone near or in the terminator.

Fig 2-24 shows a simulation for a crooked propagation path between Western Europe and Europe's black hole (centered on KH8) in the Pacific. The regular path between ON and KH8 is almost right across the North Pole. At both ends of the circuit the stations are *not* beaming in the genuine great circle direction (which is North). On their way the signals, following a straight line propagation path, encounter an area where a steep horizontal ionization gradient in the ionosphere bends the signals and acts as a kind of reflector. The exact place where the bending occurs can sometimes be reconstructed using ionospheric data (see Chapter 1). The areas of reflection shown on both maps are just the result of a simulation and not to be considered as the exact points where this always happens. I know it's still very controversial exactly how "long path" really works, but the mechanism as explained above has been confirmed on several occasions where the ionospheric data were analyzed and found to prove the existence of such areas where the required steep ionization gradient could have been responsible for the bending phenomenon.

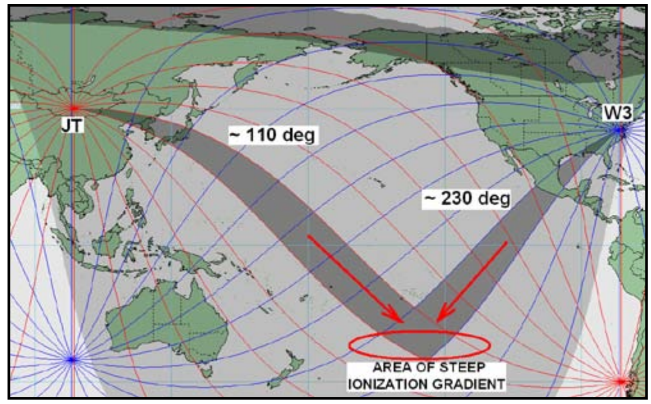


Fig 2-23 — This map shows the great circle lines centered on JT (Mongolia) and W3. From the US East Coast, stations were receiving the signals from ~230°, a path going just south of FO (French Polynesia). It is likely that in that area of the Pacific a steep horizontal ionization gradient caused the wave to be bent in the direction of JT. Beaming from JT the signal heading would have been slightly south of east. (Map generated by *DX Atlas*, with additions by ON4UN)

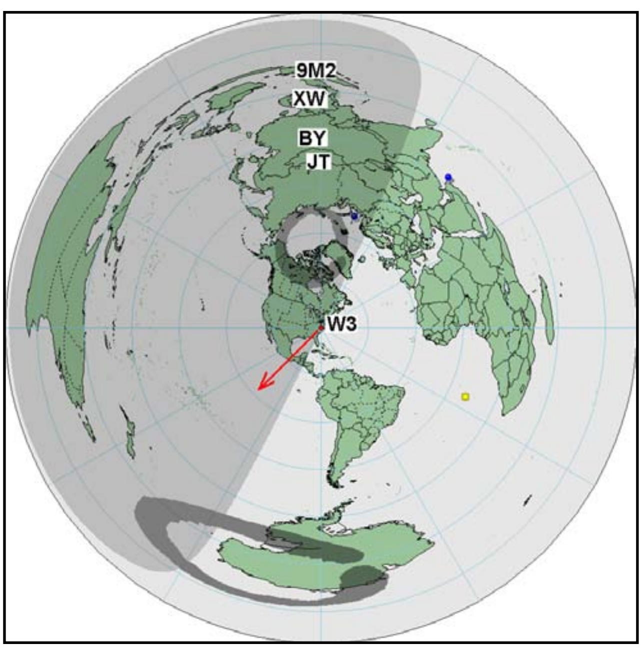


Fig 2-22 — 160 meter bent long path propagation from US East Coast to the mid and east Asia black hole around January 15, at 1220 UTC. (Map generated by *DX Atlas*, with additions by ON4UN)

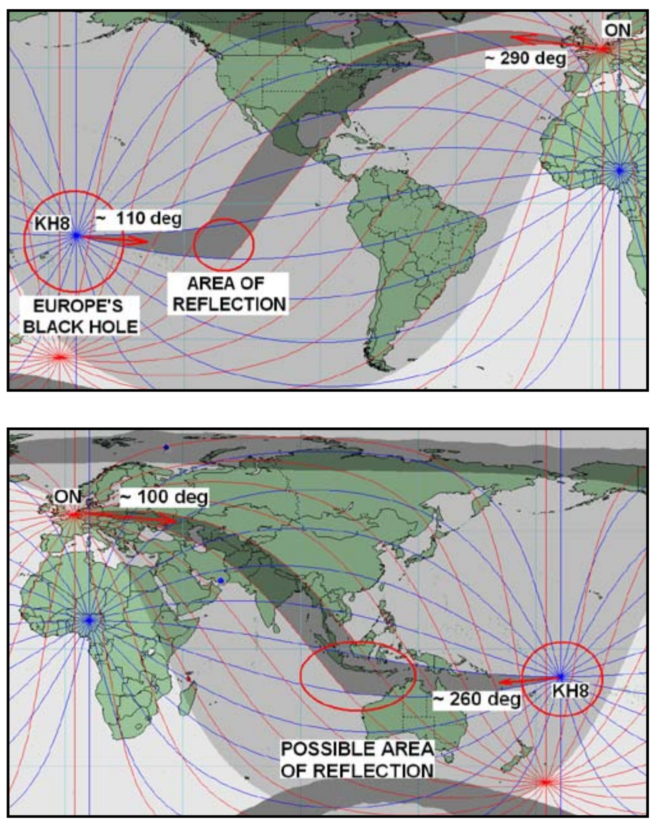


Fig 2-24 — Simulation of possible sunrise (top) and sunset (bottom) "crooked" propagation paths between Western Europe and the Central-South-Western Pacific (Europe's black hole on 160 meters). The maps were drawn for mid-winter at sunrise and sunset in Belgium. (Map generated by *DX Atlas*, with additions by ON4UN)

17. DXING AND CONTESTING ON THE LOW BANDS

One often hears "contesting and DXing do not go together." I think this is certainly much more true on the high bands than on 80 and especially 160 meters. Why? The high bands have reasonably predictable and stable propagation conditions. On the high bands a DX chaser is after difficult countries, uninhabited islands, entities near his antipodes on the globe. The "rare" stuff.

On 80 meters and much more on Top Band the story is quite different. An example: The majority of stations on the West Coast of the USA are still chasing for European countries they have never heard so far, not because they are not being activated, but because there almost never is really good propagation on that difficult path. New countries for the Top Band DXer do not necessarily come in the shape of "rare" countries, as is the case on the higher bands.

Contests on 80 meters and more so on Top Band offer enhanced chances to work some new stuff, provided conditions are willing, simply because a great number of stations are on the band during the contest. Take the example of the now famous propagation during the 2009 CQ World Wide 160 Meter CW Contest. I have seen reports from many stations on the US West Coast saying that, in just a few hours, they worked anywhere from 5 to 10 new countries in Europe "right across the North Pole." What if these conditions did not occur during such a popular contest? Just a few of the regular night owls who live on the band would have witnessed these great conditions and maybe would have worked one or two new countries.

So, contests are sometimes a good place to work new countries, conditions willing. That's the gambling part. Being in the contest increases your chances to work new stuff, if propagation is nice to you.

That's why most of the low-band DXers, and more specifically the Top Band DXers are also contesters. Maybe they are not avid or die hard contesters, but they know it's a place and a time where chances to work a new country are higher than any time else.

One of the problems with working the relatively weak DX stations is that, during the contests, the QRM level is high. On the receive side we now have excellent radios that can cope with very strong signals that are very close to the frequency. These transceivers also excel by the high-quality click-free CW signals they generate. Unfortunately, not all radios are that good yet. Users of lower quality transceivers will suffer on the receiving end when operating close to other strong signals.

A worse problem is the fact that some radios have poor quality transmit signals (key clicks and noise sidebands) will disturb their neighbor operators in the band. During one of the recent big contests on Top Band I noticed some stations with well known calls, and running high power and big antennas, had inferior quality signals on the band. Let me appeal to all the stations that get onto the low bands, even if it is only in the contests, to make sure the quality of their signals is on par with the state-of-the-art of the available technology. Get rid of those radios that produce clicking signals and noise sidebands. Make sure you can be proud, not only of the number of QSOs you made in the contest, but also of the quality of the signal you transmit on the air.

Let's behave as gentlemen not only when chasing DX, but also during contests. I am talking about ethics and operating procedures. One particular situation seems to repeat itself over and over in every contest. It happens all too frequently that a "big gun" contest station has seen a multiplier on the DX cluster. He's been working on his run frequency for a long time, but now he has to leave it for chasing the multiplier. If, in the mean time, after having (yes or no) worked the new multiplier somewhere else on the band, he finds his former run frequency taken over by another station, too bad for him. That's the name of the game.

It happens in every contest that, while I am doing a search and pounce session on the bands (just moving across the band looking for stations I have not yet worked), I suddenly find a large clear spot. I ask QRL? and often the station running on the frequency will resume calling CQ. No problem. Every now and then the QRL? query and a second QRL? result in nothing but full silence on a perfectly clear frequency. This, you can be sure is where one of the big stations was running before he went away chasing a multiplier. Too bad for him but now it is "my" frequency. After I already have made one or two or even more QSOs on this frequency, all of a sudden a big gun will start calling right on top of me, as if he never left the frequency. This is very unethical. And most of the time these guys won't listen to reason, and would rather have a fight than behave ethically. If you go chasing multipliers, there's a fair chance you will lose your running frequency — that's part of the game. There's a French saying that describes this situation wonderfully: "*He who goes on the chase, loses his place.*" Come on guys, don't do this. Let's behave like gentlemen. After you worked your multiplier, it maybe is a good idea to go on a search and pounce tour yourself.

18. THE SUCCESSFUL LOW-BAND DXER

If we want to analyze what's required to become a successful low-band DXer, we must first agree on what is success. Success is very relative. If you have only a ⅛-acre city lot and you want to work the low bands, your goals will have to be different from someone with 10 acres and a well-filled bank account. But you can be successful in your own way, relative to your own goals.

There are a few essential qualities that make good low-band DXers, I think. They also apply to the low-band DXer with a modest setup. Three of the most important qualities are called the "3 Ps" principles by Bill, W4ZV: Propagation, Patience and Persistence.

Understanding and experience in matters related to propagation: Don't expect to turn on the radio any time of the day on 80 or 160 meters and work across the globe. You must understand that you are trying to do something very difficult, something that requires a lot of experience to be successful. You'll have to be able to predict "possible" openings, sometimes with an accuracy of minutes. The successful low-band DXer must build up his propagation expertise over a long period of time.

Perseverance, persistence: If you are not prepared to get up in the middle of the night five days in a row to try to work your umpteenth country on 80 or 160, you won't make it to the "top 20" in our sport. If you think it's too hard to go out at night in the fields or through the woods to roll out a special

Fig 2-25 — Contests offer good opportunities for newcomers on the low bands to work new countries. XE1RCS (managed by XE1KK), a top-notch phone contest station from Mexico, has given a new country to many 160-meter DXers worldwide.

one-time Beverage for that new country you have a sked with in a few hours, then you better forget about becoming really successful in the game, or rather the art, of low-band DXing!

Patience and dedication: Bill gave as an example that it took him many seasons to finally work JT1CO and UK8DAN. I can say exactly the same about the countries in my black hole in the center of the Pacific. From Europe it is just not possible to work 25 of the very difficult countries in the central Pacific in one or two years.

But there are others:

Understanding antennas: For the low bands, it is not like opening a catalog and ordering an antenna. You have to understand antennas — the Whys and the Why Nots. You will have to become an antenna experimenter to be successful, even more so if you'll have to do it from a tiny city lot!

Willingness to learn: Isn't improving our technical knowledge and ability what our hobby is all about? Working DX on 160 meters makes you feel like you are doing it like Marconi!

Equipment and technologies: Receivers are getting better at every vintage, even if the evolution isn't moving as fast as we might like. The successful low-band DXer uses the best equipment available and uses it in a professional way. He gets involved with the latest technologies in radio communication, such as the Internet, DX Clusters etc. These provide real-time information about activity on the different low bands.

Good QTH: The top scores in DXCC and in low band contesting are set from excellent QTHs. Different from VHF and higher, excellent QTHs for the low bands are often flat, rich lowlands. This does not mean that a successful low-band DXer has to be a rich land owner. I, for one, have just over an acre, but my location is excellent. The neighbors are nice and I can use the fields in the winter for my Beverage antennas.

Operating proficiency: Your "know-how-to-do-it" is probably the best weapon that can make or break a low-band DXer with a modest station.

Willingness to become a good CW operator. I don't think this needs to be explained! As VO1NA put it: "on 160, CW shines."

19. THE 10 LOW-BAND COMMANDMENTS

Mark Twain once said: "If we were supposed to talk more than we listen, we would have two mouths and one ear." How true this is for low-band DXing — and for most other human endeavors.

Jeff, K1ZM, published in his excellent book *DXing on the Edge* (Ref 511) a set of rules, from the hand of Bill, W4ZV, and which had been published earlier on the Top Band Reflector. It goes without saying that these rules apply equally well to the other low bands. A chapter on operating would not be complete without these rules, which I like to call the 10 Low-Band Commandments:

Rule #1: When the DX station answers someone else, listen — do not call. Instead try to find where *he* is listening. Most good operators spread the pileup over at least 1 to 2 kHz. If you listen for the station he is working, you will maximize your probability of being heard since you will know where he is listening. You may also recognize the pattern the operator uses. That is, is he slowly moving up in frequency, down in frequency or alternating picks on either side of the pileup? You will also know when to transmit (ie, when *he* is listening). It's very hard for him to hear you calling while he is transmitting!

Rule #2: Listen carefully! He may change his QSX (split) frequency or move his transmit frequency. If you're calling continuously, you will never know it. I can't tell you all the good stuff I've worked easily because I was one of the first on a new QSX frequency. If you're transmitting continuously, you'll be one of the last to know. For those of you with QSK, you have an advantage here. If you don't, use a foot switch so that you can listen between calls and stop sending when he starts.

Rule #3: Do not transmit on top of the station answering. Why? Because a good operator will stay with that station until he finishes the QSO. Repeats necessitated by your QRM just reduce the amount of time *you* will have to work him before propagation goes out. The name of the game is for the DX to work as many stations as quickly as possible. Continuously calling only slows down the whole process and reduces *your* probability of a QSO. It might also encourage some DX operators to make a mental note in their head to never "hear" you again!

Rule #4: Learn your equipment so you know how *exactly* to place your transmit signal properly on frequency. No, this does not mean exactly zero beat on the last listening frequency where all the other guys are. It's far better to offset by a few hundred Hz based upon which way you think the DX is tuning (see Rule #1). Also *please* learn to use your equipment so you don't transmit on the DX frequency inadvertently. This only slows things down for everyone and wastes precious opening time on 160 meters.

Rule #5: If you have limited resources on 160, focus on your receive-antenna capability. You will work far more 160 meter DX with good ears than with a big mouth. Being an "alligator" that cannot hear anything is not productive on Top Band.

Rule #6: Always send your full call. Partial calls only slow things down. (from Rolf, SM5MX, XV7SW)

Rule #7: Use proper and consistent spacing when sending your call on CW. There are some very well known DXers who don't understand this. They will break the cadence of their calls with pregnant pauses — this can confuse the DX station trying to decipher your call through 160-meter QSB and QRN.

Rule #8: Send the DX station's call if you are in doubt about who you are working. You will not be happy if you log a DX station while you actually worked another station! This is especially important if more than one DX station is working split in the same general area of the band. (from 4S7RPG)

Rule #9: Listen to the DX station's reports and match his sending speed. If he is giving 459 at 18 WPM, don't reply at 35 WPM! If the DX station is missing part of your call, or if he has incorrectly copied part of your call, repeat only that part of the call several times, at a constant pace. (from 4S7RPG)

Rule #10: Listen, ... listen, ... listen!

20. THE SURVEY

During January 2009, e-mails were addressed to over 2000 hams worldwide. Recipients included active low-band DXers and contesters as well as a large number of hams who subscribe to the Top Band reflector. These e-mails contained a range of questions concerning equipment and antennas used, as well as operating results (DXCC, WAZ), for 160, 80 and 40 meters. Within a few weeks over 400 replies were received. I used more than 95% of the responses (those that hardly answered any of the questions were not used). The data were collected in an Excel file which is included on the CD that comes with this book (survey-2009.xls).

I would like to thank all the participants in this survey for taking the time to help me collect these valuable data in order to make a "picture" of the low-band operator and his station/antennas.

20.1. The Analysis

All data were carefully analyzed and in several areas compared to the survey results from five and ten years ago. The results are very interesting. During the analysis of the vast amount of information (assembled in a spreadsheet with 109 columns and over 400 rows) I made a clear distinction in two fields:

1) Between each of the three low bands: 160, 80 and 40 meters.
2) Between the *total group* of respondents and a selected band group, which I call the *top group*. The top group is composed only of stations having met a certain DXCC criterion. For the 160-meter band, stations are part of this top group if they have worked at least 225 DXCC entities. On 80 meters the lower limit is 275 entities, and on 40 meters it is 300 entities.

In order to obtain meaningful results, the sample size should be big enough. With a total of well over 400 low-band DXers who filled out the questionnaire and sent in their data, we came to the following sample sizes:

- 160 meters: total group: 397 stations; top group: 88 stations
- 80 meters: total group: 397 stations; top group: 128 stations
- 40 meters: total group: 393 stations; top group: 170 stations

With such large sample size, the conclusions of the analysis will be meaningful.

20.2. Operator's Age and Activity

Approximately 120 survey participants indicated their exact age: the oldest one was 91 (congrats Chas, WØCD), the youngest one 30. The average age of the group was 59, although most of the low banders seem to be between 60 and 70 years.

Fig 2-26 shows the age distribution of the sample of low-band DXers compared with the age distribution of the total ham population in Belgium (thanks ON7TK). Note that the average low bander's age is approximately 10 years higher than for the "average ham" group. It means we desperately need young blood in the ranks of low-band DXers! (The older distribution may also mean there are simply more older hams who have accumulated the resources for the antenna acreage needed on 80 and especially 160.)

As I did for previous editions of this book, I asked how long the respondents had been active on each of the three low bands. **Fig 2-27** shows the results. Note that the curve for 40 meters is shifted most to the right, which means that 40-meter DXers have been around the longest time: on average 33 years.

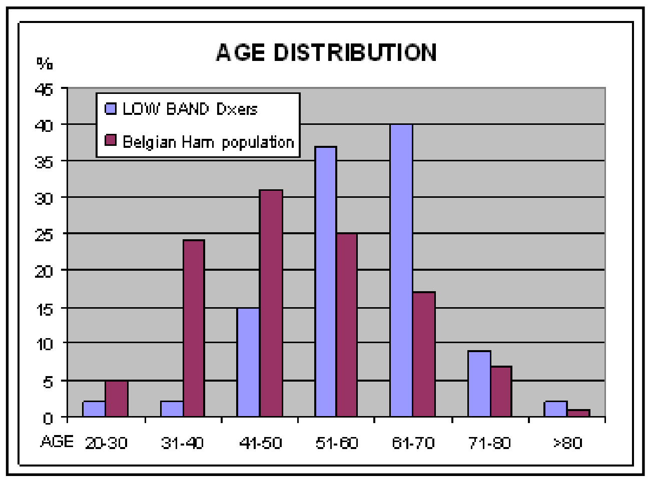


Fig 2-26 — Graph showing age distribution for low-band DXers compared to the overall ham population in Belgium.

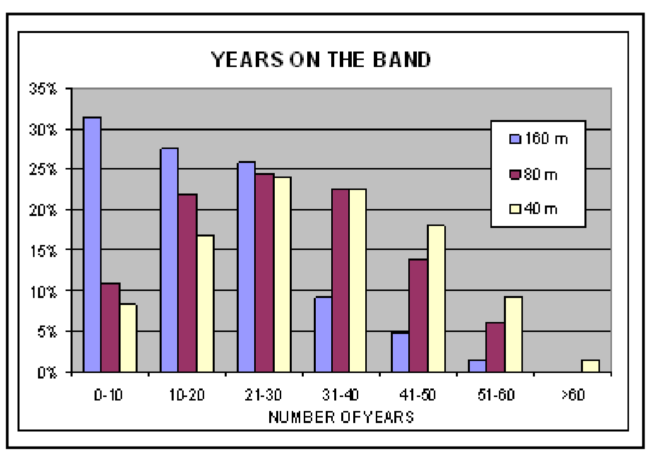


Fig 2-27 — Graph showing the number of years low-band DXers have been active on each band.

The 80-meter curve is shifted a little to the left, which means there are a number of hams of more recent vintage on 80 meters. On average they have been around 29 years.

The 160-meter curve resembles a cut-off Gauss curve, which in this case means that the numbers of 160-meter DXers are still growing. The 0 to 10-year group is the largest group and represents 29% of the active 160-meter DXers. That is very good news. Top Band apparently still attracts a large number of "newcomers," and I think this is mainly because this is the band where you can get most satisfaction (thrill, joy, pleasure) out of your DXing and also contesting activities.

The average time Top Band DXers have been around is only 20 years (vs 29 and 33 years for 80 and 40 meters). Compared to the survey from five years ago, this average has gone down from 22 to 20. This change in percentage is quite meaningful, as the sample is reasonably large (nearly 400 stations) and selected fairly at random.

20.3. Equipment

20.3.1. The Total Survey Group

For the total group, 414 stations listed 507 rigs. This means you cannot add up the percentages shown in **Table 2-3**, as they will total more than 100% (some stations listed two and even three radios).

It is no surprise that the Yaesu FT-1000 series of transceivers (D, MP, MP MarkV) score the highest as this series of radios has been available for almost 20 years. The big surprise comes from the fact that the Elecraft K3 scores a distinct second (tied with the IC-756), only just over a year after it came on the market. The Orion (I and II) score third, which is no surprise as it is, together with the K3 considered by many of the experts as the best transceiver on the market. Of the "expensive" radios the Icom IC-7800 and IC-7700 scored well, while the Yaesu FTDX9000 and the FlexRadio SDR transceivers are almost totally absent (mentioned only four times in nearly 400 surveys).

20.3.2. Stations Active on Top Band

The total group of 397 stations listed 493 rigs; for the top group 112 transceivers were listed by 88 stations. The most significant difference between the total group (397 stations) active on 160 meters and the top group (stations having worked at least 225 DXCC countries) is the fact that the K3 and the Orion are used by a higher percentage of the sample (see **Table 2-4**).

Table 2-3
Transceivers Used by Total Low Band Survey Group

Transceiver	*Total Group*
Yaesu FT-1000 (D-MP)	27%
Elecraft K3	12%
Icom IC-756	12%
Ten-Tec Orion	9%
All Kenwood equipment	7%
Other Icom equipment	7%
Yaesu FT-2000	6%
Icom IC-7800, IC-7700	5%
Other Ten-Tec equipment	5%
Other Yaesu equipment	4%
Elecraft K2	4%
Various	1%
FlexRadio SDR	<1%
Homemade	<1%

Table 2-4
Transceivers Used by 160 Meter Operators

Transceiver	*160 Meter Total Group*	*160 Meter Top Group*
Yaesu FT-1000 (D-MP)	26%	28%
Elecraft K3	13%	16%
Icom IC-756	12%	6%
Ten-Tec Orion	9%	11%
All Kenwood equipment	8%	7%
Other Icom equipment	7%	6%
Yaesu FT-2000	6%	6%
Icom IC-7800, IC-7700	5%	6%
Other Ten-Tec equipment	5%	4%
Other Yaesu equipment	3%	4%
Elecraft K2	3%	5%
FlexRadio SDR	<1%	0%
Various	1%	1%
Homemade	0%	0%

Table 2-5
Transceivers Used by 80 Meter Operators

Transceiver	*80 Meter Total Group*	*80 Meter Top Group*
Yaesu FT-1000 (D-MP)	27%	31%
Elecraft K3	13%	12%
Icom IC-756	11%	7%
Ten-Tec Orion	9%	8%
All Kenwood equipment	7%	7%
Other Icom equipment	7%	7%
Yaesu FT-2000	6%	4%
Icom IC-7800, IC-7700	5%	7%
Other Ten-Tec equipment	5%	5%
Other Yaesu equipment	3%	5%
Elecraft K2	3%	6%
FlexRadio SDR	<1%	0%
Various	<1%	1%
Homemade	0%	0%

Table 2-6
Transceivers Used by 40 Meter Operators

Transceiver	*40 Meter Total Group*	*40 Meter Top Group*
Yaesu FT-1000 (D-MP)	27%	30%
Elecraft K3	13%	12%
Icom IC-756	12%	12%
Ten-Tec Orion	9%	10%
All Kenwood equipment	7%	5%
Other Icom equipment	7%	7%
Yaesu FT-2000	6%	5%
Icom IC-7800, IC-7700	5%	7%
Other Ten-Tec equipment	5%	6%
Elecraft K2	4%	4%
Other Yaesu equipment	3%	3%
FlexRadio SDR	1%	<1%
Various	1%	<1%
Homemade	0%	0%

20.3.3. Stations Active on 80 Meters

The total group active on 80 meters (397 stations) listed 491 rigs, while the top group (128 stations) listed 165 transceivers. The figures (**Table 2-5**) do not differ a great deal from those collected from the 160 buffs. Only the use of the K3 and the Orion seems to be a little less pronounced, maybe because on Top Band buffs seem to be more critical and because the top-notch performance of these radios pays off even more on the difficult 160-meter band than it does on 80 meters.

20.3.4. Stations Active on 40 Meters

The total group active on 40 meters counted 393. They listed 486 transceivers. For the top group 217 transceivers were listed by 171 stations. Here too, the overall picture is not very different from the other bands (**Table 2-6**).

20.3.5. What's the Best Transceiver?

Besides asking what was used (at the time of the survey), I also asked what was considered the best transceiver on the market (at that time, January 2009). A total of 166 stations from the overall group (413 stations) had a clear opinion. Of the 160 meter top group (88 stations having worked at least 225 countries on Top Band), 36 stations expressed their view. The answer was very straightforward and simple (see **Table 2-7**).

Whereas 16% of that group are already using a K3, 59% were clearly considering getting one or had it already on order. It is short of amazing to see that 81% of the top group were of the opinion that the US-made Elecraft K3 and the Ten-Tec Orion were the best. Excellent, and let it continue that way!

The exercise was repeated using the total group (166 stations answered the question out of the total group of 413). The results roughly confirm the trend, with the only difference that out of the total group a larger percentage thought the FT-1000 was the best radio and a smaller group was using the Ten-Tec Orion.

Tony, T77C, gave a very pertinent answer to the question "What do you think is the best transceiver for the low bands?" He said "Your ears." Right on he is!

20.3.6. Evolution and Summary on Equipment

- Since the previous survey done in 2003, overall the FT-1000 which, at that time, had a share of almost 60% (vs 44% in 1998), has dropped to less than half of that percentage. All of this means there are still many FT-1000s around. I hope all of those have by now been modified to avoid the key clicks that are so typical for these transceivers. It is clear that the FT-2000 will replace some of the FT-1000s, although some commented that the FT-2000 would not be as good as the FT-1000.
- The Kenwood share has continued to dwindle from 20% in 2003 (30% in 1998) to merely 7% today.
- The total Icom share which had grown from 17% in 1998 to 27% in 2003 is now 24%.
- Ten-Tec has grown from 7% in 1998 and in 2003 to 14% now.
- It is obvious that the new and bright star in the sky is the Elecraft K3, which I expect to at least double its share or do even better than that in another five years. The survey was done in early 2009. I am sure that at the time this book will be out (late 2010), the Elecraft K3 will be at or near the top.

Amazing observation: *only one* of the JA stations in the survey admitted using US-made equipment (a K2, which he said was "the best"). The 245 US hams who participated in this survey specified 296 rigs, of which only 30% were US made (Elecraft, Ten-Tec or FlexRadio). Seventy percent of the radios they use are made in Japan, and this despite the fact that 79% of the total survey group state that the best equipment (for the low bands) is made in the USA (K3 and Orion).

Table 2-7
What is the Best Transceiver?

Transceiver	*160 Meter Top Group (36 responses)*	*160 Meter Total Group (166 responses)*
Elecraft K3	59%	61%
Ten-Tec Orion	22%	18%
Icom IC-7800, IC-7700	8%	6%
Yaesu FT-2000	4%	3%
Yaesu FT-1000 (D-MP)	4%	10%
Icom IC-756	2%	3%

20.4. Antennas Used by the Low Banders

20.4.1. On 160 Meters

For 160-meter antennas, I compared this survey with the results from six years ago. There seems to be little change in the picture, except maybe that the top group uses higher performing transmit antennas (more vertical arrays) and that receiving arrays (eg the DX Engineering receiving Four Square) have more than doubled (**Table 2-8**).

20.4.2. On 80 Meters

It is amazing to see that 41% of the top group on 80 meters do not use Beverages, but these are the guys that use quads or Yagis (11%) or vertical transmit arrays (23%) or receive-only arrays (8%). If you add all these percentages you come to 52%, which means that all of the top group 80-meter DXers do use some kind of directional antennas on receive, even if they don't use Beverages. See **Table 2-9**.

Table 2-8
160-Meter Antennas

Transmit Antennas	*Total Group*	*Top Group*
Verticals	57%	59%
Dipole	21%	17%
Half sloper	10%	6%
Vertical array	7%	18%
TX loops	4%	6%
Sloping dipole	2%	6%
Yagi, quad	1%	0%
Horizontal array	<1%	0%
Other antennas	6%	0%
Receive Antennas	*Total Group*	*Top Group*
Beverages	39%	62%
RX loops	29%	21%
RX array	6%	9%
Other RX antenna	5%	15%
None	31%	15%

Table 2-9
80-Meter Antennas

Transmit Antennas	*Total Group*	*Top Group*
Verticals	36%	27%
Dipole	30%	23%
Vertical array	16%	26%
Half sloper	10%	10%
TX loops	4%	2%
Sloping dipole	4%	7%
Yagi, quad	4%	11%
Horizontal array	1%	1%
Other antennas	6%	4%
Receive Antennas	*Total Group*	*Top Group*
Beverages	33%	42%
RX loops	22%	14%
RX array	5%	8%
Other RX antenna	3%	1%
None	43%	41%

Table 2-10
40-Meter Antennas

Transmit Antennas	*Total Group*	*Top Group*
Yagi, quad	45%	62%
Verticals	25%	16%
Dipole	25%	19%
TX loops	5%	4%
Sloping dipole	2%	2%
Vertical array	2%	2%
Half sloper	1%	1%
Horizontal array	0%	0%
Other antennas	5%	4%
Receive Antennas	*Total Group*	*Top Group*
Beverages	21%	22%
RX loops	8%	7%
RX array	2%	2%
Other RX antenna	2%	0%
None	66%	68%

20.4.3. On 40 Meters

Almost 50% of all 40-meter DXers use some kind of beam (up about 5% in both the total group and the top group as compared to 2003). Two-thirds of the 40 meter DXers do not use specific receive-only antennas, because so many DXers use a beam. See **Table 2-10**.

20.4.4. Overall Conclusion on Antennas

The use of different antenna types on the three low bands has not changed a lot over the past five years. The main difference is that receiving arrays made of short vertical elements, as abundantly described in the previous edition of the book, have made their entrance on the floor, together with the availability of a commercial receiving type Four Square.

20.5. Time Spent on Each of the Low Bands

Table 2-11 shows the average time the low bander spends on each of the three low bands. Notice that the popularity of 160 meters is slightly on the increase to the detriment of 80 meters. What is the split for time spent on the 40, 80 and 160-meter bands by low-band DXers?

Table 2-11
Time Spent Per Band

Band	*2009*	*2003*
160 meters	47%	45%
80 meters	28%	31%
40 meters	25%	24%

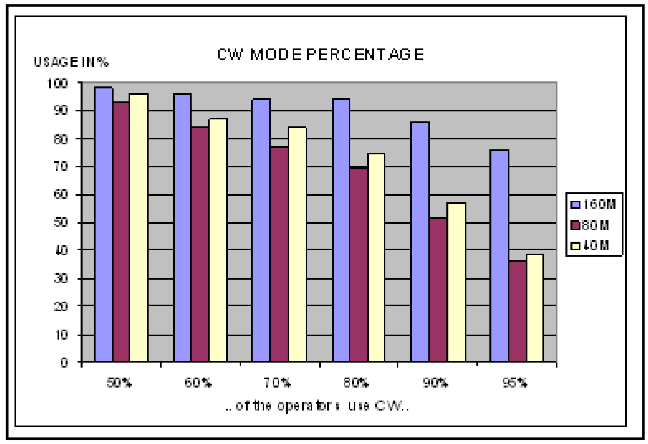


Fig 2-28 — Graph showing percentage of CW operation among low-band DXers on 160, 80 and 40 meters.

Forty percent (33% in 2003) operate 40 meters only occasionally (less than 10% of the time), while 14% work 40 meters more than half of the time. This obviously means that 40 is not really a low band: this band clearly does not represent enough challenge for the die-hard low banders.

The number of low-band operators who spend more than 50% of the time on 80 meters has gone down from just under 10% in 2003 to just over 7% in 2009! The apparent drop in popularity of 80 meters with the low-band operators is likely due to the fact that an ever-increasing number of operators (nearly 20%) have worked 300 countries or more and have moved to 160 meters for more challenges and new opportunities to work new countries.

The picture is very different for Top Band: Over 49% (was "only" 30% in 2003) spend more than 50% of their time on 160 meters. Twenty percent even spend more than 80% of their operating time on 160 meters, while only 14% spend less than 20% of their low-band operating time on Top Band. Only 3% of the total population of participants in the survey (approximately 400) never operate 160 meters.

20.6. Modes

It is more than obvious: The low bands are the bands where DX is worked mainly in CW (see **Fig 2-28**). How do you read the chart? It says that 50% of the low-band operators use CW on 160 meters 98% of the time, and 95% use CW 76% of the time. The use of CW on 40 meters is a little more pronounced than on 80 meters and that is of course because of the restricted bandwidth on 40 meters (until 2009) for CW.

On the average, the 160-meter operators use CW 93% of the time. On 80 meters this figure is still 79%, and on 40 meters, 82%. Let's not forget that the challenge of working

Table 2-12
Use of CW

Band	*Total Sample*	*Top Group*
40 meters	82% (79% in 2003)	83%
80 meters	78% (70% in 2003)	74%
160 meters	92% (90% in 2003)	95%

DX on the low bands is for many also related to the necessity of working CW, and being a good CW operator.

It is clear that Top Band is a CW band: More than 82% operate more than 90% of their Top Band time on CW (28% work nothing but CW on Top Band). Less than 10% operate more than 20% on phone on 160 meters.

Actually all three low bands are CW bands for the low-band DXers. On 80 meters exactly 50% use CW 90% of the time, and almost 90% say they use CW at least 50% of the time. These high percentages may come as a surprise.

The SSB activity by the low banders on 80 meters is down considerably from six years ago. Whereas 33% of the respondents in 2003 said they work phone 50% of the time or more, this figure is now down to only 13%. The die-hard phone operators who use SSB exclusively on 80 meters are very few (2%).

Forty meters is still mainly a CW band for the DXers: 52% use CW for their DX contacts at least 90% of the time (25% at least 99% of the time!). On the other hand, only 3% of the 40-meter DXers use SSB more than 50% of the time (less than 2% use SSB more than 90% of their operating time on 40 meters). Finally, two-thirds use SSB more than 10% of their operating time on 40 meters.

If we examine the split for stations with a high DXCC score (the top performance group, see Section 20.1), the preference for CW becomes even more pronounced as shown in **Table 2-12**, which lists the average time spent using each mode for the top group as well as for the entire sample.

We can safely conclude that for the top group as well as for the total sample, DXing on the low bands means operating CW almost all the time. It certainly is untrue that CW is on the decline, at least among low-band DXers. Now that in most countries the knowledge of CW is no longer mandatory, hams work CW not because it's a "must," but because they know it's a superior mode when trying to make contacts under marginal circumstances. Also, you seem to find more gentlemen on code than on phone.

Digital modes: If you look at the *Excel* file on the CD that comes with this book, you will notice that for a given call sign the sum of the column listing the CW and the column listing the SSB percentage is not always 100%. The difference is the percentage spent on digital modes. On total average this represents less than 1%.

20.7. Achievements

The achievements by station can best be viewed on your PC using the CD that comes with the book. The listing is done alphabetically by call sign, but you can of course sort out the table in any way you want. The WAZ and the DXCC scores for each of the bands were used in the survey. Some respondents only gave the confirmed score, in which case I entered the same score in the "worked" column. Where sorting was done according the DXCC score (to determine the top group), the "worked" column was used. If someone filled out ~200 or >200, I entered 200.

20.7.1. Achievement Summary

Percentage of respondents holding award:
5-Band DXCC: 68%
5-Band WAZ: 36%
160-meter WAZ: 34%
These figures are almost identical (within a few percent) of those published six years ago.

Average all-time DXCC count:
40 meters: 274 (up 14 countries vs 2003)
80 meters: 235 (up 10 countries)
160 meters: 170 (up 6 countries)

Average WAZ count:
40 meters: 38.6 (38.2 in 2003)
80 meters: 36.6 (36.2 in 2003)
160 meters: 32.0 (31.0 in 2003)

The purpose of the listings is not to give an accurate DXCC status report, but to show what some of the leading low-band DXers have achieved. Using the *Excel* file on the CD, you can link the results to antenna and equipment data.

Figs 2-29 and **2-30** give the distribution of the DXCC and WAZ status on 40, 80 and 160 meters for the respondents

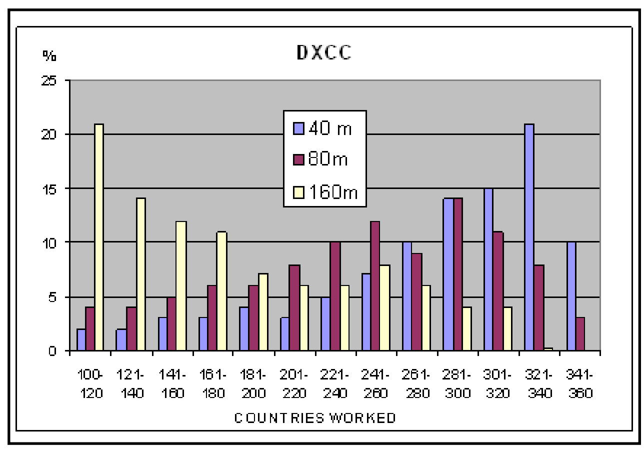


Fig 2-29 — Graph showing DXCC countries worked by low-band DXers on 160, 80 and 40 meters.

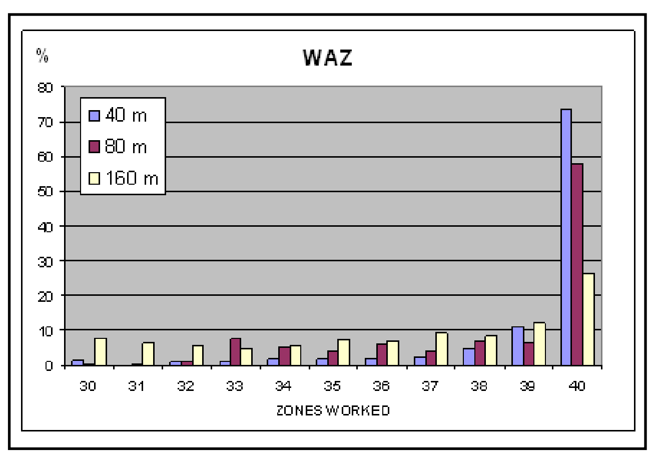


Fig 2-30 — Graph showing CQ WAZ zones worked by low-band DXers on 160, 80 and 40 meters.

to the survey (~400 entries for each band).

The DXCC chart shows the distribution of all stations having worked at least 100 countries on the three low bands (40 meters: 273 stations, 80 meters: 265 stations, 160 meters: 205 stations, out of a total survey population of ~400 stations).

It is truly amazing to see that on 80 meters at least 153 stations (out of ~385 stations) have worked all 40 zones. Back in 1979 when the 5 Band WAZ was created, it was considered very difficult to get all 40 zones on 80 meters.

Even more amazing is that on Top Band, at the time of the survey 54 stations in the sample have worked all 40 zones (26% of the group). Before 1987, the year when access was given to 160 meters in Belgium, there were only two 160-meter WAZ (all 40 zones) awards issued (I have #3). All of this illustrates the tremendous growth in interest in DXing on the low bands, and more in particular on 160 meters, we have seen over the past 20 years.

20.8. Why Do We Want to Work DX on the Low Bands?

This is a rhetorical question, we all know the answer, but as in the past editions of the book I wondered if anyone would find a more original way of saying "it's the challenge." The comments below were collected from the surveys done in 1998, 2003 and 2009.

- The feeling of achievement (G3NKC)
- The challenge, the feeling of camaraderie, better operating standards (GM3YT)
- Satisfaction, accomplishment doing it with QRP (K1HTV)
- To experiment with antennas (IK7JTF)
- Because it requires operating skill and knowledge about antennas (N4JJ)
- Achievement as a result of study and experimentation (EY8MM)
- Excellent bands for a night owl like me (DF3KV)
- No challenge on the high bands, Top Band is still to be figured out (K3UL)
- Experimenting with antennas, Top Band ops are a cut "above" the rest (N1BAA)
- The fascination of 160 (N2WM)
- Unpredictable propagation on 160 (N2NT)
- The necessity of know-how to be successful (AD8P)
- Challenge and camaraderie (W3UM)
- The thrill of doing it with less than ideal antennas (W2GDJ)
- The unique propagation on 160 meters (WA1Z)
- Takes more patience and skill (K4LTA)
- Because I enjoy it (K5OVC)
- Top Band is the last frontier of Amateur Radio (IKØYVV)
- There's always something unexpected (W7IUV)
- Harder, thus more satisfying. You also meet the better operators on the low bands (NN4T)
- Greater challenge, thus greater satisfaction (SM4CTT)
- It's not so easy… (PA3AAV)
- Challenge. 160 meters fits my time (availability) table … (W8RU, KA9FOX, K1KP, JA2VPO)
- Because CW excels for DXing on the low bands (SM2CEW)
- Being able to take pride in a home-designed home-made 160-meter antenna system (KK3AN)
- Challenge, room for experimentation, leaning about propagation (N8PR)
- I love digging weak signals out of the noise (K2XA)
- W1BB got me interested on Top Band…and it's still my favorite band (WØAIH)
- The thrill of catching the unexpected opening (AA7A)
- Challenge, satisfaction, camaraderie (K1DT)
- Just because it's not easy (VE7SL)
- Because it's a lot of fun (NØTT, I4EWH, WB9Z, HAØDU)
- The thrill: it is fantastic (JA1EOD)
- To join the gentlemen's league on Top Band (DL7ZZ)
- Challenge, camaraderie and generally good operating practices (VE7VV)
- Low band DXing requires a good knowledge of propagation and CW (K2ONP)
- The technical challenge (antenna building), more courteous operators, more friendship on 80 and 160 (K3SV)
- The challenge to do better, experimenting, because I suffer from the "Top Band disease" (K7TJR)
- Top Band teaches me to become a patient person… (YCØLOW)
- The challenge, possibility to develop, build and test transmit and receive antennas, Top Banders are a 'different' kind of ham (SP5EWY)
- The challenge, the unexpected surprises, the fun (K4CIA)
- Because it is not mainstream Amateur Radio. It entails all that real Amateur Radio should: home brewing, antenna design/work, contesting, competition, CW (VE3XZ)
- The challenge; the low-bands ops are generally good ops, makes more fun (K1KO)
- The challenge; the pain (K4UTE)
- It is very noisy and very difficult, but I love it (JO7MKB)
- Amazing ! (EA5BY)
- The challenging battle against noise and QRN (YV1DIG)
- I like the "sound and climate" of the low bands (SP2XF)
- Friendly operators in combination with challenging conditions (SM7BIC)
- Because I am a masochist…Top Band is EXCITING! (SV3RF)
- 160 meters is one of the most challenging aspects of Amateur Radio (KR4OW)
- The challenge. If it's too easy, it loses appeal. (N4CC)
- The combination of technical challenges (propagation, antennas, etc) and the generally great group of ops especially on 160 meters (WØFLS)
- The fun and the room for experimentation (JA6BZI)
- The courteous operators on 160 meters
- Because it's hard (SP3DOI)
- More challenge, less lids (N2LQ)
- The challenge and being with other like-minded guys (W4ZV)
- Because I like it the hard way (OZ1LO)
- More difficult = more fun (I4EAT)
- Challenge: need to develop skills in a number of fields: propagation, antennas etc. Not like shooting fish in a barrel (K7FL)
- DXing on the high bands is like shooting fish in a barrel. (AA4MM)
- Low-band DXing is the greatest challenge in Amateur Radio. (ABØX)
- I love a good "static salad." (K1UO)
- Anyone can do it if it's easy. (K4PI)

- I experience the same thrills as 40 years ago that hooked me on radio, high bands are too easy. (K4TEA)
- Top Band is the only band that still gives me a thrill. (K6ANP)
- 160 is an addictive band, 160 is not easy to be good at. (KO1W)
- Worked a new one on 160 is not so cut-and-dried. (KX4R)
- On 160, CW shines. (VO1NA)
- Fewer lids than on high bands. (WØGJ)
- I am a masochist (NI6T)
- See how much pain one can endure before taking the headphones off. (W7TVF)
- Fun. (WB9Z) — *These guys should get together.*
- The low bands are where you can test the station and the operator's skills. (UA3AB)
- Challenge of hearing, silence the utility poles, be ready all the time. (N7RT)
- Best demonstration of operating skill, station design and knowledge of propagation (like 6 meters). (W4DR)
- Pushing the operator and the station to the limits. (N4KG)
- 160: No nets, no lists, no deliberate QRM, moving on the edge, alone with QRN... (IV3PRK)
- 160: This is a new mountain to climb (the tallest one). (N6RK)
- 160 requires more technical skills and operating skills: The ultimate DXing challenge. (K9RJ)
- On the low bands success comes through knowledge (antennas), not money. Few do it well. (K1VR)
- It's not that easy but I like difficulties (easy things are for everyone). (RA3AUU)
- DX nets on high bands make many contacts phony; playing field on low bands is more level. (ZS6EZ)
- Why do you climb mountains? …because they're there. 160 is the highest mountain with no worn path. (KØHA)
- Try to get the impossible, work all countries on all bands. (HB9AMO)
- I like difficult things, and... if you can't hear them you can't work them. (ON7TK)
- 160 is like the BC band, I was a BC SWL as a child. (N5SV)
- 160-meter DX requires the best of everything: antennas, equipment, QTH, operator skills. (4X4NJ)
- I think I was dropped on my head when I was a baby. (K4SB)
- On 160 you can be competitive using your hands, not your checkbook. (NW6N)
- Doing the impossible from a city, camaraderie on West Coast. (K6SSS)
- Satisfaction of achieving the seemingly impossible. (PA3DZN)
- The intellectual challenge of dealing with all the odd variables of propagation makes it a thrilling activity. (NØAX)
- I feel more at ease with my fellow low-band DXers than some of the "stuffed shirts" that hang out on 20 meters. (WØFS)
- Ties with early pioneers who did so much with so little. (K8MN)
- 160 is the absolute end in DXing, the last frontier. (K9UWA)
- To make the impossible possible: 160 DXCC from the worst place on earth. (YB1AQS)
- K6SE (SK) got infected at an early age: "As an 8-year old in Detroit I would stay up late at night do DX on the AM broadcast band."
- Chance to do something everybody thinks is impossible. (G4DBN)
- 160 is more a gentleman's band: Lids are too lazy to fight QRN. (W9WI)
- No pain, no gain, and no nets on 160 yet. (GW3YDX)
- Creates great friendships. (W6KW, ex-W6NLZ, ex-K2RBT)
- It helps to be insomniac. (W8RU)
- I am a man whose life begins after sundown. (AA4V)
- The challenge both on the technical side (antenna design and propagation) and DX techniques. (CT1EEB)
- More difficult and more value for each QSO. (EY8MM)
- You tend to find better operators on LF. (GM3YTS)
- For contest multipliers that are harder for others to get. (K1TTT)
- Operating skill as important as hardware. (K2RD)
- The challenges make it fun. (KØXM)
- Unusual and unexpected propagation and openings. (K3NA)
- Unpredictable! Fun! (K4CIA)
- Success is not automatic… (K4TEA)
- The challenge of the fight. (K6EID)
- It's a challenge! Plus it gets back to the roots of ham radio DXing. (K8BHZ)
- On 160 most operators are DX-oriented gentleman, good camaraderie. (K9FD)
- The engineering needed to be competitive on these bands. (K9JF)
- The sense of accomplishment, especially for my limited antennas and real estate. (K9KU)
- It's not something the average ham can do well, with the high noise, strange DX hours required, the skill and dedication needed to be successful. Nothing like working a new one" on Top Band (except for receiving the QSL!). (KG6I)
- The challenge doing it from my mobile. (W6/KH6DX/M)
- Results indicate antenna competency and operator savvy. (N4JJ)
- The lower the frequency, the higher the challenge. (NX4D)
- The challenge of propagation, the valuable awards. (S5ØA)
- Frees up some daylight time! (VE7BS)
- ANYONE can work DX on the high bands. (WØGJ, W1JZ, VE7ON, etc)
- 160 is the most challenging in terms of propagation and technicality. (W4ZV)
- Working the seemingly unworkable. (W9AJ)
- Testing antenna systems. (WXØB)
- Is there any challenge left in high-band operating? (ZS6EZ)
- The challenge on the low bands reminds of my early days as a new ham. (K4UEE)
- It has a charm of its own, and reminds me of early days with W1BB and W2EQS. (WØAIH)
- The lower the frequency, the higher the challenge. (NX4D)
- It's the only challenge left. (K2UO)
- I was inspired by W1BB; Stu gave me my Novice license test when I was about 12 years old. (AJ1H)
- It's fun and it's a challenge (PY2RO)
- It makes me happy when I work a new DX on these difficult bands (LY2IJ)
- The three-way challenge: Antennas, RF equipment and operating skill (WØCD)

I logged the reasons why the 349 participants of the 2009 survey operate the low bands, and classified them in a series of categories:

- 62% mentioned the challenge

- 14% stated that on the low bands you can or have design and build your own antennas
- 10% mentioned the unpredictable propagation on 160.
- 9% mentioned thrill and excitement.
- 8% mentioned the competition aspect (contests, getting awards etc)
- only 5% mentioned it was *more fun*.
- 4% mentioned that the low bands (especially 160 meters) require special skills
- 4% mentioned they like the company of the low-band operators better (160 is a gentleman's band).
- 5% mentioned that better operators are required to work the low bands.
- 3% mentioned they're on the low bands because they have worked just about everything on the higher bands
- 2% mentioned they operated the low bands because they are night bands
- 1% mentioned because that's all they can do during the low part of the sunspot cycle

In his response to the survey VE3MGY wrote: "DXing on the low bands is by far the ultimate technological challenge combining all of the classics — propagation, antennas, skill, equipment and know how. It is most definitely not an appliance operator sport!

"This is where you find the 'Special Ops' of the Ham Radio world. After all there are only two kinds of operators in my mind, the 160-meter operators and those who wish they were!

"No matter where I go 160 always comes up. Regardless where people spend their time — PSK, VHF, HF or moonbounce, 160 always comes up in discussions. It seems to hold everyone's interest just by the nature of how hard it is. Also on 160 there are no lists, no QRM, no lids and no unsure QSOs. It's just you and the band.

"Few operators will go to the extremes involved in getting that next 160 QSO or obtaining the next dB. Working a 'new one' on any other band is 'boring' by comparison when you have to consider all of the physics that have to come together on 160 at the right place and the right time to make a QSO happen. Working RZØAF in Zone 18 with 5 watts right over the pole from VE3 took 11 years of listening and trying to predict openings. It was also the highlight of my 160 meter career to date — and will be hard to beat!"

VE3MGY sure seems to know what he is talking about! Well said Brian!

20.9. QSL Cards

About 91% of the low-band DXers said they collect QSL cards, which is roughly the same as six years ago. About 85% say they answer all cards received. Six years ago E-QSLs were used by only 15%, today it is still only 30%. To make it clear: Logbook of The World (LoTW) is not E-QSL.

21. THE FUTURE

Competitions in DXing (DXCC, WAZ etc) and contesting have always been a driving force for improving antennas, equipment, operating procedures and Amateur Radio technology in general.

The hobby of Amateur Radio is all about communicating *by radio* with our equipment and antennas. Using the telephone to ask for a QSO in a contest has never been accepted as a normal way of making a QSO. Using the Internet to communicate with the DX station while trying to make a difficult QSO is not the way to do it. It seems there is a very large consensus among radio amateurs on this subject that this should not be done.

A long time ago (25 years) hams exchanged DX information on DX nets and used 2 meters to do that. A little later they did it via packet radio, still via exclusive Amateur Radio means.

Then came the Internet, the greatest blessing and, in a very few domains, the greatest evil. The greatest blessing because we now have all the information we can dream about at our fingertips. We can call our friends across the world via the "Internet telephone" almost (if not completely) for free. We can order our radio components, order our equipment via Internet, and do "Internet window shopping" without even moving a foot. We are spoiled.

These fantastic means and services that are provided to us via the Internet can be great evil as well. I am not referring to criminal scams. I am merely talking about the danger it may bring to Amateur Radio. Some people seem to be of the opinion that if you cannot make the contact (the entire stretch) via *radio* on the Amateur Radio bands, you can substitute part or all of it by some kind of "virtual radio" done via the Internet.

To me this cannot be Amateur Radio. Take as an example the so-called *Xtreme contesting* category, where almost anything seems to be permitted to make a contact. If we can set up remote receiving stations across the world, and cover most of the distance via the Internet, we are not playing *radio* any more, we are playing games, *virtual Internet games*. I think that remotely-controlled stations, whether for receiving or for transmitting, should be at a maximum distance of around 100 km from the base station, so that propagation is more or less the same at the two locations. Remote stations should be used only to overcome limitations imposed by urbanism (limited space, high noise, high interference risk) or geography (living at the bottom of a deep valley). In other words, don't use remote stations to overcome propagation limitations.

If you can set up a remotely-controlled receiving station anywhere in the world, it will no longer be necessary to try to build a good receiving site with top-notch receiving antennas for 160 meters. One would be able to make contacts in the "all mouth and no ears" style.

With all due respect, to me this is all just like Echolink/IRLP, with the only difference that at the end of the Internet link there is not the usual VHF repeater but a "closed circle" or private receiving or receiving/transmitting station at the other side of the world. All of this may, technically, be very challenging and require top notch technologies, know-how and expertise, but, in my humble opinion, this is *not* Amateur Radio, certainly not if any means like that are permitted in contests and DXing. If these technologies are permitted in the world of contesting and/or DXing, we enter the world of *virtual* Amateur Radio.

Some say "let's have an open mind" and allow all these new things, be it in a very closed and separate contest category. OK, as long as these stations only work other stations that play the same virtual game. But this is impossible. The stations playing in this new category will of course be making contacts with "normal" stations (call them little guns by comparison). Most, if not all of these little guns want to make "real" QSOs on the air, though. If the little gun in Europe makes a QSO on Top Band with a W7 who's living in Oregon, he wants to be copied in Oregon, and not via a remote station somewhere in Europe. A contact made for most part via the Internet is not a QSO for him.

What will be the value of your 160-meter WAZ, or your 340 countries worked on Top Band, via such virtual QSOs? What will be the satisfaction derived from such "contacts" (no, these are *not* "QSOs" by any reasonable definition).

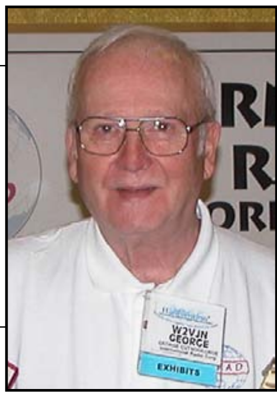
I believe that QSOs involving remote stations (RX and/or TX) should only count in contesting and toward DX awards if these remote stations are located less than 100 km from the home station. But how will that be checked? What about the little gun who works a great part of his DXCC countries during contests? Will he ever be sure that he made a "real" QSO?

Maybe I sound a little old fashioned, but in my opinion, Amateur Radio is all about RADIO, and all about QSOs made entirely by Amateur Radio means.

Contents

CHAPTER 3

Receiving and Transmitting Equipment



I would like to thank George Cutsogeorge, W2VJN, for being my critic and much trusted proofreader for this chapter. George has held the same call since 1947. He is an electronics engineer, retired after a 40-year career, including 17 years with RCA designing spacecraft and ground-station equipment and another 17 years with Princeton University, where he was involved with fusion energy research. George is heavily involved with electronics and radio amateurs. Thank you also, George, for letting me quote from your publications.

The performance of communication equipment has progressed by leaps and bounds over the years. About 10 years ago the Yaesu FT-1000 series transceivers were considered by many low banders to be the best available equipment, as judged from their popularity among low band fanatics. They were popular despite limited dynamic range compared to current transceivers and bad key clicks (see Section 2.5.1).

In recent years, major breakthroughs in receiver performance came from Ten-Tec (Orion) and Elecraft (K3), two American based companies that seem to live by marketing strategies that are very different from those followed by their main Japanese counterparts. The nice thing about this is that you don't have to be rich to own the really top notch transceiver.

While at the time this is written (February 2009) FlexRadio still seems to be the only "all in one box" software defined radio (SDR) transceiver on the market, it remains uncertain that a transceiver for which you need a bulky top range PC is what the market asks for. Also, at this time less than 1% of the more than 400 low band DXers who replied to the survey (see Chapter 2, Section 20) are using such a transceiver from FlexRadio. In the latest SDRs, the signal processing that used to be done by a PC with sound card is done by built-in hardware, although a PC is still required to run the SDR software. The evolution in the field of SDR is so fast that what I write today may be outdated by the time it is printed.

It is undeniable though that SDR transceivers will provide a major terrain for further developments and breakthroughs. Already we can make a very competitive 160-meter receiver with a waterfall or spectral display with excellent performance using only a small Softrock SDR circuit board (less than $20), a good sound card ($100) and a standard fast PC, which most of us have anyhow! I have been amazed at the performance of such a little SDR receiver on top band. I can literally see a very weak signal on my waterfall screen that I would probably never have detected on my regular radio, unless it had been announced on the DX Cluster!

1. THE RECEIVER

1.1. Receiver Specifications

Over the years, receiver performance and the expectations of amateurs have changed. Until about 40 years ago, receiver performance was almost exclusively defined by sensitivity and selectivity. In the 1950s and early 1960s a triple-conversion superheterodyne receiver was a status symbol. It was not until the mid-1960s that strong-signal handling became an important parameter (Ref 250). Ten years ago a transceiver using DSP in the last IF was the latest technological advancement. Today the focus is on performance in the presence of nearby strong signals.

We should understand that it is not necessarily the type of technology used that makes a receiver a good performer. Performance criteria and measured test data are what counts, regardless of the technology.

Let's have a look what is really important for the low band DXer.

1.2. Handling the Weak Signals: Sensitivity and Signal-to-Noise Ratio (SNR)

The nomographs in **Fig 3-1** show the voltage and power relationships of the RF signals found at the receiver input. The table can be used to convert between the many different units used to express signal strength. See also the spreadsheet *Receiver_Levels.xls* that comes on the CD with this book. In the spreadsheet you can change the system impedance (for example, to 75 Ω instead of 50 Ω).

Sensitivity is the ability of a receiver to detect weak signals. The most important concept related to sensitivity is the concept of *signal-to-noise ratio (SNR or S/N)*. Reception will be good or bad depending on the strength of the signal with relation to "the noise." It is generally accepted that comfortable SSB reception requires a 10-dB signal-to-noise ratio. CW reception requires a lower S/N, and any moderately experienced CW operator can rather easily deal with a 0 dB S/N. A really good operator can dig CW signals out of the noise at –10 dB S/N in a 500-Hz bandwidth, mainly because his built-in "brain filter" narrows the noise bandwidth much further. This shows the inherent advantage of CW over SSB for weak-signal communications.

What does "noise" mean in the context of signal-to-noise ratio?

1.2.1. Thermal Noise

The noise present at the receiver audio output terminals is generated in different ways. Internal receiver noise is the noise we hear when there is no antenna but a dummy load connected to the receiver. The internal receiver noise is produced by the movement of electrons in any substance (such as resistors, transistors and FETs that are part of the receiver circuit) that has a temperature above absolute 0 Kelvin (0 K or –273º Celsius). Absolute zero is where all electrons have stopped moving. Above 0 K, electrons move in a random fashion, colliding with relatively immobile ions that make up the bulk of the material. The final result is that in most substances there is no net current in any particular direction on a long-term average, but rather a series of random pulses. These pulses produce what is called thermal-agitation noise, or simply thermal noise.

The Boltzmann equation expresses the noise power in a system. The equation is written as:

$$p = kTB \qquad \text{(Eq 3-1)}$$

where

p = thermal noise power, watts

k = Boltzmann's constant (1.38×10^{-23} joules/Kelvin)

T = absolute temperature in Kelvin

B = bandwidth, Hz

Notice that the power is directly proportional to temperature, and that at 0 Kelvin the thermal noise power is zero.

Expressing equivalent noise voltage, the equation is rewritten as:

$$E = \sqrt{ktBR} \qquad \text{(Eq 3-2)}$$

where R is the system impedance (usually 50 Ω).

For example, at an ambient temperature of 27 ºC (~300 K), in a 50 Ω system with a receiver bandwidth of 3 kHz, the thermal noise power is:

$$p = 1.38 \times 10^{-23} \times 300 \times 3000 = 1.24 \times 10^{-17}\ \text{W}$$

This is equivalent to $10 \log (1.24 \times 10^{-17}) = -169$ dBW or –139 dBm (139 dB below 1 milliwatt), and is equivalent to 32 dB below 1 µV or –32 dBµV (Ref 223). This is the theoretical maximum sensitivity of the receiver under given bandwidth (3000 Hz) and temperature conditions. More sensitivity can

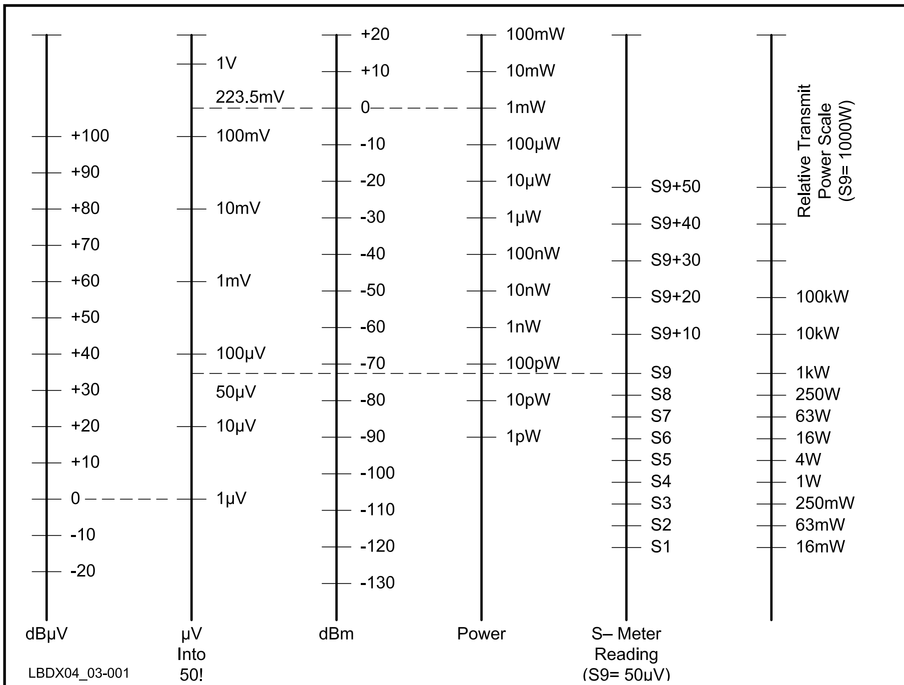


Fig 3-1 — These nomographs show the relationship between receiver input voltages, standard S-meter readings and transmitter output power.

be achieved by reducing the bandwidth or by cooling your equipment. If in this example the bandwidth is reduced to 300 Hz, the noise floor level becomes –149 dBm, simply because there is 10 times less noise power in a window that is 10 times more narrow. We have gained 10 dB by reducing the noise by 10 dB! This explains the big advantage of CW over SSB when signals are weak.

1.2.2. Quantifying Receiver Noise

How do we quantify the noise performance of a receiver? This quality is commonly expressed by the receiver *noise factor (F)* or the receiver *noise figure (NF)*. Both of these are a measure of degradation of the signal-to-noise ratio, caused by components in the RF signal chain.

The *noise factor* is by definition the ratio of the total output noise power to the input noise power when the receiver's input is at the standard temperature of 290 K (27 °C). Being a ratio, F is not denominated. It also is independent of bandwidth, temperature and impedance.

$$F = SNRin / SNRout \qquad \text{(Eq 3-3)}$$

where SNR = signal-to-noise ratio.

The *noise figure* is the logarithmic expression of the noise factor, in dB:

$$NF = 10 \log (F) \qquad \text{(Eq 3-4)}$$

where F is the noise factor.

A spreadsheet calculator (*RX_Noise_Factor_and MDS_calculator.xls*) is available on the CD that comes with this book.

1.2.3. Minimum Discernable Signal (MDS)

MDS is a measure of *sensitivity*. The *minimum discernable signal* (MDS) produces an output that is the same as the internal noise level of the receiver. The value of the MDS is by definition the same as the value of the receiver *noise floor*. MDS can be expressed as:

$$MDS\ (dBm) = -174\ dBm + 10 \log(BW) + NF \qquad \text{(Eq 3-5)}$$

where –174 dBm is the minimum noise level in a 1 Hz bandwidth at 27 °C and BW is expressed in Hz.

Is the MDS a very important receiver parameter on the low bands? Not usually because the level of the input noise to the receiver (atmospheric and manmade noise, see Section 1.2.4.) is relatively high, and generally much higher than the internal noise level of the receiver.

The real acid test as far as receiver sensitivity is concerned is as follows: If you hear a very noticeable noise increase when you connect an antenna to the receiver, your receiver is sensitive enough! You should check sensitivity at the quietest time with the narrowest selectivity you use on every antenna you use. Any more sensitivity will just make it more difficult for the receiver to handle very strong signals (see Section 1.3.4).

A conversion tool (*RX_noise_figure_and_MDS_calculator.xls*) that calculates the level of the receiver internal noise as well as the receiver's noise figure (given the receiver's bandwidth, temperature and MDS) is available on the CD included with this book.

1.2.4. External Noise and Receiver Sensitivity Margin

Besides the noise generated inside a receiver, the factor that usually limits the sensitivity of the overall receiver system on the lower HF bands is the *noise coming from the antenna* (external noise). This noise is mainly atmospheric and/or manmade noise.

Let me quote from W8JI's Web site:

"The noise that limits our ability to hear a weak signal on the lower bands is almost always an accumulation of many signal sources. Below 18 MHz, the noise we hear on our receivers (even at the quietest sites) comes from terrestrial sources. Receiver noise is generally a mixture of local ground wave and ionosphere propagated noise sources, although some of us suffer with dominant noise sources located very close to our antennas.

"In ***urban*** locations, noise arrives from multiple random sources through direct and ground wave propagation from local sources. One or more sources can actually be the induction-field zone of our antennas (in most cases the induction field dominates at distances less than ½ λ). Urban locations are the least desirable locations because typical noise floors average 16 dB higher than suburban locations. There is often no evidence of winter night noise increase on 160 meters, since ionosphere-propagated noises are swamped out by the combined noise power of multiple local noise sources. Much of the noise sources are utility distribution lines, because of the large amount of hardware required to serve multiple users. Other noise sources are switching power supplies, arcing signs, and other unintentional manmade noise transmitters.

"***Suburban*** locations average about 16 dB quieter than urban locations, and are about 20 dB noisier than rural locations. Noise generally is directional, arriving mostly from areas of densest population or the most noise-offensive power lines. Utility high-voltage transmission lines are often problematic at distances greater than a mile, and occasionally distribution lines can be problems. The recent influx of computers and switching power supplies has added a new dimension to suburban noise. There is often a small increase in nighttime winter noise at exceptionally quiet suburban locations. This increase occurs when propagated terrestrial noise equals or exceeds local noise sources.

"***Rural*** locations, especially those miles from any population center, offer the quietest environment for low-band receiving. Daytime 160-meter noise levels are typically around 35-50 dB quieter than urban, more than 20 dB quieter than suburban locations. Nighttime brings a dramatic increase in low-band noise, as noise propagates in via the ionosphere from multiple distant sources.

"Primary local noise sources are electric fences, switching power supplies, and utility lines. I can measure a 3 to 5 dB daytime noise increase in the direction of two population centers, Barnesville (population 7500) and Forsyth (population 10,000) both 10 km from my QTH.

"Typical daytime noise levels, measured on a 200-foot omnidirectional vertical, are around –113 dBm with a 350-Hz bandwidth (noise power is directly proportional to receiver bandwidth). Noise power increases about 5 to 15 dB at night, when the band 'opens.' As in the case of suburban systems, directional antennas reduce noise power. Nighttime is the 'big equalizer,' reducing the advantage of location as distant noises increase with improved propagation."

Table 3-1
160 Meter Manmade and Atmospheric Noise Data

	Daytime Noise Level (dB)	*Nighttime Noise Level (dB)*
Rural	0	5 to 15
Suburban	20	20 to 25
Urban	35 to 50	35 to 50

Measured by Tom Rauch, W8JI.
Total noise reference level 0 dB = –113 dBm in 350-Hz bandwidth.

While we are at it, and as we will need to understand noise and its characteristics in our fight against it, let us get rid of a common misunderstanding which says that noise signals are always vertically polarized. This is incorrect.

W8JI writes: "Noise arriving from the ionosphere is randomly polarized. It arrives at whatever polarization the ionosphere happens to favor at the moment. It has the same ratio of electric to magnetic fields (also called field impedance) as a 'good' signal. Sources within a few wavelengths of the antenna combine and produce a randomly polarized noise. It has *no* dominant field. It can either be electric or magnetic field dominant. Noises arriving from ground wave sources some distance from the antenna are vertically polarized. The fixed polarization occurs because the earth 'filters out' horizontal components. Horizontal electric field components are 'short circuited' by the conductive earth as they propagate and are eliminated."

Some time ago the ARRL Laboratory asked Tom, W8JI, to make some manmade noise-level measurements at his super-quiet QTH, in preparation for ARRL's filing comments about BPL (Broadband over Power Lines). I visited W8JI to find out firsthand just how quiet a rural environment can be on 160 meters. During the daytime I could switch his Beverage antennas and tell from the rise in noise level the directions to the nearest towns about 10 km away.

Tom measured a typical daytime manmade noise level on 160 meters of –113 dBm in a 350-Hz bandwidth. Using a large omnidirectional vertical antenna, Tom has a noise level between S2 and S3. These readings are in the absence of local QRN (static and thunderstorms). W8JI's margin over the receiver MDS (–141 dBm) is thus 28 dB during the day. This assumes that Tom would be using a large vertical for receiving, although this is usually not the case since he usually uses Beverage antennas, which can discriminate against noise coming from other directions. **Table 3-1** summarizes W8JI's noise numbers for 160 meters. On 80 and 40 meters, typical manmade noise levels are lower by about 8 to 9 dB on 80 meters and 17 to 18 dB on 40 meters.

The manmade plus atmospheric noise-level data in Table 3-1 are referenced to a level of –113 dBm (at 0 dB) for the best-case (rural) daytime local manmade noise propagated by ground wave. For this level of noise, a receiver noise figure of 35 dB would be adequate to maintain a S/N of 0 dB in a 350-Hz bandwidth. This means that a 25-dB attenuator could be placed at the input of a typical receiver with a 10-dB noise figure and the desired signal would drop down to the thermal-noise level of the receiver itself. The output S/N would thus still remain more than adequate for good copy on CW.

In quiet areas the total noise will be higher during the night, since additional noise arrives by atmospheric propagation, adding to the local manmade noise found during the daytime. In urban residential areas the local manmade noise is so high that you never can hear such propagated noise. This means that the receiver sensitivity is essentially a moot point. Almost any receiver is sensitive enough in such a hostile environment!

Noise levels can, and do, vary tremendously from one location to another. We have many bad 160 meter locations, and only a few very good ones. Indeed, the difference in manmade noise levels can be up to 50 dB! The figures published by W8JI are the first ones I have seen from an active and knowledgeable radio amateur. Therefore they should be considered very important.

Assuming that under most circumstances all noise is evenly distributed in all directions, a well-engineered special receiving antenna (see Chapter 7) will receive up to 12-15 dB less noise than an omnidirectional antenna. To benefit from this directivity advantage, the receiver will require 12-15 dB better noise figure (more sensitivity). In addition, such directive noise-reducing antennas are usually low output antennas (such as Beverages with typically –10 dBi gain) and often used with long "lossy" feed lines (say, 1 dB loss). Add all of this together, and you need 12 + 10 + 1 = 23 dB more sensitivity (lower noise figure), so a lot of your surplus sensitivity for 160 meters has gone away. Sometimes we may need a preamp (typically with 10-15 dB gain). The use of preamplifiers for low-noise receiving antennas is covered in more detail in the Chapter 7 on Receiving Antennas.

Don't forget that a preamplifier amplifies the atmospheric noise as well as the signal, so the ratio will remain the same. Such preamps cannot amplify only the signal (and not the noise); hence they should only be used to compensate for loss in (very) long feed lines, and not to receive a stronger signal.

1.3. Handling the Strong Signals

On the low bands, more than on the higher frequency bands, receiver sensitivity (MDS) is not the most important characteristic. What is much more important is how well the receiver behaves when trying to copy a weak signal in the close vicinity of a very strong signal or multiple strong signals.

In plain language: What will happen if a signal 50 dB over S9 (approximately –20 dBm) shows up on a frequency near where I am trying to copy a very weak signal, riding on the noise (–120 dBm). Or what will happen if two very strong signals are close to my frequency?

The mechanisms involved are:

- *Blocking* or *gain compression* (covered in Section 1.3.1).
- *Intermodulation distortion* (covered in Section 1.3.2).
- *Reciprocal mixing* with local oscillator noise (covered in Section 1.3.5).

1.3.1. Blocking or Gain Compression

Blocking, also called *gain compression*, is the unwanted effect of one or more *very strong* signals in the close vicinity of a *weak* signal we are trying to copy. Gain compression occurs when a strong signal drives an amplifier stage (for example, a receiver front end) so hard that it cannot produce any more output. The stage is driven beyond its linear operating region and is saturated.

Gain compression can be recognized by a decrease in the background noise level when saturation occurs (Ref 223, 239, 281). Gain compression can be caused by other amateur stations nearby, for example in a multioperator contest environment. Outboard front-end filters are the answer to this problem. (See Section 1.8.)

1.3.2. Intermodulation Distortion

Intermodulation distortion (*IMD*) is an effect caused by two (or more) strong signals that drive one (or more) of the stages in the receiver beyond their linear range, so that spurious (phantom) signals, called intermodulation distortion products (IMD products), are generated because the nonlinear stage acts as a mixer. When these IMD products are stronger than the noise floor in the receiver we shall hear them, and they can (and will) cause interference to weak signals riding near the noise floor.

On the low bands the noise floor is usually set by atmospheric noise rather than the receiver's internal noise, but for generic testing purposes we will always refer to the receiver's own (internal) noise floor.

Third-order IMD (IMD3) is the most common and annoying front-end overload effect. It's called third order because it's the result of mixing $(2 \times f_1) - f_2$ and $(2 \times f_2) - f_1$.

What does this mean in plain language? Imagine we have two equally strong CW signals spaced 1 kHz apart, one at 3600 kHz and another at 3601 kHz. See **Fig 3-2**. As the level is increased, the receiving system shows an increasingly nonlinear response. The second harmonic of 3600 mixes with the fundamental at 3601, and the result is a new signal at $(2 \times 3600) - 3601 = 3599$ kHz. Another signal appears at $(2 \times 3601) - 3600 = 3602$ kHz. The level of the mixing products increases faster than the level of either individual interfering signal. When we can hear the "ghost signal" above the noise floor of the receiver, it adds interference to the weak signals we are trying to hear on that frequency.

Poor IMD performance shows up in a receiver as splatter on SSB and on CW as "bloops, bleeps, and random musical thumps or phantom signals" as W8JI puts it. These phantom CW signals have puzzled me for years during 160 and 80 meter contests, when, in between strong signals, one always seemed to hear undecipherable bits and pieces of CW. Since the better generation receivers (with much better closed spacing IMD performance) became available some years ago, I have never heard these phantom signals again.

In a transmitter, IMD products are generated when multiple tones in an SSB signal mix with one another in a non-linear stage. Such IMD products are very well known by the name of *splatter*.

1.3.2.1. Test Spacing

When we have on-the-air interference problems working weak signals, it is almost always with stations a few kHz or less away. Despite this, until maybe 10 years ago, most manufacturers used to test for blocking and IMD at relatively wide spacing, 20 to 100 kHz. Why care about a test or data at 20 kHz or wider spacing, when the bothersome signals are a few kHz up or down from us? Wide-spaced tests inflate performance, and give us meaningless numbers for real-world performance. If we do an IMD3 test at 20 kHz spacing, we are checking for problems created by two strong signals that are 20 kHz and 40 kHz away from "our" frequency!

It is clear that the frequency separation of the two input signals can greatly influence the IMD and blocking test results. The worst case applies when there is no selectivity in the front end to attenuate one or both signals. This happens when both input signals are within the passband of the first-IF filter and the intermodulation products are produced in the second mixer. Until about six years ago, all commercial receivers were using a high first IF using wide filters (usually 10 to 20 kHz wide) to accommodate narrow-band FM and general HF coverage without requiring filter changes. Such wide so-called "roofing filters" with poor shape factors are like wide-open barn doors where many undesired signals can pass through on their way to the first mixer stage. Serious low band DXers are not interested in doing FM on HF or using their communications receivers for listening to shortwave broadcasts.

It is clear that we need to know the receiver's performance when the interfering signals are close to the wanted signal. The industry standard for "close" spacing has become 2 kHz.

If we test at 2 kHz spacing using a wide first IF filter (eg a 10 kHz wide roofing filter), we will actually measure the "performance" of your second mixer, not the radio's overall performance.

If we test at close spacing (2 kHz) using a narrow first IF (roofing) filter, say 500 Hz, the roofing filter will attenuate the strong parent test signals before they get to the second mixer, and the results will be significantly better (presuming the second filter is functioning properly). Obviously, if the close-spaced performance is good, wide-spaced performance is just as good or better.

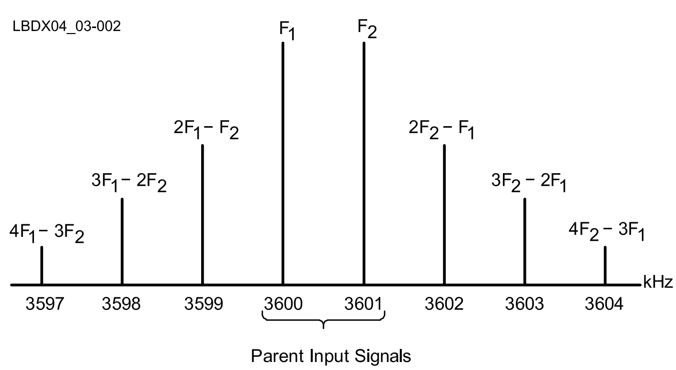


Fig 3-2 — Third, fifth and seventh order intermodulation products generated by parent input signals on 3600 and 3601 kHz. Note that the "order" of a set of IMD products is determined by adding the multipliers for each frequency. For example, $3F_1 - 2F_2$ is 3 + 2 = 5th order. For the best test of receiver performance, *both* test signals should be within the roofing-filter bandwidth. When close-spaced performance is good, wide-spaced performance is just as good or better.

Close-spaced measurements show the real picture, because that is the situation we encounter when operating on crowded bands, especially during contests. This is especially so on the low bands where signals from “local” stations can be very strong. Therefore a top-notch low-band receiver should have a “low” first IF (usually around 9 MHz), where narrow first IF filters (roofing filters) can be used.

The real problem regarding strong nearby signal handling has always been one of *design architecture* of the receiver. Ten-Tec with its Orion transceiver and Elecraft with its K2 were the first manufacturers to tackle this problem. The Elecraft K3 followed soon after. These are all transceivers with only amateur band coverage, using a first IF of around 9 MHz. The Orion and the K3 use a whole bank of switchable roofing filters, where filters as narrow as 200 or 300 Hz can be installed. That’s right — we need the selectivity as close as possible to the antenna. These radios have the best close-in third order IMD performance of all commercially available Amateur Radio receivers at the time of writing (early 2009).

As this edition was being prepared for publication, Yaesu introduced the FTdx5000. This new transceiver also uses an IF around 9 MHz for one of its two receivers, similar to what is done in the Ten-Tec Orion, and that receiver includes switchable roofing filters down to 300 Hz. Receiver specifications are claimed to be similar to the K3 and Orion, with close-in IMD dynamic range around 100 dB. Having a second receiver with a different (higher) IF, true diversity reception will likely not be possible. (The Orion has a similar limitation compared to the K3, which has two identical receivers.) As this was written, we were awaiting independent test reports; and it will be interesting to see how the FTdx5000 performs on the low bands.

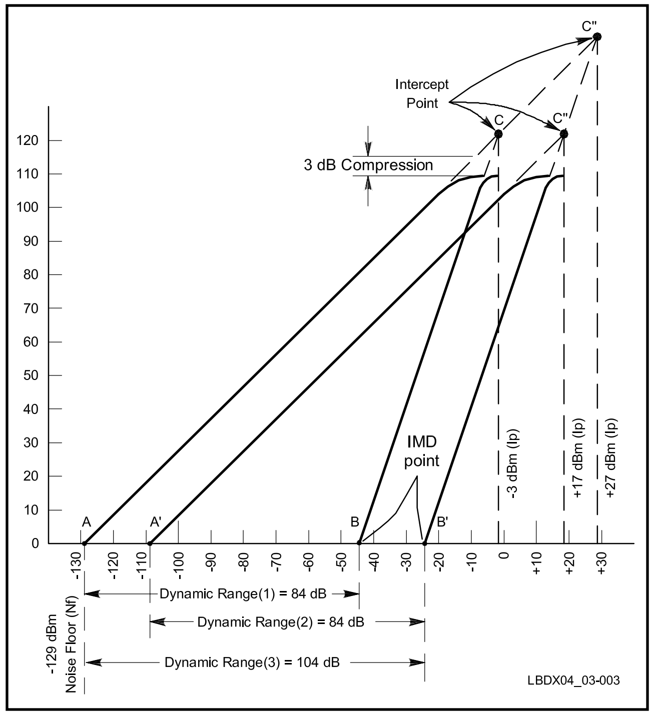


Fig 3-3 — Third-order intercept point showing three examples, with and without 20-dB front-end attenuation. The intercept point increases by the same amount as the attenuation is increased. This is for an average receiver with an 84-dB intermodulation distortion dynamic range (IMD DR) and with a 104-dB dynamic range.

1.3.2.2. Third Order Intercept Point

The third order intercept point (IP3) is the theoretical point at which the extrapolated third order intermodulation level is equal to the signal levels in the output of a two-tone test when the extrapolation is made from a point below which the third order intermodulation follows the third order law.

Fig 3-2 shows the IMD spectrum for an example where the two parent signals are spaced 1 kHz apart. (The third-order products are: $2F_1 - F_2$, and $2F_2 - F_1$.) Third-order IMD products increase in amplitude three times as fast as the pair of equal parent signals (Ref 210, 211, 213, 226, 239, 247, 255, 274, 281). **Fig 3-3** shows three examples of third-order intercept points. The vertical scale is the relative output of the receiver front end in dB, referenced to an arbitrary zero level. The horizontal axis shows the input level of the two equal-amplitude parent signals, expressed in dBm. Point A sits right on the receiver noise floor. Point A' is the floor with a 20-dB attenuator at the receiver input. Increasing the power of the parent signals results in an increase of the fundamental output signal at a one-to-one ratio. Between –129 dBm and –44 dBm, no IMD products are generated that are equal to or stronger than the receiver noise floor for the no-attenuator Case 1. At –44 dBm (point B), the third-order IMD products have risen to exactly the receiver noise floor level.

Point B is called a two-tone IMD point, expressed in dBm. Further increasing the power of the parent input signals will continue to raise the power of the third-order IMD products *three times faster* than that of the parent signals. At some point, the fundamental and third-order response lines will flatten because of *gain compression*. Extensions of both response lines cross at a point called the *third-order intercept point*. The level can be read from the input scale in dBm. The intercept point (IP) can be calculated from the IMD point as follows.

$$\mathrm{IP} = \frac{2 \times \mathrm{MDS\ (noise\ floor)} + 3 \times \mathrm{IMD\ DR}}{2} \qquad \text{(Eq 3-6)}$$

where

MDS = minimum discernible signal or noise floor

IMD DR = IMD dynamic range

By itself, the IP3 of a receiver is totally meaningless. It only becomes meaningful if we consider it together with the receiver sensitivity. This is how we come to the concept of *receiver dynamic range.*

Applying the three examples from Fig 3-3, we find:

Case (1): IP = (2 × –129 + 3 × 84)/2 = –3 dBm. This is for a receiver with an 84 dB IMD DR and no front-end attenuator.

Case (2): IP = (2 × –109 + 3 × 84)/2 = +17 dBm. This is for the same receiver as in Case (1) but with a 20-dB attenuator.

Case (3): IP = (2 × –129 + 3 × 104)/2 = +27 dBm. This is for a receiver with a 104 dB IMD DR and no attenuator.

Conversely, the two-tone IMD point can be derived mathematically from the intercept point.

$$P_{\mathrm{IMD}} = \frac{2 \times \mathrm{IP} + \mathrm{MDS}}{3} \qquad \text{(Eq 3-7)}$$

where MDS = minimum discernible signal or noise floor.

The three examples from Fig 3-3:

Case (1): $P_{IMD} = -(2 \times -3 + -129)/3 = -45$ dBm
Case (2): $P_{IMD} = -(2 \times +17 + -109)/3 = -25$ dBm
Case (3): $P_{IMD} = -(2 \times +27 + -129)/3 = -25$ dBm

What does this mean? For Case (1) this means that two signals below –45 dBm will not create audible IMD products. In other words, two signals at around S9 + 30 dB will start generating audible IMD products. In Europe this is an everyday situation on the 7-MHz band, where signals of –17 to –13 dBm from broadcast stations are common.

The value of the third order intercept point by itself is meaningless. When evaluating third-order intercept points, we must always look at *receiver noise floor levels* at the same time. When we raise the noise floor from –129 dBm to –109 dBm, for example by inserting 20 dB of attenuation in the receiver input line as in Case (2), both response lines and the intercept point will shift 20 dB to the right. This means that the intercept point has been improved by 20 dB, while the *dynamic range* remains the same. An average receiver with a +5 dBm intercept point can be raised to +25 dBm merely by inserting 20 dB of attenuation into the input. Remember that this can frequently be done with present-day receivers, since they often have a large surplus sensitivity. Case (3) shown in Fig 3-3 represents a better solution, where the improvement is obtained by designing the receiver to handle stronger signals before becoming nonlinear.

1.3.3. Receiver Dynamic Range

It is clear that IMD levels and signal blocking levels by themselves don't mean a thing. They have to be interpreted together with another important receiver parameter: MDS (receiver sensitivity).

Receiver dynamic range is the ability of a receiver to receive a weak signal without loss of readability while one or more strong signals are in the vicinity. In other words, IMD is the *range* between on one side the receiver internal noise floor (same figure as receiver MDS, see Section 1.2.2.) and on the other side the power level of the signals at which IMD products appear or signal blocking sets in, that is relevant.

1.3.3.1. Intermodulation Distortion Dynamic Range

The two-tone, third-order intermodulation distortion dynamic range (IMD DR or IM3 DR), also called the *two-tone dynamic range* gives us the *entire picture* as far as intermodulation distortion of a receiver. Refer back to Fig 3-3 for a graphical representation of dynamic range. IMD DR is calculated by:

$$\text{IMD DR} = -(\text{MDS} - P_{IMD}) \qquad \text{(Eq 3-8)}$$

where

IMD DR = intermodulation distortion dynamic range, in dB
P_{IMD} = two-tone IMD point, dBm
MDS = minimum discernible signal or receiver noise floor, expressed in dBm

Example: for $P_{IMD} = -30$ dB and MDS = –130 dB,

$$\text{IMD DR} = -[-130 - (-30)] = 100 \text{ dB}$$

If the intercept point is known instead of the two-tone IMD point we can use the following equation:

$$\text{DR} = \frac{2\,(\text{IP} - \text{MDS})}{3} \qquad \text{(Eq 3-9)}$$

where IP is the intercept point in dBm and MDS is the minimum discernible signal or noise floor, also in dBm.

The dynamic range of a receiver is important because it allows us to directly compare the strong-signal-handling performance of receivers (Ref 234, 239, 255), since it takes into account the sensitivity as well.

The IMD dynamic range is the single most important number when comparing receivers for the low bands. W8JI puts it as follows: "IMD DR is a measure of how badly your own receiver causes problems you might blame on other people."

Some figures: Assume a receiver with an MDS of –130 dBm (in a 500-Hz bandwidth) has a measured IMD DR of 100 dB (at a given spacing between the two offending strong signals). This means that unwanted signals will be detected at the receiver's noise floor for offending signal levels greater than –130 dBm – (–100 dBm) = –30 dBm, which is approximately S9 + 40 dB (see Fig 3-1). In the Lab, the noise in which the intermodulation products are buried is the *receiver internal noise*. In reality, on the low bands, the limiting factor will be the atmospheric noise and not the receiver noise.

Do we need 100 dB dynamic range? Fortunately, not under most circumstances. First of all we seldom will be in a situation where we need the full sensitivity of the receiver on the low bands, because of the atmospheric noise. If we use special receiving antennas such as Beverage antennas (gain –10 to –15 dBi), the received signals will be weaker. Tom, W8JI says that an IMD DR number above 80 dB is enough to stay out of trouble 99% of the time. If you are in a noisy location, you obviously need less performance, because the IMD products will be buried in the atmospheric noise rather than in the receiver noise.

All I can say is that when I switched from a radio with a 2 kHz IMD DR of approximately 70 dB to a radio with more than 90 dB, it made a day and night difference when operating on a very crowded 160 or 80 meter band during CW contests. Those phantom signals near the noise floor are gone. It's much harder for me to tell the difference on SSB, as most of the IMD products we hear during contests are caused in transmitters (splatter) rather than in receivers.

1.3.3.2. Blocking Dynamic Range (BDR)

We described the phenomenon of blocking or gain compression in Section 1.3.1. As with IMD, we should really look at the signal range a receiver can handle without causing blocking. Blocking is usually only an issue if you have another ham at line of sight distance from you, or if your station operates in a multi-multi contest environment.

The blocking dynamic range (BDR) is the difference between on one side the receiver internal noise floor and on the other side the power level of the signal that causes blocking. BDR can be calculated by:

$$\text{BDR} = -(\text{MDS} - P_B) \qquad \text{(Eq 3-10)}$$

where

BDR = blocking dynamic range, dB

P_B = single tone blocking level, dBm

MDS = minimum discernible signal or receiver noise floor, dBm

It is clear that, just as with IMD DR, the BDR will be very dependent on signal spacing. Whether or not the interfering signal falls inside or outside the passband of the roofing filter will make a day and night difference (see Section 1.3.3.). This means that only the close-spaced test figures (2 kHz or closer spacing) are meaningful.

How much BDR do we need on the low bands? Here's what Tom, W8JI writes on his Web site: "The number you want here is probably around 80 dB or more if you live in a reasonably quiet location and work weak signals on crowded bands. If you run two transmitters on the same band or have a neighbor who operates near your frequency, you almost certainly need more dynamic range. I'm in a very quiet rural location and have very directive antennas, and 80 dB blocking DR suits my requirements just fine most of the time."

1.3.4. Intermodulation Created Outside the Receiver

If you hear what sounds like spurious signals from a broadcast station in the ham bands, one way to tell if the product is occurring in your receiver is to insert an attenuator at the input of the receiver. If you observe a much greater drop in the garbage level than in the desired signals when attenuation is added, then you can bet the garbage comes from overload of your own receiver. If both the desired signals and the spurious signals drop the same amount with attenuation, then the generation of spurious signals is happening outside your receiver.

I have witnessed this problem with aging Beverage antennas. Sometimes it is referred to as bad ground loops or even bad connections, but nonlinearity caused by corrosion can create overload, cross modulation and intermodulation (in plain language: *mixing*). You need good connections in the system, even if you don't normally run transmitter power into a receiving antenna such as a Beverage. If you suddenly hear all kinds of alien signals pop up in the band where they don't belong, it's time to go and check all the contacts in your receiving antenna system. Also check proper grounding of the coaxial feed line.

Such products can occur in poor electrical connections in cable TV (where aluminum cable is in contact with steel support strands), telephone wires, fences, towers and even in your own antennas. The mixing products are radiated by these inadvertent antennas into your receiving system. With broadband antennas such as Beverages, you may need to use high-pass filters or preselectors when operated in the vicinity of broadcast stations (as mentioned in Section 1.8).

1.3.5. Local Oscillator Noise

Each superheterodyne receiver has at least one local oscillator. This oscillator allows you to "tune" the receiver.

Oscillators are mostly thought of as single-signal sources, but this is never so in reality. All oscillators have *noise sidebands* to some degree, caused by small phase and frequency variations (jitter) of the signal. Crystal oscillators have a very low level of phase noise; free running LC oscillators normally perform very well also. Frequency synthesizers, especially those built around phase locked loops, are critical to design in this respect and intrinsically produce more noise sidebands.

The quantity of *phase noise* is usually expressed in units of dBc/Hz, which means the noise power below the carrier in a 1 Hz bandwidth. In other words: a 1 W carrier with a phase noise power of –100 dB/Hz emits a wide noise spectrum of –100 dBW in 1 Hz and if we consider a bandwidth of 2 kHz we have to scale the value using the factor –10 log (2000) or –33 dB. This means that we have –67 dBW of noise power in 3 kHz, which is the same as –37 dBm. That's a pretty poor figure.

In recent years very significant advancements have been made in commercial equipment to design and use LOs with very low noise. The Ten-Tec Orion and Elecraft K3 achieve –130 to –140 dBc/Hz. Many older transceivers yield figures between –105 and –120 dBc/Hz. All these figures are measured at 10 kHz spacing (source: **www.sherweng.com**).

It is clear that to be complete one has to specify *where* this 1 Hz is with respect to the carrier. (With some type of oscillators the phase noise is much higher close to the carrier than far away.) In most cases the noise measurement is a swept frequency measurement showing results from a few Hz to a few tens of kHz away from the oscillator frequency.

The major effect of local oscillator noise is so-called reciprocal mixing in the receiver, which reduces ultimate selectivity, sensitivity and blocking dynamic range, as well as IMD dynamic range.

1.3.5.1. Reciprocal Mixing

In a superheterodyne receiver the main effect of local oscillator noise is known as *reciprocal mixing*. This occurs when the noise sidebands of the local oscillator (LO) mix with strong signals that are close in frequency to the wanted signal, producing unwanted noise products at the intermediate frequency and degrading the receiver sensitivity.

To look at how this occurs, take the case of a superheterodyne receiver tuned to a strong signal. The signal will pass through the receiver mixer, where it will be mixed with the local oscillator to produce the IF signal going into the IF filter. When the local oscillator is tuned away so that the strong signal we initially received now falls outside the passband of the IF filter, this strong signal will no longer be heard. However it will still be possible for the phase noise off the side of the local oscillator to mix with the strong incoming signal to produce *noise signals* that will fall inside the receiver IF passband. If the level of these noise signals exceeds the level of the noise floor it will limit the receiver's sensitivity, and weak signals may be masked.

Phase noise contributes to reduced sensitivity as well as indirectly to poor receiver BDR and IMD DR through the effect of desensitization (lower MDS).

Looking at it in another way: Assume you have a carrier (CW) outside the passband of the IF filter (this means you are not tuned to that signal). The strong carrier (CW) will mix with the noisy local oscillator (LO) and produce noise inside the passband of the filter. If the LO noise is strong enough, the noise inside the filter passband will be stronger than the noise floor of the receiver. If we now tune the receiver a few kHz to the left and/or to the right, still keeping the strong carrier (CW) outside the passband of the receiver, the noise due to reciprocal mixing will always remain inside the filter (because the wide band phase noise spreads on both sides of the LO frequency). This effectively reduces the selectivity of the receiver. In other words, if the static response of the IF filters is specified down to –80 dB, the noise in the LO must be down at least the same amount in the same bandwidth in order to not degrade the effective selectivity of the filter.

Let's take the example of a filter with a pass band of 2 kHz. According to the thermal-noise equation (Eq 3-1), the noise power is –174 dBm at room temperature for a bandwidth of 1 Hz. The noise in an SSB bandwidth of 2 kHz can be scaled to a 1-Hz bandwidth using the factor 10 log (2000) which means subtracting 33 dB. If we have a receiver using a local oscillator where the phase noise is 130 dB/Hz (eg at 2 kHz, which means inside the passband of a SSB filter) the effective ultimate stopband of the IF filter is 130 – 33 = 97 dB. If we look at CW selectivity (eg 300 Hz), this floor becomes 105 dB. These are very respectable figures. But if the LO was noisy (eg 100 dB/Hz), as we have seen them years ago when the first transceivers with phase locked loop oscillators came about, the ultimate effective stop band rejection will only be 100 – 33 = 67 dB, which is quite poor!

Fig 3-4 shows the levels of the noise sidebands produced by a specific local oscillator. These are the close in values. At greater distances these values usually remain constant.

The phase noise limited dynamic range is a function of the

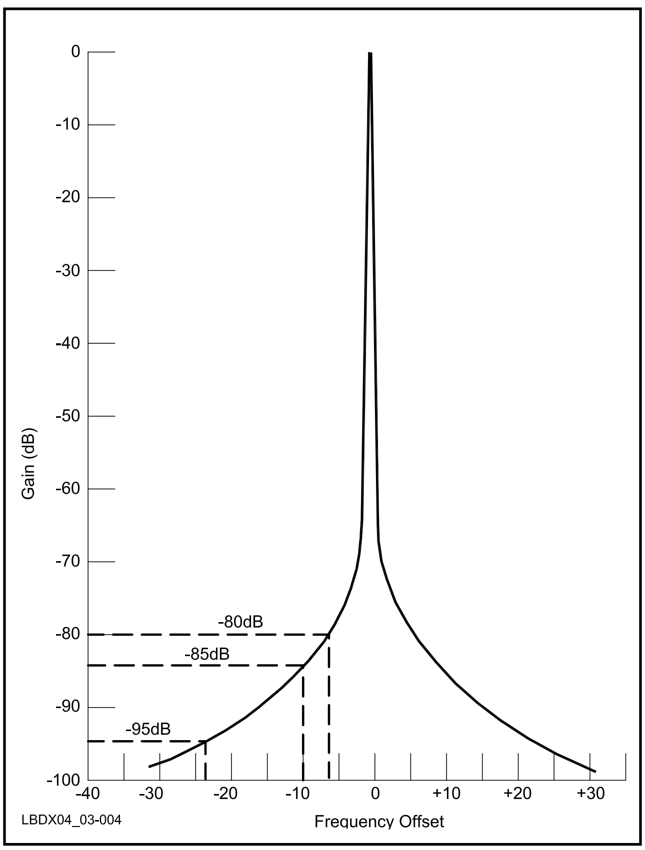


Fig 3-4 — Output spectrum of a voltage-controlled oscillator. If the measurement is made in a 3-kHz bandwidth, the oscillator sideband performance referenced to a 1-Hz bandwidth is (–85) + (–34) = –119 dBc/Hz (dB referenced to the carrier per Hz).

cleanliness of the frequency synthesizer or local oscillator(s). In conjunction with the selectivity-defining filters, it determines performance with respect to strong off-frequency signals.

Peter Chadwick, G3RZP, wrote an excellent article in *NCJ* (Ref 187) explaining that phase noise performance can, when the band is loaded with many very strong signals, be more problematic than IMD dynamic range or blocking dynamic range. The problem with reciprocal mixing is that the noise generated by various strong signals all adds up as Peter explains.

1.3.5.2. The Pitfalls of Many AGC Systems

Automatic gain control (AGC) systems amplitude modulate signals in the receiver chain. On a "dead" frequency (only noise, noise which does trigger the AGC), the noise envelope will modulate noise, and create extra noise. It is a form of intermodulation distortion. Most avid Top Band DXers have been aware of that, and switch off the AGC, and manually adjust the RF gain of the receiver to eliminate this phenomenon. This problem is inherent to almost all AGC systems, and the only cure so far was to apply the technique described above. Watch out, simply switching the AGC off and manually adjusting the RF gain may kill your ear drums if a strong signal pops up. Fortunately the Elecraft K3 has an adjustable audio limiter, which prevents this from happening.

To avoid most of these problems, on the low bands I almost always reduce the RF gain to the point that the S-meter moves up to the peak value of the noise (in winter time the background noise on Top Band typically varies between S1 and S3 on my Beverages — without a preamplifier outside the receiver). In such a case the AGC (which I normally have set for "fast") would really only start working on signals stronger than the background noise. Another solution is to switch to slow AGC, with a hold time of 100 to 300 ms. This emulates the constant AGC level that I set with manual control. The problem with the slow AGC is that the decay after the 50-300 ms hold is usually very slow, and when strong signals are in the passband they will greatly reduce the sensitivity during the long decay time.

Elecraft has been aware of this problem (AGC creating "havoc" in the receiver), and has developed a novel AGC system that solves the above mentioned problems in their K3. This new approach makes it possible to reduce the in-band IMD by 15-25 dB without having to resort to manually adjusting the gain of the receiver.

1.4. Cross Modulation

Cross modulation occurs when modulation from an undesired signal is partially transferred to a desired signal in the passband of the receiver. Cross modulation starts at the 3-dB compression point on the fundamental response curve as shown in Fig 3-3. Cross modulation is independent of the strength of the desired signal and proportional to the square of the undesired signal amplitude, so a front-end attenuator can be very helpful to reduce cross modulation. Introducing 10 dB of attenuation will reduce cross modulation by 20 dB. This exclusive relationship can also help to distinguish cross modulation from other IMD phenomena (Ref 223, 247).

1.5. Selectivity

Selectivity is the ability of a receiver to separate (select) a desired signal from unwanted nearby signals. For weak signal conditions narrow filters also reduce the noise power contained in the window of the filter, and thus a narrow filter improves the S/N ratio.

1.5.1. Selectivity for SSB Operation

On a quiet band with a reasonably strong desired signal, the best sounding audio and signal-to-noise ratio can be obtained with a filter selectivity between 2.1 and 2.7 kHz (at –6 dB). Under adverse conditions, selectivities as narrow as 1.5 kHz can be used for SSB, but the carrier positioning on the filter slope becomes very critical for optimum readability. The ideal selectivity for SSB reception will of course vary, depending on the degree of interference on adjacent frequencies.

1.5.2. Selectivity for CW Operation

There are two schools of thought in this area: those who swear by the narrowest possible bandwidth and those who like to keep it wide (500 Hz or even more). We should remember we use filters with high selectivity not only to discriminate against other nearby signals, but also to reduce the noise level (noise power). On the low bands we are confronted mainly with propagated noise (thunderstorm QRN, clicks, etc). This is very different from the EME guys, who use really narrow filters to dig for weak signals in a different type of noise: white galactic noise.

I have talked to many low banders, and certainly a large majority of operators prefer a relatively wide filter (typically 500 Hz). They let their brains do the required signal processing. Since I use a radio with continuously adjustable bandwidth (final selectivity obtained by DSP), I have the bandwidth usually set at 200 to 300 Hz bandwidth for normal CW work. When I am running a pileup, I may go to 500 Hz. Properly designed DSP filtering does not introduce any ringing effect, so you take full advantage of a narrower bandwidth, and reduce the noise power in the passband to minimum. I find it necessary to be able to adjust the bandwidth or move the filter slopes in 10 Hz steps, especially when we are dealing with bandwidths of less than 500 Hz.

Tom, W8JI, wrote: "I use 250-Hz filters when the band is quiet with only white noise, and 600-Hz filters when there is QRN or 'rough' noise. A wider filter always works better when there are static crashes. I do have receivers with very fast, very good AGC systems, and they work very well during static crashes with AGC on, but I still find that wider selectivity helps. Wider selectivity helps because the sharp waveform of the static crash is not lengthened and blurred, and so my ears can do a better job of filtering the noise from the signal."

George, W2VJN adds: "The narrower a filter is, the more concentrated the noise is at the center frequency of the filter. In a noisy environment this can make a weak signal harder to read as the noise pulses are stretched by the filter bandwidth. A wider filter does much less stretching and no concentrating of the noise around the signal frequency. Since the human response can be as narrow as 50 to 100 Hz, nothing much is lost by opening the receiver bandwidth up."

1.5.3. CW Sidetone Pitch

In all modern transceivers one can adjust the CW sidetone pitch to suit personal preference (see also Chapter 2, Section 6 on Zero Beat). Over the years avid CW operators may have developed a sensitivity notch in their hearing, whereby certain frequencies are suppressed quite a number of dB. I, for one, have developed such a problem around 450 Hz. It often suffices to move 10 or 20 Hz off the side to solve this notch problem. Equipment designers should take this into account and make the sidetone pitch adjustable in steps of 10 Hz or less.

1.5.4. Passband Tuning

Passband tuning (IF shift) allows the position of the passband versus the BFO to be altered without requiring that the receiver be retuned. The bandwidth of the passband filter remains constant, however. In some cases interfering signals can be moved outside the passband of the receiver by adjusting the passband tuning.

In better receivers a combination of passband tuning and continuously variable bandwidth is available.

1.5.5. Continuously Variable IF Bandwidth

Before the introduction of DSP in the last IF (usually at a frequency between 10 and 50 kHz) continuously variable selectivity was achieved by moving the passbands of two filters (at different IFs), one across the other. By adjusting the position of the BFO versus resulting passband, one could change the pitch on CW or the passband position on SSB (favoring either the lows or the highs in the audio spectrum). The controls to do so were commonly called *width* and *passband.*.

In this system, producing a continuously variable bandwidth involves passing the signal through two separate filters at two different IFs (such as 9 MHz and 455 kHz). The mixing frequency is slightly altered so the two filters do not superimpose 100%, but have their passbands sliding across one another. You must understand, however, that a variable bandwidth system such as this can never have as good a shape factor as individual well-shaped crystal filters, since the shape factor always worsens when you narrow the bandwidth in this fashion.

Most transceivers developed after around 2003 use DSP techniques at a very low IF to obtain the final selectivity. If used in conjunction with narrowband roofing filters after the first mixer, this is a very good solution. These DSP filters can achieve very narrow bandwidth (50 Hz and less) without exhibiting any ringing. The ultimate solution would be a continuously adjustable roofing filter bandwidth as this moves the selectivity as close to the antenna as possible.

1.5.6. Filter Shape Factor

The filter *shape factor* is expressed as the ratio of the bandwidth at –60 dB to the bandwidth at –6 dB. Good filters should have a shape factor of 1.5 or better. This 1.5 figure is a typical shape factor for an 8-pole crystal filter. Too many transceivers are equipped with rather wide SSB IF (hardware) filters (typically 2.7 kHz at –6 dB) with mediocre skirt selectivity. On a quiet band these give nearly hi-fi quality, but is this what we are really after as low band DXers? For the average operator this may be an acceptable situation, although the serious DXer and contest operator will want to go a step further.

International Radio (Inrad) offers modification kits for modern transceivers. See **www.inrad.net**. See **Fig 3-5** for a comparison of a stock mechanical filter and an 8 pole crystal filter from Inrad. When not properly engineered, narrow (CW) filters with very steep skirts can often yield poor group delay figures, resulting in ringing (see Section 1.5.7).

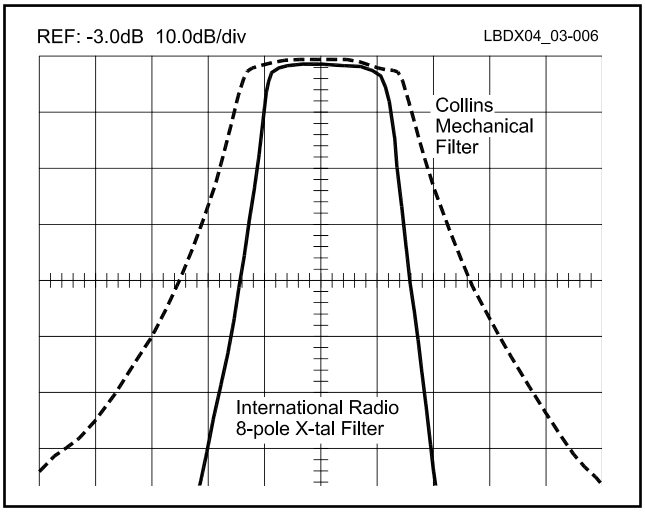


Fig 3-5 — Selectivity curves for typical 455-kHz IF, 500-Hz passband Collins mechanical filter and the 400-Hz crystal filter offered by International Radio. Notice that at –60 dB the replacement crystal filter has only half the bandwidth of the stock mechanical filter.

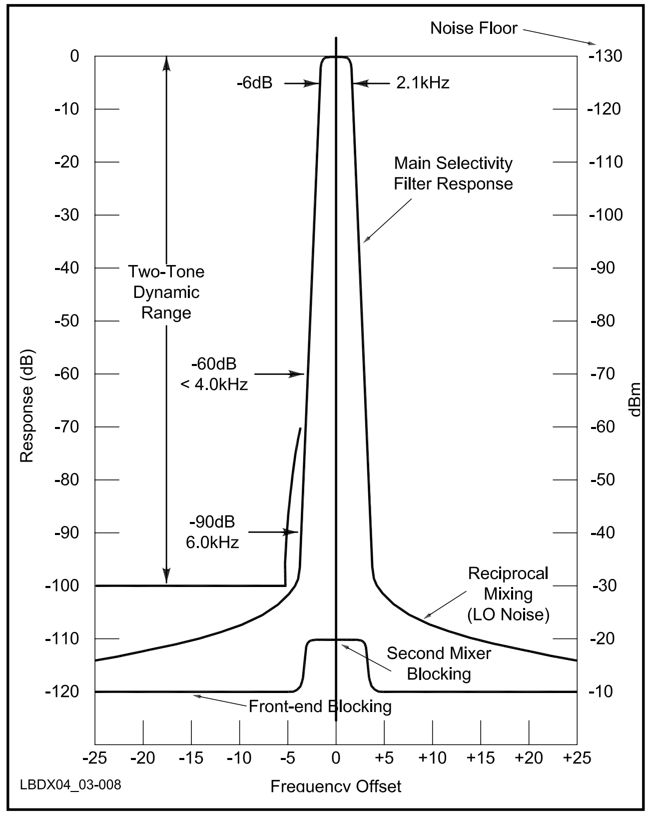


Fig 3-6 — Merit graph for a "dream receiver," which was first published in 1987 in the First Edition of this book. As this is written in 2009, only the Ten-Tec Orion and the Elecraft K3 have met this merit graph with a two-tone 3rd-order dynamic range of 100 dB and sharp selectivity skirts.

1.5.7. Static and Dynamic Selectivity

The *static selectivity* curve of a filter is the curve we measure on a network analyzer. It shows the performance of the filter by itself. The *dynamic selectivity* of a receiver is the combination of the static selectivity and the effects of reciprocal mixing (see Section 1.3.5.). Note that the static selectivity can be degraded significantly by the effect of reciprocal mixing.

If the amplitude of the reciprocal mixing products is greater than the stop-band attenuation of the filter, the ultimate stop-band characteristics of the filter will deteriorate. State of the art frequency-synthesizer designs yield a noise output that is down 110 dB (measured in 3 kHz bandwidth) or –144 dBc/Hz, at a 10-kHz offset. With such an excellent low noise local oscillator one can take advantage of a filter with ultimate rejection of over 100 dB. If however, the noise of the LO is only 80 dB down (114 dBc/Hz) it is senseless to use an excellent IF filter with a 100-dB stop-band characteristic.

Hart, G3SJX, uses an interesting graphical representation of the main receiver parameters. **Fig 3-6** shows the characteristics of a "dream receiver" using Hart's graph technique. Fig 3-6 was first published in 1987. The specifications for such a high-performance receiver, which was a dream rather than reality more than 20 years ago, now read:

- 3rd order IMD dynamic range: ≥100 dB.
- Noise floor: MDS ≤–130 dBm (500 Hz bandwidth).
- Blocking dynamic range: ≥110 dB.
- LO sideband noise performance: Better than –135 dBc at close spacing (2 kHz).

It took almost 20 years before we saw a couple of transceivers on the market that just about reach these requirements. When this was written in 2009, these are the Orion from Ten-Tec and the K3 from Elecraft. (Watch for reviews to see if the new FTdx5000 meets the requirements as well.)

1.5.8. IF Filter Position in the Receiver Chain

The filter providing the bulk of the operational selectivity can theoretically be inserted anywhere in a receiver between the RF input and audio output. However, when considering parameters other than operational selectivity, it is clear that the filter should be as close as possible to the antenna terminals of the receiver. In Section 1.3 we saw that narrow filters immediately after the first mixer (the "roofing" filters) will help reduce IMD products and blocking effect.

Transceivers that cover all HF frequencies (1.5 through 30 MHz) need to use a first IF above 30 MHz (usual choice is between 45 MHz and 75 MHz). As explained in Section 1.3 it is beyond the state of the art to make narrow filters with a reasonably steep skirts on such high frequencies. In addition the designers of these transceivers usually want to cover all modes (including FM) with a single first IF filter. In general, we find rather simple 2-pole crystal filters with a nominal selectivity of 15 to 20 kHz (at –6 dB) in the first-IF chain. This means a lot of compromising as far as the dynamic range of the receiver for close in signals is concerned.

The Ten-Tec Orion and Elecraft K2 and K3 use a first IF around 9 MHz (didn't we old timers do that 40 years ago with the McCoy 9-MHz filters?). Both the Orion and the K3 provide for a series of switchable roofing filters, ranging from 250 Hz all the way up to 20 kHz. As a consequence, these radios are not general-coverage receivers, but who cares? The Orion has a second receiver built in with the traditionally high first IF, which is general coverage but with reduced IMD dynamic range specifications. The K3 has an option for a second receiver that is identical to the first receiver, using separate roofing filters. That is one of the main differences between the Orion and the K3. These top-of-the-range transceivers for low band DXers of course use "last IF" DSP for obtaining the final operational selectivity (see Section 1.5.10).

When in the previous edition of this book I wrote "I am convinced that somehow the designers have to move away from these very high first IFs, and compensate for the loss of image rejection by going back to tuned input circuits, or narrowband filters (not octave filter) covering just the amateur bands." Well, it has happened. We low band DXers are happy to see transceiver design moving in that direction.

1.5.9. Switchable Sideband on CW

Switchable CW sidebands is a very useful feature that was introduced in the Kenwood TS-850. Nowadays this feature is available in all the "better" transceivers. The user can switch CW reception from lower sideband to upper sideband, just like in SSB.

Although the terminology of lower and upper sideband is not so common on CW, CW signals are indeed received with the beat oscillator frequency either above (as an LSB signal) or below (as a USB signal). This feature can be quite handy in the daily fight against QRM. Together with bandpass tuning, sideband switching can often move an offending signal down the skirts of your filter to a point where no harm is done.

The default on commercial transceivers is usually for CW reception to be LCW (lower sideband). On LCW, when you tune your dial clockwise, the audio pitch goes up. Note: I have always preferred the opposite as the higher pitch is quicker to get my attention.

1.5.10. Digital Signal Processing (DSP)

All current transceivers use DSP, some more than others. One of the nice things about so many features being implemented in DSP in software in modern transceivers is that the software that controls the DSP is upgradeable at all times. No need to send in your transceiver for an upgrade, just download the latest software version and you are all set.

In the mid 1990s, manufacturers started putting their DSP circuits in the last IF of the receiver, mostly around 10 kHz. This very low IF is still necessary today, since CPUs operating at frequencies high enough to allow operation at much higher IFs are either still too expensive or still under development.

The newest transceivers do many functions in DSP: operational bandwidth filtering, noise reduction, automatic notch, AGC, detection, noise reduction etc. Using narrow roofing filters right after the first mixer in combination with IF DSP can give the best of both worlds: very high dynamic range at close signal spacings and utmost flexibility regarding operational bandwidth.

Most transceivers nowadays use DSP filters for obtaining the "final" receiver selectivity (ie, bandwidth filtering appropriate for the mode in use). However, multiple pole crystal filters are widely used as first IF roofing filters, and ringing or pulse widening can be caused in those roofing filters. Ten-Tec seems to be conscious of group delay problems in the Orion receiver, judging from the fact that they advise their Orion users *not* to use the 250 or 500-Hz roofing filters when digging for weak signals in static crashes.

In digital filters as well as hardware filters, an effect called the *Gibbs phenomenon* causes an overshoot (or "ringing") at simple discontinuities (for example, the make and the break of a CW signal, the make and the break of a sharp noise pulse). The only way to eliminate ringing caused by this phenomenon is to increase the shape factor or filter bandwidth. In the Elecraft K3 you can switch from IIR (Infinite Impulse Response) filter design to FIR (Finite Impulse Response) when using bandwidth of 100 or 50 Hz, in order to minimize ringing.

With DSP filters we can now continuously vary the filter bandwidth. Now, what is the optimum bandwidth on CW? Alex, VE3NEA, wrote: "for a CW signal, the optimal bandwidth in terms of signal-to-noise ratio in white noise is 1.5 × WPM — eg, 45 Hz for 30 WPM. Under heavy QRM, even narrower filters should be used. You will have a bit lower signal-to-noise ratio but much better signal-to-interference ratio."

This means that if we have properly designed DSP filters we can use bandwidths down to 50 Hz or less. I remember having a perfect copy on VP6DX on 160 one morning when he was deliberately jammed by a station about 30 Hz off his frequency. Bringing the bandwidth down to 20 Hz (in VE3NEA's *Rocky* SDR software), I was copying VP6DX 100%.

1.6. Noise Blanker and Noise Reduction

A noise blanker (NB), by nature of its principle of operation, only works on short-duration type pulses, such as ignition type pulses). Noise blankers detect strong noise pulses and block (gate) the receiver's IF chain when these pulses are present. To detect these short pulses, we use wide roofing (first IF) filters, because most narrow filters (with group delay problems) lengthen and distort the noise pulses and make noise blanking impossible. Noise blankers are one of the reasons why until recently transceivers used very wide (much too wide) first IF filters, leading to poor IMD performance on strong nearby signals. As the noise pulses are detected on our receiving frequency, rather than on any other frequency outside the busy amateur bands, noise blankers usually are ineffective when the band is fully loaded, such as during contests. Strong adjacent signals can gate the receiver, instead of noise pulses. Using a frequency outside the amateur bands to sense the noise, as was done in the Collins KWM-2 receiver more than 40 years ago, would be a solution to that problem.

In the past decade some top-end transceivers, such as the original Yaesu FT-1000D series, had a design flaw in the noise-blanker circuitry that reduced the IMD performance significantly. The modifications to overcome this flaw in the FT-1000D are described on W8JI's Web site (**www.w8ji.com**).

In the FT-1000MP and MkV this problem exists when the NB gain pot is not set to minimum. In the FT-1000MP, if the gain control is set to CCW the NB has no effect on the receiver IMD. In the MkV, if the menu is set to A=1 and B=1 (under software control), there is no IMD added by the noise blanker. In the newer FT-2000 the noise blanker problem has been solved.

State of the art transceivers such as the K3 are now using very effective and efficient noise blankers that operate both in the DSP part of the receiver (at low IF) as well as in the first IF right after the first mixer.

1.7. Software Defined Radio (SDR)

In a *software defined radio* (*SDR*), functions that have typically been implemented in hardware (such as mixers, filters, amplifiers, modulators/demodulators or detectors) are implemented using software. While the concept of SDR is not new, the rapidly evolving capabilities of digital electronics are making many processes that were once only theoretically possible practical.

1.7.1 Hardware

Let's keep it simple and limit the discussion to the receiver section of an SDR for the moment. When we think of an SDR, we envision a PC equipped with a sound card and some hardware. SDR, for most of us, is a little hardware box and a "large" PC. This can be completely correct. The newest high-end SDRs however do not require a PC sound card for signal processing and have DSP hardware built into the radio. Some examples are the current FlexRadio transceivers and Perseus, SDR-IQ and SDR-14 receivers.

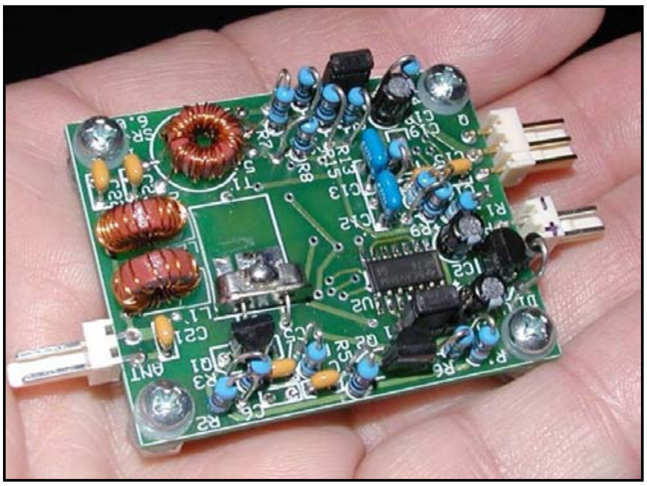


Fig 3-7 — Softrock V6.0 hardware used by the author together with high-end sound card and the *Rocky* and *CW Skimmer* software from VE3NEA. Performance is simply "fantastic."

KB9YIG developed what we can call the "minimum" hardware for a single band SDR receiver, the Softrock SDR board. The entire circuit contains but three transistors and four ICs (see **Fig 3-7**). A number of versions of the basic design have been built all over the world. At the same time a number of software programs were written to support SDRs using a classic PC sound card as the core of the digital operations and the interface to the analog world. The evolution in both hardware and software in this world of SDRs is so breathtaking that by the time this book will be available, things that I describe may be old fashioned.

The little Softrock board is connected to the antenna on one side (input) and to the stereo input of your sound card on the other side (output). And yes, there also is a connection to a 12 V supply. The secret of good performance is in the sound card. Most of the built-in sound cards are mediocre performers. You will need a good sound card with 96 or preferably 192 kHz sampling rate (such as Delta 44, Creative Labs E-MU 0202 or equivalent) if you want good performance. These boards are a little expensive but worth the effort.

FlexRadio Systems (**www.flex-radio.com**) offers a line of complete SDR transceivers. In January 2009 I did a survey among over 2000 active low band DXers (see Chapter 2, Section 20). To my amazement only four out of over 400 stations (less than 1%) that responded use a FlexRadio transceiver.

1.7.2 Software

One of the better and most successful SDR programs is no doubt *Rocky*, a freeware program written by Alex, VE3NEA, (**www.dxatlas.com/Rocky/**). Another is *PowerSDR*, originally developed for the FlexRadio SDR transceiver, which is open source software that also provides an excellent human interface. It works with all of the FlexRadio SDR transceivers as well as with the Softrock hardware and other SDR receivers.

If all of this this is new for you, you're in for a unique experience. When you think of a receiver, you probably think of *listening* for signals. With SDRs you may first be *looking* for signals, rather than listening. The human interface between the SDR receiver and you is the screen of your PC, as well your mouse (and maybe your keyboard, if you prefer). Now, watch and listen. I said "watch" first, because you will be able so "see" weak signals before you "hear" them by just casually tuning across the band. This

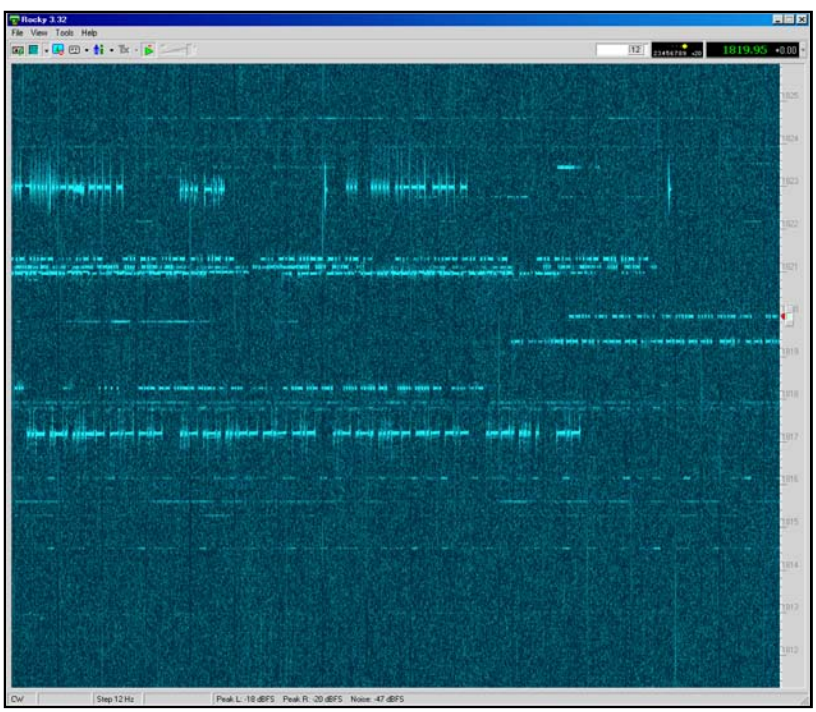


Fig 3-8 — VE3NEA's *Rocky* software in "waterfall" display, in this example showing the spectrum between 1812 ands 1825 kHz. Look at the clean signals and the slicks of others! No other display gives you this much information.

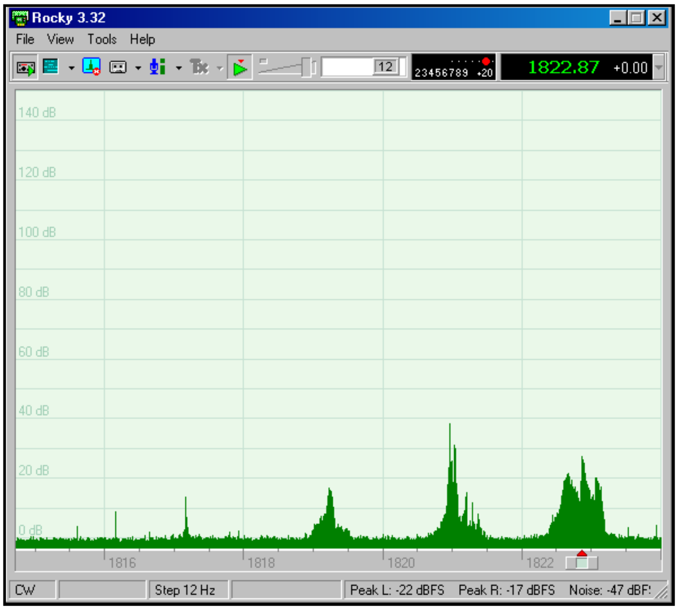


Fig 3-9 — The spectral display for the *Rocky* software With the proper sound card a dynamic range of >130 dB can be reached.

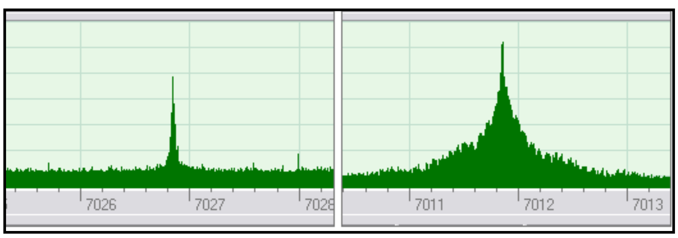


Fig 3-11 — Left: a clean, well shaped CW signal with minimum bandwidth; at right: a CW signal with heavy key clicks occupying three times as much spectrum. All of this can readily be seen on the *Rocky* band scope display. Note that the signal on the left (approximately S7) is approximately 13 dB weaker than the signal to the right (slightly stronger than S9), but at 35 dB below signal peak CW. The signal on the left appears to take less than 150 Hz while the signal to the right takes approximately 400 Hz at –35 dB. With the signal on the left the wider sideband pedestal is hidden by the band noise. The signal on the right with a bandwidth of approximately 1.4 kHz is a very broad signal. (Compare the width of this signal with the SSB spectra shown in Fig 3-22.)

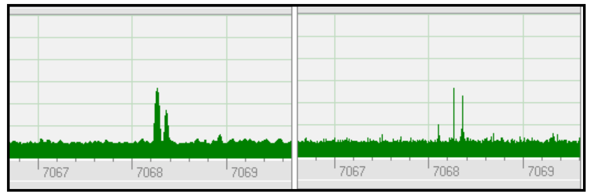


Fig 3-10 — Left: *Rocky* spectral display using the standard FFT algorithm. To the right a similar display using the polyphase algorithm. The latter display has significantly better resolution and allows for detecting of very weak signals "in" the noise.

is a unique experience. Weak signals are easily detectable on the waterfall display where they show up as a thin low contrast (broken) line. Once you see that line, you place your cursor on it, click, and you will hear the weak signal. If you were casually tuning the band you likely would have tuned across the signal without hearing it!

The SDR software provides you with both a *time domain* display (*waterfall* display, see **Fig 3-8**) and a *frequency domain* display (spectrum analyzer or band scope mode, see **Fig 3-9**). VE3NEA's *Rocky* software uses a *polyphase* FFT algorithm to calculate the displayed spectrum. **Fig 3-10** shows the difference between the standard FFT and polyphase FFT: the latter has much higher resolution and does not introduce extra spectral leakage, resulting in crisp and sharp spectrum displays. **Fig 3-11** shows how *Rocky* can be used to evaluate the quality of received CW signals.

These SDR receivers are much more than super spectrum displays. *Rocky* uses a fantastic AGC system, and you really need to look at the S-meter to know whether the signal is S2 or S9++; you simply cannot hear the difference!

1.7.3 Using SDR Software with Other Transceivers

Larry, N8LP, developed LP-PAN (**Fig 3-12**), a pan-adapter interface for the Elecraft K3 transceiver (**www.telepostinc.com**). LP-PAN is a dedicated SDR designed to use the 8.215 MHz IF output from the K3. The specs of the LP-PAN are impressive: 100 dB dynamic range, –130 dBm noise floor and 90 to 100 dB image rejection when used with VE3NEA's *Rocky* or *CW Skimmer* software (discussed below). There is no doubt that similar adapters will be developed for other transceivers.

Larry also developed *LP Bridge* software to interface

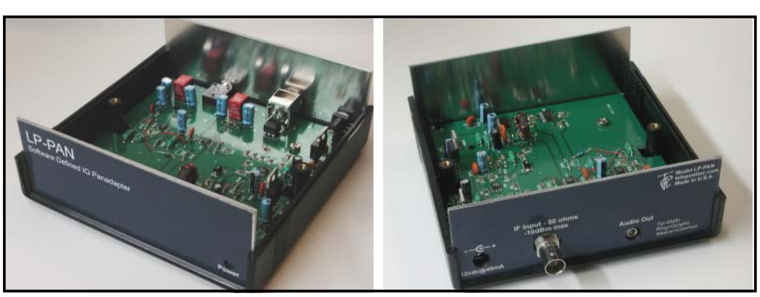


Fig 3-12 — LP-PAN is a specially designed interface for the Elecraft K3 that provides superb spectrum and waterfall displays for the K3. In addition this unit, together with SDR software such as *Rocky* or *PowerSDR* adds another radio chain to the K3.

the K3 and *PowerSDR* (for interactive control) with your logging program. This means you can control the K3 from your logging program (eg N1MM contest logging software), and also from the *PowerSDR* window. It's interactive in that if you tune the VFO of the K3 (or change bands or modes), these same changes will be applied to your logging program and to *PowerSDR*. It even goes one step further. You can simultaneously run *PowerSDR* (which gives you a very nice frequency spectrum display) and another piece of SDR software such as VE3NEA's *CW Skimmer*. *LP Bridge* actually creates an image of the K3 data in the program, an image which is then made available to the different user programs that want to receive data from the K3. Commands to be sent to the K3 are passed on in proper sequence by *LP Bridge* to avoid any timing conflicts so the K3 can talk with numerous programs, without any problem. *LP Bridge* is the Master of Ceremonies.

Scott, WU2X, (**www.wu2x.com/sdr.html**), developed a version of the *PowerSDR* software aimed at working with the IF output of complete transceivers, such as the Elecraft K3 or the Ten-Tec Orion, through an interface such as *LP Bridge* or *Ham Radio Deluxe*. The software can control all major functions of the transceiver and, using a quality sound card, acts as a complete receiver that taps its input RF from the first IF stage of the transceiver it is connected to.

1.7.4 CW Skimmer

A piece of SDR software that made waves in early 2008 is *CW Skimmer*, also by Alex, VE3NEA (**www.dxatlas.com/CwSkimmer**). *CW Skimmer* is a panoramic, multi-channel CW decoder and analyzer. The use of *CW Skimmer* is explained more in detail in Chapter 2, Section 14.3.

CW Skimmer will list the call signs it has heard on the CW band, sorted by frequency or by call sign, and display them exactly like the familiar logging software bandmap driven by the DX Cluster (see **Fig 3-13**). During a DX pileup, *CW Skimmer* will show you the DX station and the pileup stations in the waterfall display, with the call signs of the calling stations next to the waterfall display.

I have been using the WU2X version of the *PowerSDR* software with my K3, using LP-PAN, *LP Bridge* and the Creative 0202 sound card. The sound card can simultaneously handle both *PowerSDR* and *CW Skimmer*, which means that my K3 (which has the sub receiver built in) now actually has four receivers. I have been testing this setup thoroughly on the low bands and have found that under marginal conditions (weak signals) the K3 receiver is always significantly better than the SDR combinations. *PowerSDR* selectivity is very gentle, which means that the filter slopes are not very steep. I also noticed that, when there are very strong signals on the band, *PowerSDR* seems to have problems coping with them, and intermodulation appears in the band.

CW Skimmer has a very good CW filter, with much steeper and hence better performing skirts. To make a long story short, it was not my intention to have another receiver. Interfacing the K3 with *PowerSDR* was only to have a panadapter (frequency domain) display.

Interfacing it with *CW Skimmer* was done to be able to make use of *CW Skimmer*'s *raison d'être*, the simultaneous decoding of all the CW signals and the resulting call sign list, in addition to the waterfall display that let you look at all the signals of a pileup, and at the quality of the signals. With these objectives in mind, both pieces of software are a great complement for a K3, and make the transceiver even better.

In the summer of 2010, Elecraft introduced the P3, a matching standalone DSP spectral/waterfall display with point and click QSY facilities, and even an extra buffered IF output for additional IF processing (see **Fig 3-14**). This makes the K3-P3 combo the best visual display I have used, even compared to

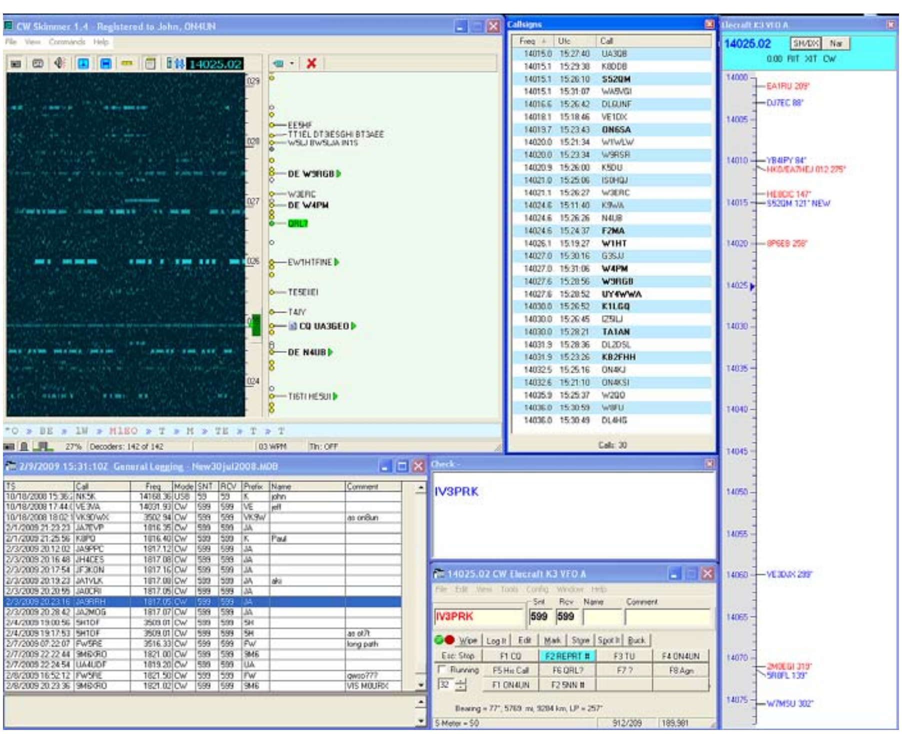


Fig 3-13 — This PC screen shows *CW Skimmer* (top left) and the *CW Skimmer* call sign list (top center). Also on-screen is *N1MM Logger*: band map along the right-hand edge; log data, bottom left; QSO entry window, bottom center; and above that the call sign check window (master.dta file checking). You can click on a call in either the N1MM band map or the CW Skimmer call sign list and send the K3 to the relevant frequency. In case of a pileup, the waterfall display lets you identify where the station is that just made a QSO with the DX station (the station where the 599 report pops up).

Fig 3-14 — The ever-so-popular Elecraft K3 with its own optional spectral and waterfall display (the P3). You can control a lot of functions of the K3 from the P3.

the luxury transceivers that cost five times as much and weigh five times as much. The advantage of such a system is that no extra high-quality sound card is required, and the operational interfacing with the K3 is optimized. It also has better dynamic range than any other system I have operated.

1.7.5 The Future of SDR

When today I write a few pages in this book about SDR, I know that most of what I have written may already be outdated by the time the book is published. But I obviously could not "not say" anything about this wonderful new technique, that will undoubtedly evolve with leaps and bounds in the coming years, and make possible things we hardly could dream about.

1.8. Outboard Front-End Filters

The transceivers that provide general coverage (100 kHz to 30 MHz) reception generally use half-octave front-end filters, which do not provide "narrowband" front-end selectivity. Older amateur-band-only receivers used either tracked-tuned filters or narrow bandpass filters, which provide a much higher degree of front-end protection, especially in highly RF-polluted areas. The newer generation of top notch transceivers now also use better and more selective front end filters.

Some excellent articles describe selective front-end receiving filters (Ref 219, 221, 251 and 266). Martin (Ref 219) and Hayward (Ref 221) describe tunable preselector filters that are very suitable for low-band applications in highly polluted areas (Ref 294).

Whether or not such a front-end filter will improve reception depends on the presence of very strong signals within the passband of the half-octave filters. Several outboard narrow-tuned-filter designs have been published over the years by K4VX (Ref 295), W3LPL (Ref 2953), K1KP (Ref 2954), and N6AW (Ref 2952). Another popular and rather simple filter was designed by the members of the Bavarian Contest Club (Ref 2951).

International Radio (Inrad) also offers front-end crystal filters, which are the ultimate solution for multi-multi contest stations for protection against interference from a multiplier station operating at the same location on the same band. Information can be obtained at **www.inrad.net**.

Many Top Banders experience problems with overload from local broadcast (BC) stations on 160 meters. Each situation requires a different approach to solve the problem. If the problem occurs with a special receiving antenna, such as a Beverage, then a tuned preselector (as described above) may help. **Fig 3-15** shows the schematic diagram, the layout and the bandpass curve of a highly effective and popular passive BCI filter designed by W3NQN (Ref 298, 299).

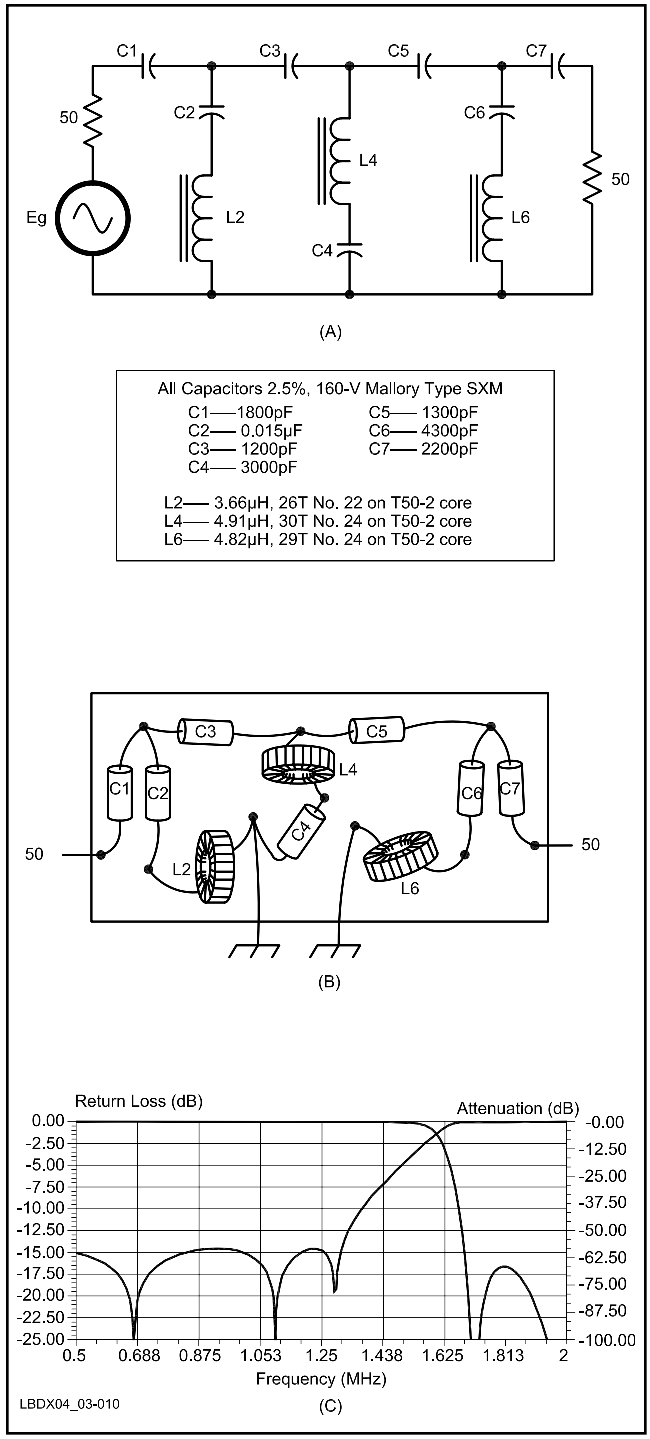


Fig 3-15 — High-pass filter designed by W3NQN. Layout at A, schematic diagram at B and response curve at C.

If you use no separate receiving antenna, and the problem exists when listening on your transmit antenna, you will need to install a similar filter that you must bypass while transmitting.

1.9. Band Splitter for Beverages

All modern transceivers have a separate input for a Beverage or other type of receiving antenna. But what if you want to split one receiver antenna among several different receivers? This is a very common situation in contests, for example in a SO2R (single-operator, two-radio) or multioperator setup. If you simply connect the Beverage coax in parallel to the two receivers operating on different frequencies, you don't really know what will happen. The input impedance of the receiver at the other frequency may be very low and may result in heavy swamping. Also, there is a danger of one of the radios getting too much RF when the other transmits.

Using a 3-dB splitter is technically okay, but you do lose 3 dB of signal. A neat solution I have used for some time now was developed by DL7AV and is shown in **Fig 3-16**. Three band filters are designed in such a way that the load impedance on the other frequencies is very high, effectively uncoupling the three band-outputs.

1.10. Adding Input Protection to Your RX Input Terminals

If Beverages or other special receiving antennas are installed very close to the transmit antenna, dangerously high voltages could destroy the input circuitry of the receiver.

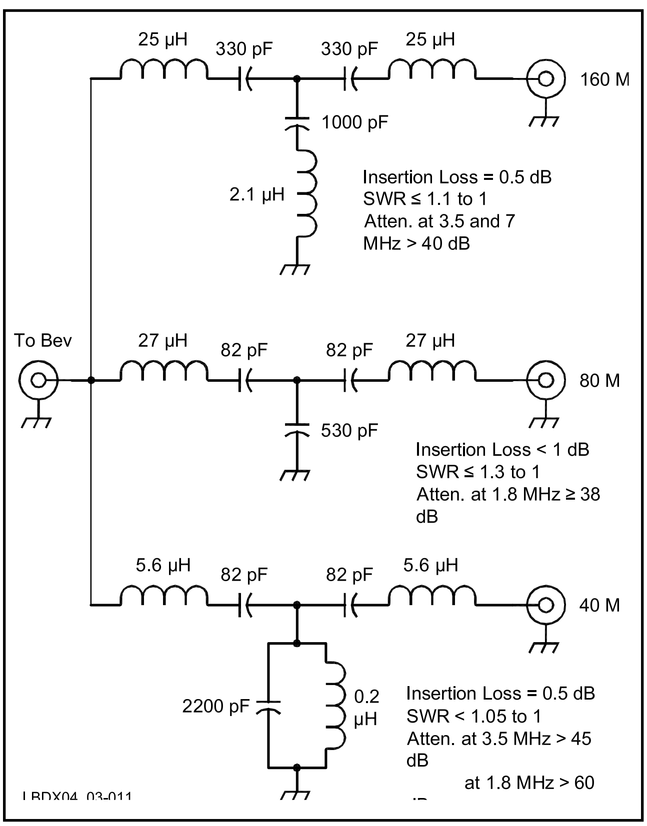


Fig 3-16 — This three-band Beverage splitter makes it possible to feed the signals from one Beverage to three receivers, operating on 160, 80 and 40 meters with minimal splitter loss (design by DL7AV).

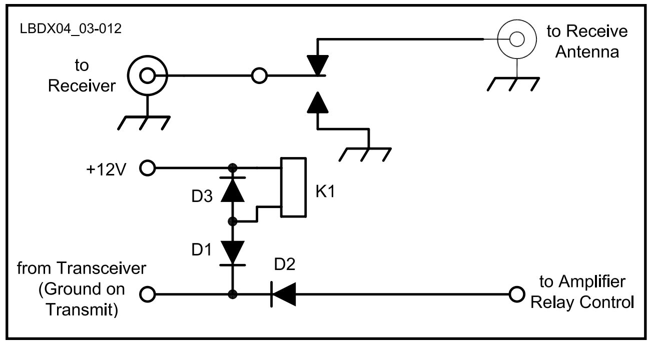


Fig 3-17 — Receiver input protective circuit. Receiver input is grounded during transmit. D2, D2 and D3 are silicon diodes.

That can happen if the transceiver circuitry does not provide for sufficient isolation of the receiver input circuitry during transmitting. Since equipment manufacturers do not always incorporate a suitable protective circuit, it may be wise to build one of your own.

Fig 3-17 shows a suitable protective circuit. A small relay shorts the input of the receiver during transmit. The voltage for the relay can be obtained from any 12-V source (usually from the transceiver itself), while the relay is activated by the amplifier control line. Two diodes make it possible to switch the amplifier and the protection circuit from the same line. It is clear that this circuit only protects your equipment from RF coming from the same transceiver. Where more than one transmitter is used (as in a multi-transmitter contest operation), a different approach — such as using bandpass filters — must be taken. It is also important to have a dc path to ground on the antenna jack. This can be a 2.5-mH choke or a 1-MΩ resistor.

1.11. Noise-Canceling Devices

When you are plagued with a local single-source manmade noise, you can often dramatically improve, or even eliminate it, using a so-called *noise-canceling device*. In a noise-canceling device, signals from two antennas (one is the regular receiving antenna, the other one is called the *noise source antenna*) are combined in such a way that the phase of the noise received on the noise antenna is of equal amplitude as on the normal receiving antenna, but exactly 180º out-of-phase. Details for such noise canceling devices can be found in Chapter 7 on Receiving Antennas.

1.12. Receiver Evaluations

Side-by-side (A/B) testing of radios is very tricky unless done at exactly the same time. On the low bands the type of noise in which we are listening for weak signals changes continuously. The tester's brain (doing the final decoding) may work differently at different times (such as when you are tired) and many other circumstances can make results of so-called A/B testing vary quite a bit.

Of course, you would like to A/B-test all radios before buying one! However, results of such very subjective testing, done under totally uncontrolled circumstances, are just very subjective. It is not possible for most of us to perform exhaustive, laboratory-quality receiver tests ourselves. To minimize confusion, I have refrained from quoting test measurement data.

If you are interested in knowing more about how receiver performance measurements are made, check W8JI's Web site (**www.w8ji.com/receiver_tests.htm**). Tom also lists a whole range of measurements he did on various transceivers (**www.w8ji.com/receivers.htm**).

Another good source of receiver test data comes from Sherwood Engineering (**www.sherweng.com/table.html**).

For *QST* Product Review testing, ARRL publishes two-tone IMD and blocking test reports at 2, 5 and 20 kHz spacing (**www.arrl.org/product-review**). As a low band DXer you should only look at the figures for 2 kHz spacing! The very detailed test procedure used by the ARRL to evaluate radios is also available to members on that Web page.

On its Web site (**www.elecraft.com**), Elecraft publishes a table comprising test data performed by ARRL and Sherwood Engineering on a range of popular radios.

Tadeusz, SP7HT, wrote an excellent article in *QEX* (Ref 446) reviewing the results on blocking and IMD dynamic range testing done by G3SJX and W8JI on several radios.

1.13. Diversity Reception

Diversity reception — as we know it on the low bands — is a method to improve the readability of a signal by using two different conditions of receiving on the same frequency. These different conditions are usually determined by using two different antennas, antennas with clearly different characteristics. In general we apply either *space diversity* (antennas not in the same space, usually separated by several wavelengths); *polarization diversity* (one horizontally polarized antenna and one vertically polarized antenna); or *direction diversity*; or any combination of those three.

What you need for diversity reception is two really "diverse" antennas. In practice that might be antennas shooting in (slightly) different directions (say, two Beverages) or antennas using different polarizations (a Beverage and a "low" dipole), or identical antennas that are widely spaced.

Sophisticated diversity systems as used in commercial links have hardware and software deciding which of the "diverse" systems gives the best S/N ratio at any given time, and reception is done only with that system (at that time). The diversity we use is normally a so-called *stereo diversity* system: two receivers are tuned to exactly the same frequency, operating in the same mode of reception, connected to separate antennas, with their audio routed to the left and the right channel of our headphones. The audio signals in those two channels are not *combined*, *weighted* or *voted* as in some sophisticated commercial systems — the weighting, voting and combining is done in our brains.

Diversity is like using a single receiver with different antennas that you switch between at a very fast rate to see which one is better. The main difference is that you can really hear gradual changes in propagation leading to one antenna being better now, and the other one better in just a few seconds or minutes.

For stereo diversity reception one needs two *identical receivers* tuned to exactly the same frequency. *Exactly* means no frequency difference and no phase difference. If you tune the two-receiver system to a carrier, all you should hear in both channels is a perfectly stable carrier, and no signs of warbling (flutter, rapid fading) caused by phase difference between the frequency of reception of the two receivers.

The first commercial transceiver that makes real diversity reception possible is the K3 from Elecraft. When equipped with the KRX3 subreceiver, the K3 has two *fully identical* receivers, driven from *the same frequency source*. The Ten-Tec Orion came close, but the two receivers were not phase locked, which limits the possibilities of such a setup.

An effective diversity system has been found to give substantial improvements in readability of noise-floor signals. As Tom, W8JI says "The difference can be worth as much as a signal being nearly readability 5 (perfect) in stereo to readability 2 without. When a signal is marginal, it is all the difference in the world."

Bill, W4ZV, has been an enthusiastic user of diversity reception with the K3: "Since I installed the KRX3 subreceiver in my K3, I would not want to be without it on 160 meters. I distinctly recall several QSOs in the Stew Perry contest where diversity saved repeats due to QSB. I remember one QSO where the DX signal shifted from left to right in my headset as he was sending the exchange. If I had been listening on either antenna alone, I would not have copied it without asking for a repeat due to the rapid QSB. And I have no idea how many more QSOs I made due to hearing people calling from different directions than my primary Beverage. The KH6LC crew reported good results by transmitting on an omnidirectional vertical and listening toward JA and USA simultaneously on Beverages using the K3 in diversity. There is no other radio on the market today that can do this as well as the K3. None at all."

1.14. In Practice

Now that you understand what makes a receiver good or bad for low-band DXing (and contesting) and after you study all the available equipment reviews, remember that what really counts is how the radio operates at your location, in your environment and with your antennas. You need to know how it satisfies *your expectations* and how it compares to the receiver you have been using so far.

It would always be ideal of course if you could test a new radio before buying it. You would not want to test it on a quiet mid-week evening but in the middle of a big contest. One of the problems is of course that most of us must go by specifications and published test data, because the vendors usually don't want you to take a radio for a test drive.

As far as I am concerned, and assuming I want to go shopping for a new radio, and if that radio would *not* be an SDR radio but a superheterodyne receiver (probably using DSP in its second IF) here is what I would look at (from a receiver point of view), and the figures I like to see:

1) The receiver should use switchable roofing filters, with selectivities down to a few hundred Hz (an SDR receiver does not have roofing filters).

2) Intermodulation dynamic range (IMD DR) must be >90 dB, preferably >100 dB (at 2 kHz spacing).

3) IMD DR for 2 kHz spacing, using test signals that both fall inside the roofing filter passband, must be >85 dB (essentially testing the second mixer performance).

4) The blocking dynamic range (BDR) must be >125 dB at 4 kHz spacing.

5) Local oscillator noise: better than –135 dBc/Hz at 4 kHz spacing.

6) The receiver should have continuously variable bandwidth (preferably in 10 Hz steps) down to approximately

50 Hz. The filter shape factor should increase with bandwidth to avoid ringing.

7) The receiver should have a wide band IF output to drive an external panadapter display (spectrum and waterfall, see Section 1.7.).

8) The CW beat tone must be adjustable down to 200 Hz (yes, some listen this low), in steps of 10 Hz maximum (see Section 1.5.3).

1.15. Receiver Areas for Further Improvement

In the Third Edition of this book, which I wrote more than 10 years ago, I begged for a series of modifications and gave the manufacturers a detailed wish list. I must admit that since then very substantial performance improvements have been made in the first place by two American companies, Ten-Tec and Elecraft. Thank you for listening to what we wanted and doing something about it. The Japanese manufacturers that traditionally held the largest chunk of the market have been slower to improve receiver performance that will help low band DXers. It is my impression that they still think that adding more bells and whistles (more weight and a bigger price tag as well) is more important. They are very obviously interested in a different market.

The success that both Ten-Tec and Elecraft have had is well deserved. These companies can and should be proud of what they have done; they are now the technology leaders when it comes to radios that meet the tough requirements of us, the low band addicts. They have clearly won the battle against strong signals artifacts on the low bands.

In this "tribute" I should not forget FlexRadio, which is, to my knowledge the only manufacturer (again, based in the USA) that has pursued the development of a full SDR transceiver (should we call it a SDT?). SDR technical performance has improved by leaps and bounds over the past five or seven years, and the latest FlexRadio transceivers have joined the ranks of the top notch equipment manufacturers for the serious radio amateur.

Do we have more desires for our receivers?

1) I would like to see a built-in adapter for large screen spectrum and waterfall display, using software like we now have from VE3NEA (see Section 1.7). Plug in a mouse, a keyboard and a LCD screen, no need for a "bulky" PC. Everything should be onboard on a plug-in optional module of course.

2) Develop systems that can effectively reduce or eliminate manmade or static noise (something radically different from noise blankers).

3) Develop true diversity receiving systems with automatic selection (including computer controlled auto-tune noise canceling systems).

4) IMD DR (2 kHz spacing), ≥100 dB and BDR (2 kHz spacing), ≥140 dB.

I must admit there is not so much left over from the list I made up 10 or 12 years ago. Not that the whole industry listened and responded, far from that, but I want the readers to know who were the driving forces behind the improvements we've witnessed: two or three American companies. Keep that in mind, dear readers, and act accordingly.

But wait a minute, many of the problems we encounter on reception are caused by *our* transmitters. I guess there still is room for a little more improvement in that department…

2. THE TRANSMITTER

2.1. Technical Issues Concerning the Transmitter

When the receiver does not meet the state-of-the-art technical standards, the user will suffer from it. Too bad for him. The solution is to switch to a receiver that meets these standards.

If your transmitter does not comply with the state-of-the-art technical standards, you may not be the only one to suffer from it. You may cause interference to other hams on the band. Using a transmitter that meets the state-of-the-art standards is also a question of ethics: don't let your signal spoil the fun for everyone on the band. Do not transmit an SSB signal that is overmodulated. Do not use a transmitter with linearity that does not meet present day standards. In both cases you will transmit splatter and take too much space in our valuable spectrum. Do not transmit a CW signal that clicks all over the place. Transmitting a bad quality signal on the bands is *bad*.

I will first cover transmitter characteristics that can cause signals to be of bad quality. These are issues you should be concerned about, and you should take corrective actions if it appears that your signal suffers from one of these defects:

- linearity problems on SSB (3rd to 9th order distortion products too high)
- badly adjusted microphone gain or speech processor (overmodulation causing splatter, excessive background noise, etc)
- key clicks on CW

2.2. How Much Power?

It should be the objective of every sensible ham to build a well-balanced station. Success in DXing can only be achieved if the performance of the transmitter setup is well-balanced with the performance of the receiving setup. It is true that you can only work what you can hear, but it is also true that you can only work the stations that can hear you. It is indeed frustrating when you can hear the DX very well but cannot make a QSO. There are some who cannot hear the DX, but they go so far as to make fictitious QSOs by "reading the *Callbook*." Fortunately those bad guys are rare.

A well-balanced station is the result of the combination of a good receiver, the necessary and reasonable amount of power and, most of all, the right transmitting and receiving antennas. To be successful with such a well-balanced station, you need good operating techniques as well. It is, of course, handy to be able to run a lot of power for those occasions when it is necessary. In many countries in the world, amateur licenses stipulate that the minimum amount of power necessary to maintain a good contact should be used. Most countries have a limitation on the maximum power that amateurs can use.

There are modes of communication where we have real-time feedback of the quality of the communication link (PACTOR and other similar error-correcting digital transmission systems). In CW and SSB operation, we can only go by feeling and by reports received, and therefore we are most of the time tempted to run *power*.

There are a number of dedicated operators who have worked over 250 countries on 80 meters or 100 countries on 160 meters without running an amplifier. But a large majority of active low-band DXers run some form of power amplifier, and most of them run between 800 and 1500 W output.

2.3. Power Amplifiers

Today the newest technologies are utilized in receivers, transmitters and transceivers to a degree that makes competitive home construction of those pieces of equipment out of reach for all but a few. Most high-power amplifiers still use vacuum tubes, however, and circuit integration as we know it for low-power devices has not yet come to the world of high-power amplifiers. At any major flea market you can buy all the parts for a linear amplifier.

The amplifier builder will usually build more reserve into his design. He will have the option of spending a few more dollars on metal work and maybe on a larger power supply transformer to build a product that runs cool all the time and never lets him down. Maybe he will use two tubes instead of one, and run those very conservatively so that the eventual cost-effectiveness of his own design will be better than for a commercial black box.

Power amplifier designers and builders have their own reflector on the Internet, where very interesting information is exchanged among builders. For more information about the reflector and instructions for joining, see **lists.contesting.com/mailman/listinfo/Amps**. If do you consider building your own amplifier, make sure you carefully study *The ARRL Handbook*, as well as W8JI's Web site, which has lots of good information on this subject (**www.w8ji.com/Amplifiers.htm**).

One of the problems we seem to face in the coming years is the availability of power transmitting vacuum tubes, at least at a reasonable price. As commercial designs are moving to solid-state amplifiers, sources of new and surplus vacuum tubes for amateur power levels are dwindling. It appears that even the Russian made high-power tube sources have all but dried up.

Solid-state HF power amplifiers are commercially available for power typically up to 1 kW, and as this is written several manufacturers have announced 1500 W versions. It seems as though it would be possible to develop a modular power amplifier system where a number of high power (500 W to 1 kW) modules could be used together. This technique is widely and successfully used for FM broadcast transmitters. I would not be surprised to see Amateur Radio solid-state amplifiers made according to this principle.

Fig 3-18 — The Expert 2K-FA solid state amplifier is a state-of-the-art solid state amplifier using a built-in power supply and an automatic antenna tuner.

One solid-state amplifier I am familiar with is the Expert 1K-FA (**Fig 3-18**), manufactured by SPE in Italy (**www.linear-amplifier.com**). This 800 W amplifier, about the same size as a K3, weighs just 20 kg (less than an FTdx9000 or IC-7800!) and that includes the power supply as well as an automatic tuner. The amp covers 160 through 6 meters, and can be remotely controlled by a computer. I have been using this amplifier quite extensively for a number of years with no problems at all. The nice thing about a solid state PA is that it has zero warm up time.

The same company has announced the Expert 2K-FA, which is the 2 kW version of the 1K-FA. The amplifier, including power supply and antenna tuner also weighs no more than 20 kg! You can switch the amplifier between 500 W, 1 kW and 2 kW output. Voltage and bias are adjusted so that best linearity is maintained at each power setting. The power supply is a switched mode supply, which includes automatic power factor correction. Even at 2 kW carrier output (into a dummy load), there is no trace of ripple on the envelope. The amplifier covers 160 through 6 meters and appears to be the first commercial solid state amplifier delivering that kind of power across the frequency range with that kind of automation level. It is an ideal candidate to install remotely and control and monitor via a USB link to your computer. No heat, no noise, just performance.

2.3.1. Using ALC with an External Amplifier

In a properly designed and operated station there should be no need for external automatic level control (ALC) between the amplifier and the transceiver. I never use ALC. I find ALC evil. Controlling the output from a 200 W exciter by means of ALC from an amplifier that only requires 50 W of drive is a bad thing because of the attack time constant inevitably associated with an ALC circuit. Most ALC circuits start regulating *after* a problem has occurred. The problem is usually a leading edge signal power overshoot, which often shows up as an adjacent channel "spit" or "pop" on leading or rising edges of voice or CW. All of this results in extra sidebands — that is, clicks on CW or SSB splatter. In a good transmitter you can adjust the output power very precisely, which means you do not need an ALC.

Amplifiers with sensitive overdrive monitoring systems (such as the ACOM 2000A) will sense the overshoot and shut off. Don't blame the amp, blame your transmitter or transmitter setup.

If you do use ALC, turn the drive control of your transceiver down to the point where the ALC just starts working. Don't let the ALC adjust the power output level of your transceiver, do it manually!

I never transmit without watching the envelope of the transmitted signal on a monitor scope. You can best adjust the driving power to your amplifier by watching your monitor scope.

2.4. Operating SSB

If you choose to play the DX game on phone (SSB), there are a few points to which you should pay great attention.

2.4.1. Splatter

SSB transmitters can generate intermodulation distortion (IMD) products — splatter — that interfere with other stations

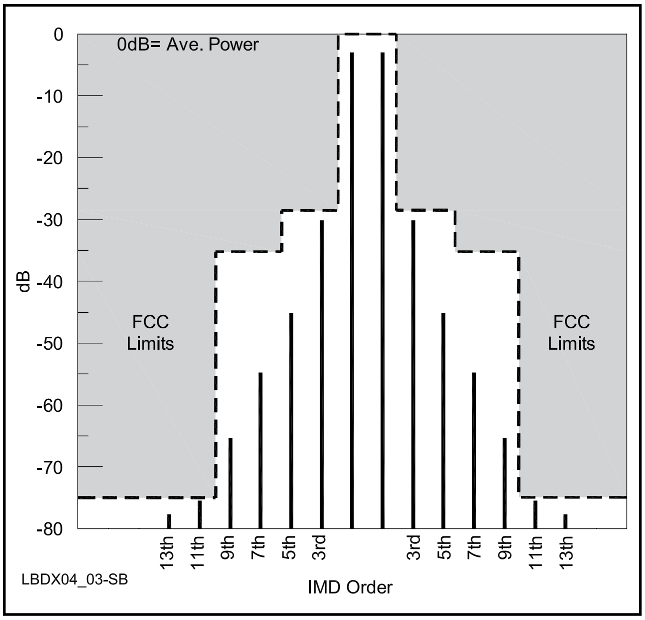


Fig 3-19 — FCC commercial specifications for transmitted IMD. For 11th and higher-order IMD products, the required suppression for a 1500-W transmitter is 75 dB below average power (the FCC definition), or 78 dB below PEP (the convention used by amateurs).

up and down the band. We have all been infuriated over and over again by stations moving perhaps 5 kHz below our frequency (on USB) and creating strong splatter, especially during contests. If the offending station is S9 + 20 dB on-channel, and his buckshot is S7 off-channel, his splatter is only suppressed 32 dB. That is not good enough, I think.

Even if manufacturers would all install pure Class A (low distortion) finals in their transceivers, very poor quality signals (with poorly suppressed IMD) could still be a problem if the transmitter is not properly adjusted. In general, we should expect that the ham who buys a rig with a Class A final is concerned about signal quality and will do all he can to properly drive his transmitter.

Look at **Fig 3-19** to understand what is meant by 3rd order, 5th order etc. If you test a transmitter with two signals 2 kHz apart, the 3rd order products are 6 kHz apart, the 5th order 10 kHz , the 7th order 14 kHz and so on. An SSB signal contains audio components that are easily 2 kHz apart (300 Hz to 2300 Hz spectrum).

Now how much are these products suppressed in an average commercial transceiver using a class AB1 transistor power amplifier? Third order: 27 to 37 dB, 5th order: 40 to 49 dB, 7th order: 46 to 52 dB, 9th order: 50 to 60 dB (these figures come from reviews of a range of modern transceivers, as published in *QST*). What's an "average transceiver"? I went through reviews of all major brands and types and noted the IMD ranges.

Are these figures good? If your neighbor, who is 45 dB over S9, has 7th order products down 46 dB, it means he will cause splatter that is S9 over a range of at least 14 kHz. Is that good?

One could wonder how transmitters with 3rd order IMD products less than 30 dB down meet today's technical specifications. Fig 3-19 shows a graph representing the current FCC specifications. These specs are far from stringent, and as long as the rule makers will not tell the industry to do better, we may not see much improvement. See also **Fig 3-20** and **Fig 3-21**.

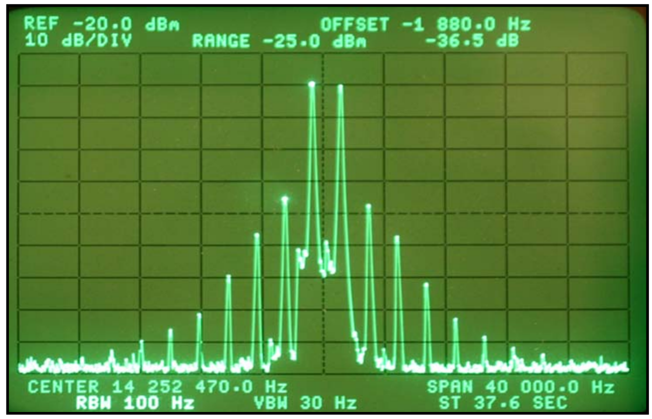


Fig 3-20 — A 1970 vintage Collins 32S3 transmitter using a pair of 6146B vacuum tubes in class AB1. The 100 W final amplifier yields 3rd order IMD of –36 dB and 5th order of –47 dB, which is significantly better than modern semiconductor type AB1 finals (source: NCØB, Sherwood Engineering, measured on 20 meters)

We have come to the point in receiver performance where it now will take better quality transmitted signals to be able to take full advantage of the receiver improvements. This is a call to the industry to make transceivers with much cleaners signals on SSB. For more details on this subject see also W8JI's Web site: **www.w8ji.com/transmitter_splatter.htm**.

What about Class A finals? In Class A the IMD figures are substantially better, especially the higher order distortion products: 3rd order: better than –40 dB; 5th order: better than –65 dB; 7th order: better than –70 dB. See **Fig 3-22**. Note that a Class A driver, followed by a Class AB1 final amplifier, will lose some of that performance as shown in **Fig 3-23**.

The problem with Class A finals is their heat dissipation. Maybe we have to design transceiver with 20 or 30 W peak

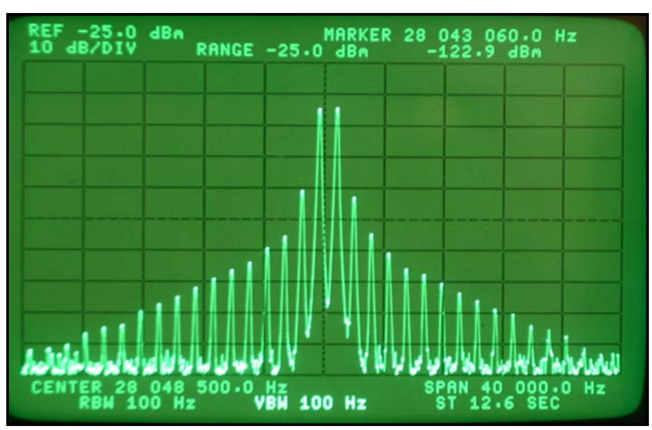


Fig 3-21 — The Elecraft K3 100 W final amplifier measures –29 dB 3rd order IMD and –43 dB 5th order IMD. Compare this with the Collins 32S3 from 40 years ago! (source: NCØB, Sherwood Engineering, measured on 20 meters)

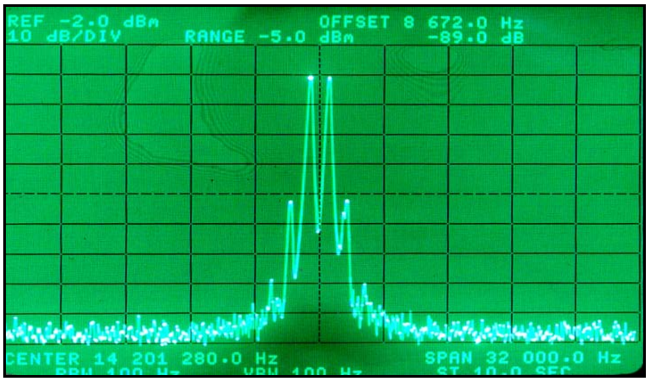


Fig 3-22 — Yaesu FT-1000MP MkV with PA at 75 W out, operating in Class A: –42 dB 3rd order IMD and –70 dB 5th order IMD. (source: NCØB, Sherwood Engineering, measured on 20 meters)

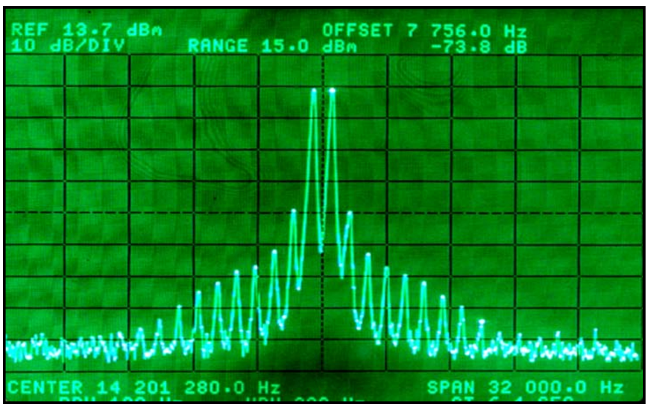


Fig 3-23 — The FT-1000MP MkV running in Class A, followed by an 8877 amplifier: –40 dB 3rd order IMD and –52 dB 5th order IMD. (source: NCØB, Sherwood Engineering, measured on 20 meters)

power in Class A, and design companion amplifiers that require that much power for 1 kW or even 1.5 kW out.

As far as improving the quality of SSB signals, a number of things can be done, I think:

1) Educate the users to properly adjust their transmitters. This is a process that takes time, and we should not forget that it's always "the others" that need to be educated… Tom, W8JI writes on his Web site: "The three greatest sins creating unnecessary bandwidth are: A) Turning up a radio's internal power or 'drive limit' pot. B) Enhancing bass and treble. C) Under-loading an amplifier."

2) Discourage people from internally adjusting the maximum power output from their transceivers. This is the best way to guarantee a poor quality signal.

3) If your station does use ALC (which I discourage), make sure it barely moves the transceiver's ALC indicator on peaks. Don't use the ALC to set the right driving power for your amp; do it manually.

4) Make sure you have properly tuned your amplifier. Too light loading causes flat topping and splatter, and is shown by excessive grid current.

5) Ask the manufacturers to design equipment with better high order IMD rejection (which means Class A finals). Forty years ago a Collins KWM2 yielded –36 dB 3rd order IMD products using a tube final. Today we see top notch transceivers exhibiting a very mediocre 3rd order IMD figure of merely 25 dB. Figures below 30 dB seem to be the rule, rather than the exception! If one of the manufacturers would take the initiative to substantially improve transmitter IMD performance, perhaps the others would join in.

6) Ask the ruling bodies to make more stringent specifications regarding high-order IMD products for amateur equipment. If point 5 does not work, maybe this is what will have to be done.

7) Ask the manufacturers to develop systems that control the quality of the SSB transmissions, systems that make it impossible to misadjust or overdrive the equipment.

8) Make it mandatory for each station operating with power of more than 100 W (peak) to be equipped with a monitor scope, and make it a must to monitor the transmitted envelope at all times.

9) Include a quality rule in contests which says that stations with really dirty signals could have their scores reduced. We could have an independent jury note the quality of the signals, and then give a kind of multiplier to the score. Excellent signal: Multiplier = 1; Bad signal: Multiplier = 0.7; Very bad signal: Multiplier = 0.4. Using SDRs we can easily record the entire band spectrum (on each band) for the entire duration of the contest, so that verification is possible at all times.

2.4.2. Microphones (and Headphones)

Never choose a microphone just because it looks pretty. Never choose a microphone just because it is expensive either. And never choose one just because it's from a well known brand. Always chose a microphone because it sounds good with *your* voice and *your* transceiver.

Until recently transmitters did not have elaborate audio tailoring facilities built in. Many newer transceivers using DSP have audio equalization built-in. The Elecraft K3, for example, has an 8-band equalizer. In each of the sub bands (centered around 50, 100, 200, 400, 800, 1200, 2400 and 3200 Hz) one can boost or cut up to 16 dB. That is similar to the popular W2IHY equalizer, only this capability is built-in.

Why is equalization important? It means that, whatever your real voice sounds like, you can tailor it to sound like BBC quality broadcast audio. Or you can make it sound sharp, penetrating, high pitched (and awful as far as I am concerned) if that is what you prefer. Don't forget the power is in the low frequencies, the intelligibility in the high frequencies. You can do all of that with one and the same microphone.

The same performance can be achieved with the W2IHY equalizer (see **www.w2ihy.com**). This unit can turn a studio-quality microphone (such as the Heil Goldline mike) into an efficient DXing and contesting microphone.

Although I am much more a CW operator than a phone man, I like to transmit a nice, full and round audio with lots of body (like me…). In the past five years I have tried all kinds of microphones — Heil HC4 and HC5, Goldline studio condenser microphones, you name it. With the W2IHY equalizer and various types of rack mounted Behringer equipment, I must tell you at one time my shack looked like a professional audio studio… and that for the few contacts I ever make on phone.

Some time ago I was looking for a headset/microphone combination that would suit my taste. What does that mean? It should be lightweight, semi enclosed ear cushions (not ear "crushers"), full headphone quality, reasonable microphone quality (you can shape that up anyhow), and, yes, not too expensive.

There are literally hundreds of types and brands of mike-headset combinations available for the computer industry. After some testing I found one particular headset (Wintech WH-41), which I bought for less than $5 each, and which works really great for me. Note: I was unable to find a dealer in the USA. By the time you read this, this particular type of microphone will likely no longer be available, but other similar units will surely come along and be equally as good or even better. And maybe even cheaper (BTW, now I have five spare units).

The WH-41 uses an electret type microphone, so I just needed to make a little adapter box to supply the 5 V to the microphone. For more info see **www.epanorama.net/circuits/microphone_powering.html**. With this microphone going into the W2IHY 8-band equalizer I can produce sharp contest quality or broadcast quality audio that rivals the audio from the Heil Goldline microphone! And that for the price of $5 for the microphone and the pair of headsets! And I get nothing but good compliments about my audio.

If you use a K3, you can simply plug in a Heil or PC-type headset-microphone combination using an electret cell into two jacks on the rear panel of the unit. This avoids the usual clutter caused by the headphones and microphone jack on the front panel.

Much more important than the choice of the microphone is the tailoring of the audio in the transmitter. A multi-channel equalizer can do magic. Most important of all is how you *use* the microphone. Communications microphones are made to be held close to the mouth when spoken into. Always keep the microphone a maximum of a few centimeters from your lips. A very easy way to control this is to use a headset/boom-microphone combination.

If you do not speak closely into the microphone, you will have to increase the microphone gain, which will bring the acoustics in your shack into the picture, and these are not always ideal. We often have a high background noise level because of the fans in our amplifiers. This background level, and the degree to which we practice close-talking into our microphone, determines the maximum level of processing we can use. In any case, adjust your transmitter audio gain and processing level so that the background noise level is sufficiently suppressed.

2.4.3. Speech Processing

The human voice contains a mix of tones, from low to high. The power in these components tapers near the low end and near the high end of the voice spectrum. This natural distribution of the voice results in an effective low bandwidth which in turn ensures a reasonably low distortion content.

If we process the voice with the aim of equalizing the power level at *all* frequencies within the voice passband (in order to increase the "average power"), we will inevitably increase IMD products spreading on both sides of the SSB signal. This is because the average level of lowest and highest modulating frequencies has been increased. Although processing brings the average power level of lows and highs up, it can also decrease overdrive (flat topping) problems. A modest amount of clipping or limiting removes modulation peaks and thus reduces chances of flat topping and creating splatter.

Although speech processing can reduce chances of splatter, it is mainly used to *improve the intelligibility* of the signal at the receiving station. Increasing the ratio of average power to peak power without worrying about the quality (readability) of the transmitted signal will not do much good. If the processing introduces lots of distortion you may actually hurt your intelligibility.

Although audio clippers can achieve a high degree of average power ratio increase, the generation of in-band distortion products raises the in-band equivalent noise power generated by harmonic and intermodulation distortion. That in turn decreases the intelligibility (signal-to-distortion-and-noise ratio) at the receiving end. RF clipping generates the same increase in the ratio of transmitted average power to PEP, but does not generate as much in-band distortion. This basic difference eventually leads to a typical 8-dB improvement of intelligibility over AF clipping (Ref 322). Virtually all current high-end transceivers are equipped with DSP speech processors that perform even better than RF clippers.

Adjusting the speech processor level seems to be a difficult task for some operators, if you judge from what we sometimes hear on the air. Modern transceivers have a compression-level indicator, which is very handy when adjusting the clipping level. Much better is to use your monitor scope to make a proper adjustment.

We already stated that the acoustics in the shack will be one of the factors determining the maximum allowable amount of speech clipping. By definition, a speech-clipped signal has a low dynamic range. To not be objectionable, the dynamic range should be kept on the order of at least 25 dB. This means that during speech pauses the transmitter output should be at least 25 dB down from the peak output power during speech. Let us assume we run 1400 W PEP output. A signal 25 dB down from 1400 W is just under 5 W PEP. Under no circumstances should our peak-reading wattmeter indicate more than 5 W peak (about 3 W average), or we will have objectionable background noise (Ref 305).

One way of getting rid of the background noise (from fans, etc), is to virtually extend the dynamic range of the audio by quieting the audio when the audio level drops below a certain threshold. The W2IHY audio equalizer described earlier has such a "noise gate" feature built-in.

Another solution is to have your amplifier in another room and remotely control this equipment: no noise, no heat, very comfortable…and no background noise. In my shack the dynamic range is between 45 and 50 dB if everything is properly adjusted.

In his excellent publication *Managing Interstation Interference*, George W2VJN reports the results of extended tests he did on the efficiency of speech processors in various types and brands of equipment. In his test the peak output was set for 100 W, and the clipping level adjusted as explained in the manual (or 10 dB if nothing was suggested). For the test the driving signal was white noise, in order to ensure repeatability of the test results The average power increase with different types of equipment appears to be all over the place, ranging from 5.2 dB (with the K3), to 3.0 dB for the ever popular FT-1000MP, to as little as 1.3 dB for an IC-756ProIII, 1.0 dB for the FT-2000, or 0.8 dB for the IC-7800.

2.5. Operating CW

2.5.1. Keying Waveform and Key Clicks

In older generations of transmitters, you could adjust an RC network to change the leading and trailing edges of the keying waveform to ensure a clean CW signal without clicks. **Fig 3-24** shows an example of a nicely shaped CW waveform. Transceivers from the 70s through the 90s provided no control whatsoever over the keyed waveform shape, resulting in many commercial transceivers that transmitted terrible CW. Ten-Tec, known for producing good CW rigs, was — to my knowledge — the first to provide a facility where rise and fall times can be adjusted between 3 and 10 ms (in the Orion transceiver).

For decades the majority of equipment designers have shown very little interest or very little know-how in the field of producing good CW signals. It requires knowledgeable engineering to get the proper results. For example, for years we told the designers at Yaesu that there was a serious key click problem with the FT-1000 series transceivers, but the problem continued with newer FT-1000 series radios. In the beginning of 2003 a modification was developed by Yaesu and applied to all new transceivers leaving the factory, although this was not announced to the public. Several people, including W8JI, tried the factory modification. The results were disappointing. W8JI wrote to me: “We are in very big trouble, because for every ten new radios sold, at best one will get repaired correctly. This will eventually ruin the low bands for many years to come. Many people have horrible clicks and refuse to fix the radios, or use poor corrections. We need to put great pressure on Yaesu and other manufacturers to correct radios or we will slowly lose all pleasure on low band CW.” We had to wait another four years until the FT-2000 came about to see the problem solved.

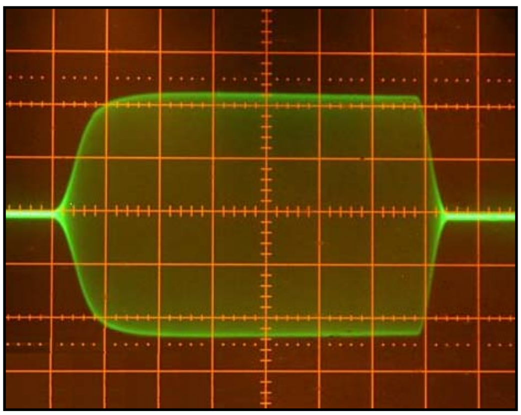


Fig 3-24 — Excellent keying waveform of Elecraft K3 (20 WPM, 100 W out, horizontal scale 10 ms/cm). (source: K3CW)

Fortunately two excellent engineers, Tom, W8JI, (**www.w8ji.com/keyclicks.htm**) and George, W2VJN, of International Radio (**www.inrad.net**) dug into the FT-1000 click problems on their own and came up with modifications that cure the problem. The bandwidths occupied in the W2VJN-modified FT-1000MP and the Ten-Tec Orion are very similar. **Fig 3-25** shows the spectral display of an unmodified FT-1000MP and one with the W2VJN modifications.

Five years later it looks like a good number of the existing FT-1000s have been modified by their users and the number of signals with horrible key clicks has dwindled. We should all be grateful to both Tom and George for helping clean up these transmitters. If you happen to have an FT-1000D, MP or MkV that has never been modified to solve the key-click problem, please apply the modifications. If you cannot do it yourself, ask a friend to do it for you. It is unethical to transmit with a

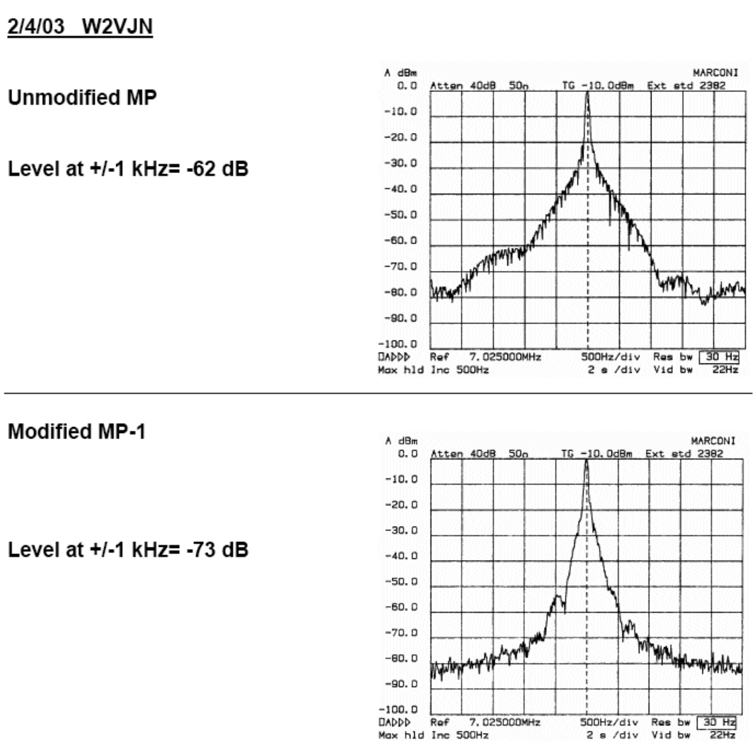


Fig 3-25 — W2VJN of International Radio measured an FT-1000MP before and after modification to reduce key clicks. At spacings from the carrier of ±1 kHz, the modification reduced clicks by about 11 dB, a very significant amount.

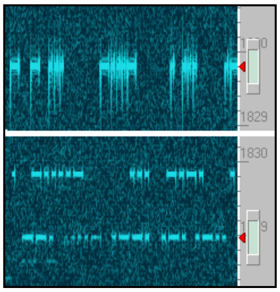


Fig 3-26 — Waterfall (time domain) display of (top) a signal with key clicks and (bottom) two clean signals (using *Rocky* software by VE3NEA).

signal that has bad key clicks, just as much as it is unethical to transmit on phone with a signal splattering all over the place. Please be considerate.

If you use a receiver with a waterfall screen (time domain display) you can immediately see the signals with bad clicks. **Fig 3-26** shows the waterfall display of a few signals on 160 meters. On top is a signal that spreads out over more than 1 kHz. The two signals at the bottom are equally strong but take only about 1/3 of the space. If you want the cleanest possible CW signal, check the ARRL Product Review test reports in *QST* and look for the keying sidebands test results (two examples are shown in **Fig 3-27**). Past reviews are available for members on the ARRL Web site.

2.5.2. Leading-edge Spikes

Another common problem with some modern transceivers is that they generate a power surge on the leading edge of the first CW character. This surge is in some cases twice the level of a constant key-down signal (up to four times with some types of ICOM rigs). This causes increased transmitted garbage, sounding like key-clicks, and can trip the protective overdrive circuits of some commercial amplifiers. If you have an amplifier with an elaborate protection system, chances are that this spike will cause the protection system to trip!

In some transceivers this problem can be overcome by turning the RF OUT knob down to the point where the output power just begins to drop. But some transceivers use an internal ALC (automatic level control) loop that is controlled by the front-panel RF OUT knob. The attack time of an ALC is designed to be fast, but it isn't instantaneous. The delay before the ALC can automatically reduce the transmitter gain allows the initial spike to appear at the output.

Do you have that problem on your transceiver? The only way to be sure is to use a high grade (large bandwidth, fast) scope to check the envelope. I have been using an ACOM 2000A amplifier which has a separate protection circuit that shuts down the amplifier in case of an important spike. I must say I have never had any problem with FT-1000s (D, MP, MkV), Ten-Tec Orion or Elecraft K3.

2.5.3. QSK, Semi Break-In and Amplifier Switching Timing

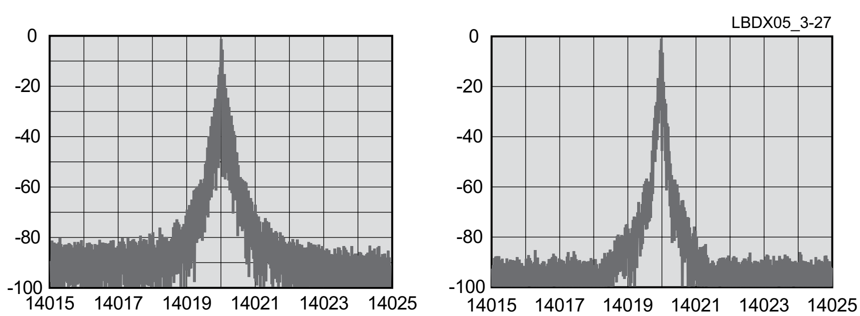
QSK (full break-in) is a nice feature, but not essential, either for the low-band DXer or for the contester. However, properly implemented QSK can be an asset to contesters and DXers. When calling, an operator can hear immediately when the DX transmits and can stop sending. This is beneficial to everyone on the frequency. QSK can help determine the DX station's pattern so that he can be called at the right time. Of course, good QSK used by two operators during a rag chew is really a pleasure.

If not properly designed and set up, however, QSK can be a disaster. It can generate severe key-clicks. It can ruin the antenna relay in the amplifier in no time, or cause component arcing and destruction of very expensive components (such as band switches) in the amplifier. Even stations operating semi-break-in on CW often exhibit poor TR timing and hot switching. In extreme cases the entire first dit is missing, so a JA call sign turns into an OA (Peru) or a W-station (USA) turns into an M-station (England).

How to make sure that you do not have this problem? There is only one way — use a scope and look at the envelope pattern of the CW. If you do not see the waveshape starting gently from zero, you more than likely have a problem. If you don't want to ruin your amplifier, or ruin the bands with clicks and clacks for your neighbors, have a close look at the timings involved in your station. You may need an external sequencer

Fig 3-28 — Tom, W8JI developed a universal time variable sequence unit, which provides properly coordinated switching for transceivers, power amplifiers and preamplifiers. See text for discussion of timing issues.

Fig 3-27 — Worst case spectral display of two 2008 grade transceivers. On the left for the Yaesu FT-2000D, on the right for the Elecraft K3 (both at 60 WPM, full power). The K3 signal bandwidth is narrower, and the far away sidebands and transmitter noise are 10 dB further down on the K3. (source: *QST* Product Review)



such as the one shown in **Fig 3-28** to help coordinate switch closures and signals among your various station components.

Poor timing and hot switching can be avoided if the manufacturers obey the following general rules. The sequence of events on all QSK modes should be:

Make side:

• Appearance of signal input (key closure or data input).
• The transmitter immediately sends "on" signal to amplifier with minimum possible delay.
• Ideally the transmitter should have an adjustable RF-on delay (1 to 30 ms), after which it allows RF output to start rising. If not adjustable, 15 to 20 ms is a must to accommodate amplifiers with slow relays.
• Wait for handshake signal (if handshake system is active), then deliver RF to the amplifier.

Break side:

• Data stops.
• RF output from transmitter stops with zero delay in the SSB mode. In CW the RF should be extended after the key is lifted by the amount of the make side delay to maintain the proper timing.
• After making sure the envelope is just at zero, the amplifier keying line unkeys.

The sequence on *semi-break-in CW* or *VOX* should be:

Make side:

• Appearance of TX signal input (key closure, data input).
• Transmitter keys the amplifier relay line without any delay.
• The transmitter should have adjustable RF-on delay (to make sure the amplifier relays are closed, see QSK mode) after which it allows RF output or checks for handshake if used. If no adjustable time, a minimum of 15-20 ms is required, and this may not be enough for some older amplifiers.

Break side:

• Data stops.
• RF envelope reaches zero.
• After independently adjustable OFF delay (0-1 second hang delay), the amp unkeys. In better transmitters there is an independent adjustment for voice operation (VOX) and for semi-break-in CW. During semi-break-in on CW, this delay is advanced to a point where the amplifier relay does not clatter at the CW keying rate. It drops out after an adjustable delay.

On *SSB* the transmitter should have standard VOX adjustments (sensitivity, anti-VOX and delay, or hang-time) then generally follow this rule:

• The VOX trips and the amplifier immediately comes up.
• After an adjustable TX delay (can be the same as CW or data, a minimum of 15-20 ms), the RF comes up.
• The VOX hang-time is set to drop out only after the RF has reached zero, even at fastest hang setting.

How can we know that the TX delay of 15 ms is enough? The best way is to watch on a scope (envelope pattern) and to listen for the transmitted signal quality in a second receiver. Listen around the transmitted signal and on its harmonics. Adjust the time delay for total cleanliness.

If the delay is longer than 20 ms it may fool very fast CW operators, and in that case it might be time to have a look at installing faster relays in the transmitter (for example, small vacuum relays).

2.6. Operating PSK

A mode that is becoming increasingly popular on the low bands is PSK. In Europe there is a lot of PSK activity in the 1838-1840 window. Because of its theoretical narrow bandwidth (less than 100 Hz), this mode is quite suitable for use on a narrow band.

It is very easy to overmodulate the transmitter, resulting in a very wide signal (up to 1 kHz and more). Therefore it is very important to adjust the equipment correctly.

A few guidelines:

• Keep audio processing and/or speech processing switched off *at all times*.
• Run as little power as necessary to have a solid QSO.
• Use an oscilloscope to monitor the waveform of your transmitted signal. **Fig 3-29** shows the waveform of a well adjusted PSK31 signal.
• When running at 100 W PEP, the power meter of the transmitter will indicate 50 W, provided the transmitter is not overmodulated. A 100 W transmitter can be run at 100 W PEP for long periods of time (the wattmeter indicating 50 W). The *duty cycle* is 50%.
• Dedicated test equipment is available for monitoring the quality of the outgoing signal. Two examples are the PSKMETER by KF6VSG (**www.ssiserver.com/info/pskmeter/**) or the IMD Meter by KK7UQ, shown in **Fig 3-30** (**kk7uq.com**). The use of such equipment or an oscilloscope is highly recommended.

2.7. Signal Monitoring Systems

You should have some means of monitoring the quality of your transmissions. All modern transceivers have some sort of built-in monitor system. The best ones are not mere audio output monitors, but instead monitor the directly detected SSB

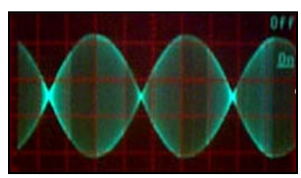


Fig 3-29 — RF envelope of a properly modulated PSK signal.

Fig 3-30 — KK7UQ developed an IMD meter to show PSK operators at all times how much transmitter IMD products are down. This is what we need for SSB transmitters!

signal, so you can evaluate the adjustment of all your relevant settings. This feature allows the operator to check the audio quality and is particularly useful for checking for RF pickup into the microphone circuitry.

A monitor scope should also be mandatory in any amateur station. With a monitor scope you can:

- Monitor your output waveform (envelope).
- Check and monitor linearity of your amplifier (trapezoidal pattern).
- Monitor the keying shape on CW.
- Observe any trace of hot-switching on QSK.
- Check the tone of the CW signal (for power-supply ripple).
- Correctly adjust the speech processor.
- Correctly adjust the drive level of the exciter to optimize the make and the break waveform on CW and to avoid leading-edge overshoot.

I have been using a monitor scope at my station ever since I got licensed almost 50 years ago, and without this simple tool I would feel distinctly uncomfortable when on the air. Many years ago Heathkit, Yaesu and Kenwood sold scopes that were developed for this particular application.

They were rather expensive and had one distinct disadvantage: You had to route the full output RF from the amplifier through the 'scope to tap off some RF, which is fed directly to the plates of the CRT. It is much easier to buy a good second hand 20 MHz professional scope, which will cost less than a new monitor scope. All you need to do is to sample a very small amount of the transmitted RF to feed to the input of the scope.

Mid-2010 Larry Phipps, N8LP (**www.telepostinc.com**) introduced a beautiful multi-function station monitoring system, the LP-500, that does all of what is listed above, and more. See **Fig 3-31**.The monitor uses a 9.5 by 5.4 cm color TFT display. It can perform the following functions:

- Power output and SWR meter, displaying simultaneous bar graphs for peak and average power with a readout of compression (in SSB).
- Signal envelope monitor (oscilloscope) with presets for CW, SSB, PSK etc, or even trapezoid (for checking linearity, in which case two RF couplers are required).
- Spectrum analyzer (dynamic range >70 dB): displays the modulation spectrum of a transmitted signal. A two-tone signal generator is built in, readout of 3rd, 5th and higher order IMD when using two-tone SSB or PSK.

In one word, this looks like the ultimate station monitoring device, which I will certainly add to my station as soon as available. It is clear that this monitor system can advantageously replace the monitor scope for the task of keeping an eye on the quality of the transmitted signal.

2.8. Transmitter Areas for Improvement

Transceivers have made more technical progress on the receive side than on the transmit side in the past 5 to 10 years. We still hear key clicks on the bands, but luckily most have been fixed by now thanks to knowledgeable people.

On SSB it still is perfectly possible to transmit a signal that splatters all over the place. It is a fact that many hams do not know how to transmit a clean signal; some may even not want to do that. Could the industry not develop transmitters that simply cannot be overdriven? Isn't it possible to design a real automatic gain control system in our transmitters that keeps the signal levels below the point where distortion sets in? How about "idiot proof" transmitters that simply can not produce splatter!

Maybe the manufacturers could start by developing a monitoring system that continuously displays how much the

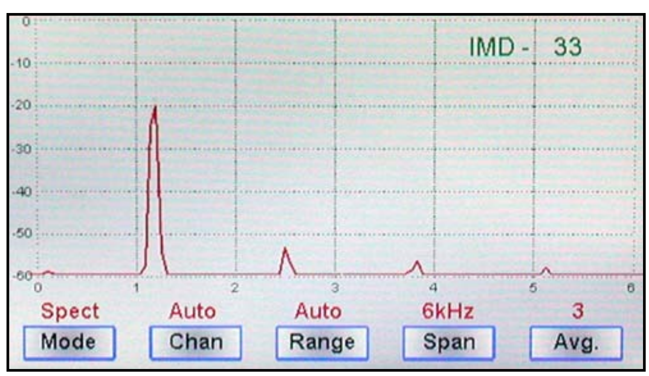


Fig 3-31 — N8LP's LP-500 monitoring system: On the left we see the display of a perfectly adjusted two-tone test or a perfectly adjusted PSK signal. The right (top) shows the leading slope of a CW signal with approximately 5 ms rise time. Below that is the spectrum analyzer screen for a two-tone SSB test, with the 3rd order IMD 33 dB down. The 5th and 7th order IMD products are also visible at –37 and – 38 dB.

intermodulation products of our transmitted signal are down. Something like the KK7UQ IMD Meter for PSK. Once we know what the IMD figure is, we could link the data to a loop system to control the gain in different stages of the transmitter, or even, in the worst case shut off the transmitter.

Here is my transmitter wish list, and it is addressed to the equipment manufacturers.

- Improve the intermodulation distortion products of the transmitter significantly.
- Make an idiot proof transmitter that cannot splatter (note that the Elecraft K3 incorporates a system that controls the signal level all through the transmitter chain to guarantee proper IMD suppression).
- Reduce noise sidebands (VCO noise to at least –140 dBc/Hz at 1 kHz separation).
- Provide an easily and precisely adjustable power output control.
- Built-in 8-channel audio equalizer.
- Provide an input for all types of microphones.
- Incorporate an automatic (background) noise reduction system that at all times ensures at least 25 dB signal to background noise ratio.
- Eliminate all leading-edge power spikes on CW.
- Provide perfectly shaped CW.
- Reduce standard SSB transmit bandwidth to 2.1 kHz (now often 2.7 kHz).
- Provide fully adjustable timing controls for QSK and semi break-in operation (to match amplifier characteristics).

3. THE BEST RADIO

A substantial number of readers will probably just want to know what's the best radio available today to the DXer who wants to score well on the low bands.

If you have read all the foregoing sections in this chapter, you should already know the answer. At this time (February 2009) I think that the Elecraft K3 is the champion. That does not mean that you will not be able to do very well with other radios. If you want to be able to hear well when your close neighbor is on the air, or hear the weakest signals during a popular low band contest, when super-big signals congest the band, the K3 is the radio to have. In my experience the Ten-Tec Orion comes close, but for me the main reason for preferring the K3 is that it has two identical receivers. That is not the case with the Orion.

My ranking is based on personal experience but of course also I also follow test reports published by W8JI, Sherwood Engineering, ARRL and others. In addition, around 80% of the dedicated and avid low band DXers who responded to my survey (see Chapter 2, Section 6) said that either the K3 or the Orion were the best radios (60% for K3, 20% for Orion). The FTdx5000 was not on the market when the survey was done, but it has the potential to be a good performer for low band DXers as well.

Yes, there are some who don't like the K3. Some say "it is too small for my shack," or "it does not have a multi-color spectral display."

When the K3 was still relatively new, the Elecraft people took a very courageous decision by equipping the — so far — greatest and most successful expedition of all time, the VP6DX DXpedition in early 2008 that made nearly 200,000 QSOs, with nothing but K3s. And the feedback from all of the operators is unanimously positive! Judging on how successful the operation was on the low bands with over 1000 160 meter QSOs into Europe (15,000 to 17,000 km distance), the K3 sure is a winner.

One area that could improve in the K3 is the sensitivity on 10 meters. At –137 dBm, it is 7 dB below the FT-1000MP MkV. Yaesu realized the need for better sensitivity on 10 meters and provided a separate RF amplifier for that band. Some degradation in dynamic range can be tolerated to provide the extra sensitivity on 10.

Gene, W3ZZ sent me the following testimony of his experience with the K3: "…Finally the strong signal handling capabilities are nothing short of amazing. I can copy weak Europeans on 160 without a receiving antenna less than 1 kHz from N3HBX who is running a kW to a 4 square less than 8 km from my QTH and is by far the loudest station I hear. I have made a test with the K3 at K3ZO's house listening to K3TW on 40 meters. Fred, K3ZO, has a 3 element Telrex at 30 meters on 40 pointed at K3TW. K3TW has a dipole at 20 meters and runs 1 KW at a distance of 350 meters from K3ZO's QTH. N3HBX estimates that the signal level at K3ZO approaches 100 mV. K3ZO's Orion — this is a fine receiver — stops working within 10 kHz of K3TW's transmit frequency. Just like a cell phone dropping a signal, all you can hear are a few squeaking noises. Beyond 10 kHz the Orion works fine. The K3 can copy weak signals that are not moving the K3 S-meter 3 kHz from K3TW — although I will say the K3 does not sound happy as there is some distortion I can hear."

4. CONCLUSION

In the previous edition of this book I tried to give some guidance for those who are new to low band DXing and pointed out some transceivers that are available on the second hand market and perform reasonably well on the low bands. Not an easy task. Things have changed nowadays. You can buy the best low band transceiver there is on the market for a very reasonable price, the Elecraft K3. And no, I have no connections with Elecraft, and paid the full price for my K3s. I would be ashamed asking for a discount for such a wonderful piece of equipment at such a very reasonable price. Next in line comes the Orion, but at a slightly higher price.

Is there still room for basic, fundamental improvements? Yes, of course. How and where do I envision these improvements?

In my view they will be mainly in the field of receiving. We now have receivers that can handle the weakest signals in close vicinity of the strongest ones. But the challenge, in my view, is to improve the signal to noise ratio by reducing or eliminating the noise that comes from outside. In my wildest dreams I see a receiver with between 4 and 8 channels, all SDR. Each of the channels is connected to an antenna, which can then be used as an element of an array. Phase and amplitude control will be done in the multiple channel receiver, not at the antenna as we do it nowadays. In addition, the receiver setup will actively adapt phase and magnitude as required for any signal, to obtain the best S/N ratio. I think we will see similar experimental setups appearing in the next few years, and being made available commercially a few years later. This will be another giant leap forward, especially for the low band addicts.

I hope I will be still there to enjoy the giant step forward.

Contents

CHAPTER 4

Antenna Design Software

L. B. Cebik, W4RNL, became a Silent Key in April 2008, one month after having helped me with reviewing this chapter. He obtained his license in 1954, and served as professor of philosophy at the University of Tennessee (Knoxville) for a quarter of a century, with interspersed administrative tasks such as serving as Assistant Dean for Research and as Director of Research Compliance. He retired early from academic life as Professor Emeritus to undertake full-time the development of his Web site (**www.cebik.com**).

He wrote extensively about antennas and antenna modeling (as well as other electronics subjects) in most of the US ham journals, including *QST*, *CQ*, *Communications Quarterly*, *QEX*, *Ham Radio*, *73*, *QRP Quarterly*, *Radio-Electronics* and *QRPp*. Besides the continuing series of antenna modeling columns he did for *antenneX*, he also wrote a column for *10-10 News* and for *QEX*.

When I asked this famous ham to be my advisor, proofreader and godfather for the section on antenna modeling, his field of expertise, L.B. immediately agreed. Well, this chapter too is in good and reputable company! Thank you, L.B., for all you did for Amateur Radio. We will miss you and your excellent articles on antenna design and modeling.

When I talk at radio clubs, I usually ask how many in the audience have a PC and how many have Internet. In 2003 the answer was somewhere around 95%. The numbers have not changed a lot in the past 10 years — it's still around 95%. The only difference is that more and more hams have more than one PC in the shack!

The PC and the Internet have both become very important tools for most active radio amateurs. There seem to be a few older generation hams (look who's talking, I'll be 70 soon) who did not jump onto the bandwagon when PCs started rolling in the late 1970s and early 1980s. Those hams who made the jump now can hardly imagine what their daily life or what their hobby would be like without one or more PCs or the Internet.

For hams, computers are not only an excellent tool to gather information via Web sites, they are also used for administrative tasks (logbooks, contest logging, etc). In this chapter we'll look at how they can be used as design tools for circuits and antennas.

1. ANTENNA-MODELING PROGRAMS

Until not too long ago, predicting antenna performance was more a black art than a scientific or engineering activity, especially in Amateur Radio circles. Building full-size models or scale models and testing them on wide-open test sites were out of reach of most amateurs. This was when some of the old myths were born and the rat race for decibels started.

What is modeling? It is evaluating the performance of a *system* that is governed by the laws of physics using a *model*. This may be a physical model (such as a scale model) or a mathematical model. Antenna-modeling programs are computer programs that via mathematics calculate and predict the performance (electrical, mechanical) of an antenna. Modeling is done in all branches of science. Modeling always has its limitations, partly because the model that we have to describe (enter into the program) can almost never be described in the same detail as the real thing (and especially its environment), and partly because of numerical limitations in the calculating code used. The final limitation is the operator, who enters the data and who interprets the results. In all cases a good deal of knowledge and experience in the field of antennas is required in order to draw the correct conclusions and take the right decisions during the process of modeling. Why do we want to model antennas?

- To understand how antennas work.
- To verify designs from literature.

• To optimize a design for your particular needs (frequency, height, application).
• To create a new design.

1.1. How Modeling Works

In an antenna model you must define the geometry of the antenna (all conductors, the feed points, the loads if any) as well as the environment in which the antenna works (free space, over perfect ground, over real ground, antenna height, etc). The basic concept is you need to describe all elements of the antenna (called *wires* in this context) by giving their X, Y and Z coordinates.

Once you have described all the elements (conductors, wires) geometrically, they will be split up into short *segments*. During modeling, the RF current in each segment is evaluated. The program calculates the self impedance and the mutual impedances (I explain what mutual impedance is in Chapter 11 covering arrays) for each of the segments. Each of these segments are considered as individual little antennas. Then the program computes the field created by the contribution from each segment. Modeling can be done in free space, over perfect ground or over real ground.

Classic antenna theory uses equations that presume a sinusoidal (cosine) curve of current distribution along an antenna element. Although the difference between the sinusoidal curve and the actual current distribution is small for relatively simple antennas, the errors become great as antennas become very long or take on complex geometries. The *method-of-moments* methods arose as a solution to the problem, since they allowed one to subdivide an element into segments and to solve essential equations taking into account the error from one segment to the next.

This section of the book is not meant to be a tutorial on how to model antennas. But it is hard to conceive that a serious Low Bander would not, sooner or later, get involved in antenna modeling. After all, the low bands are the bands where we can still do a lot of antenna building and designing. That's what makes the low bands so attractive to many.

You can learn the art of modeling by cut and try. The *EZNEC* manual is an excellent course by itself. If you are even more serious about it, visit **www.arrl.org/antenna-modeling** in the Technology section of the ARRL Web site.

Specific modeling issues, such as the required segment length, the segment length tapering technique, etc, are also covered in specific antenna chapters in this book (Verticals, Dipoles, Yagis and Quads) where relevant.

1.2. *MININEC*

1.2.1. The *MININEC* Engine

MININEC (Mini Numerical Electromagnetic Code) was developed at the NOSC (Naval Ocean Systems Center) in San Diego by J. C. Logan and J. W. Rockway. In essence it was a Basic language adaptation of *NEC* for PCs because, in the days *MININEC* was developed, we still lacked memory and space in our early PCs.

The original *MININEC* was not a user-friendly program. Several people wrote pre- and post-processing programs to make *MININEC* (now at version 3.13) more user-friendly. In later versions many initial limitations of the original *MININEC* were also overcome.

For general antenna analysis that does not press its well-known limitations, *MININEC* is a highly competent code. It handles elements of changing diameter directly, and with segment-length tapering (for example, near corners of a quad), it can accurately model a wide range of antenna geometries.

1.2.2. *MININEC* Limitations

The major limitation concerns calculations over real ground, which is limited to modeling far-field patterns. In the near field, *MININEC* assumes a perfectly conducting ground. Some of the consequences of this are that you cannot use *MININEC* to calculate the influence of radials on the feed-point impedance of a ground-mounted vertical. A quarter-wave vertical will yield a 36-Ω impedance over any type of ground. In reality the ground and the radials in the near field are important for collecting the return currents. This will influence the feed-point impedance and the efficiency of the antenna due to "lost return currents" in a poor ground. Radials can be specified with *MININEC*, but they will influence only low-angle reflection and attenuation in the far field. See Chapters 8 and 9 on dipole antennas and vertical antennas for details.

Further, *MININEC* reports the gain and the feed-point impedance of horizontally polarized antennas at low heights incorrectly. This is for horizontal antennas less than 0.2 wavelength above ground. For larger antennas the minimum height may be higher. At low heights the reported gain will be too high and the feed-point impedance too low. The shape of the radiation patterns will remain correct, however.

We thus are handicapped using *MININEC* on the low bands, where we often model antennas that are electrically close to the ground. For modeling antennas such as Yagis on higher-frequency bands, this is unlikely to be a problem because they are mounted higher than ¼ λ above ground. *MININEC* has other modeling problems with quads, which are detailed in the chapter on Yagis and Quads.

In *MININEC* wires that are thicker than 0.001 λ may not be modeled accurately due to computational approximations in the code. While low-band antennas will not be affected, this limitation may be encountered when working on antennas for 10 meters and higher. These and other limitations are very well covered by R. Lewallen in "*MININEC*: The Other Edge of the Sword" (Ref 678) and on **www.cebik.com**.

Many of the known shortcomings of *MININEC* have been compensated for in programs that use *MININEC* as an engine. Various implementations of *MININEC* show variable results, each according to the modifications introduced and the success of those modifications. L.B. Cebik, W4RNL, tested popular *MININEC*-based modeling programs and concluded that for most low frequency applications involving relatively simple antennas, most *MININEC* based programs will work well (Ref 699).

1.2.3. *MININEC*-Based Programs

Antenna Model (from Teri Software, **www.antennamodel.com**) is a full-featured and very user friendly *Windows* version of *MININEC 3.13*. The core has virtually unlimited segment capacity and uses improved algorithms to overcome many *MININEC* difficulties, fixing errors due to increasing frequency, angular junctions, wire junctions less than 28° and wires spaced closer than 0.23 λ. The program offers both 2D and 3D patterns and a variety of supplemental calculating features. *Antenna*

Model has incorporated the Sommerfeld-Norton ground simulation routine in its core. The Sommerfeld-Norton system is highly accurate, even for wires very close to the ground. This routine is also extensively used with *NEC-2*. A Smith Chart window has been added, as well as a tool that can also model transmission line losses for 70 common types of transmission line. According to Cebik, W4NRL (Ref 699) *Antenna Model* is the only *MININEC* based modeling software that passed all the benchmark tests. (Price: $90 at time of writing.)

NEC4WIN95 VM (**www.orionmicro.com**) is a *Windows* 32-bit version of *MININEC* running under *Windows* NT/2000, XP and *Vista*, using spreadsheet input page and pull-down boxes for other antenna parameters. 3D patterns are provided, as well as optimization routines. The user can vary the height of the antenna without invoking a complete recalculation of the matrix for faster results. There is a built-in loop correction feature allowing accurate modeling of square-loop antennas. The newer VM (virtual memory) version of the program permits almost unlimited numbers of segments in a model. (Price: $50 at time of writing.)

MMANA-GAL (by JE3HHT, DL1PBD and DL2KQ) is *freeware* (**mmhamsoft.amateur-radio.ca**). Based upon the *MININEC 3.13* core, the program offers a large segment (pulse) capacity and other advanced features, such as segment-length tapering, optimizing and network calculation. It is a very popular program as it is freeware, and it yields excellent results. It also includes a weighted optimization routine.

Expert MININEC Classic (**www.emsci.com**) is also freeware. It has some important limitations though: max 500 wires and no tapering.

ELNEC (**www.eznec.com**) is a DOS modeling program by Roy Lewallen, W7EL, based on *MININEC*. Note that W7EL doesn't actively market *ELNEC* any more (see *EZNEC* in Section 1.3.1.).

For more detailed information, visit **www.cebik.com/model/nec.html**.

1.3. Programs Using the *NEC-2* Core

NEC is the full-fledged brother of *MININEC*, which means that *NEC* also employs the method-of-moments to model antennas. The original versions ran on mainframe computers only, and were accessible to professionals only. They had a very unfriendly user interface. In the last decade, however, a number of user-friendly *NEC*-based programs have been developed.

NEC-2 (available since 1981), which is in the public domain, can model real ground in the near and far fields. It does away with most of the limitations described above for *MININEC*. It can model antennas quite close to the ground, as well as radials above and almost on the ground. (It cannot handle buried radials though.) *NEC-2* uses the Sommerfeld-Norton high-accuracy ground model to model horizontal wires close to the earth. One notable limitation of *NEC-2*, compared to *MININEC*, is its inability to model stepped-diameter wires (such as tapered Yagi elements) although this shortcoming has been overcome by some software providers using the *NEC-2* core.

This problem has also been corrected in the newest version *NEC-4* (available since 1992), which also has the ability to model wires in the ground. I have frequently used *NEC* to model antennas where the limitation of *MININEC* would have made the results unreliable. I will review specific modeling issues when discussing those antennas (for example, Beverages, low delta loops, elevated radials, etc).

MININEC 3.13 shows its strength in the areas where *NEC-2* displays weaknesses: stepped-diameter wire models mainly. It must be said, however, that for a large class of modeling tasks both *NEC* and *MININEC* are equally capable.

1.3.1. *NEC-2* Based Programs

EZNEC (now at version 5) is written by Roy Lewallen, W7EL, who has been writing well-received modeling software for a long time. *EZNEC* offers 3D plots, 2D slicing, ground-wave output, direct entry for trap as well as for series and parallel R-L-C loads, stepped-diameter correction, and numerous short cuts for antenna-geometry modification. Standard *EZNEC Version 5* is restricted to 500 segments, while the version *EZNEC+ V5* can handle up to 1500 segments. The *EZNEC Pro V5* handles much up to 20,000 segments but it takes a large drive and a lot of RAM to make use of. **Fig 4-1** shows the "View Antenna" screen of a model representing 300-meter long Beverage antenna for 160 meters. This model uses two quarter-wave in-line terminations at each end (see Chapter 7). For more details on this very user friendly and high performance modeling software, visit **www.eznec.com**. (Prices at time of writing: *EZNEC V5,* $89; *EZNEC + V5*, $139; *EZNEC Pro V5* for *NEC-2*, $500.) In various chapters of this book reference is made to *EZNEC* modeling files that are available on the CD that comes with this book.

One of the major assets of the later versions of *EZNEC* is that you can include real feed lines, L-networks and transformers in your model. This make it possible to make a swept frequency analysis of an antenna system (or array) that includes the elements that make up the system (or array). In Chapter 11 of this book (arrays) I make ample use of this possibility to assess the bandwidth performance of arrays.

EZNEC-ARRL is the *EZNEC* version 4 published by the ARRL as part of the CD that came with the *ARRL Antenna Book* (21st edition) . It operates as a normal *EZNEC* demo program except when a specially "signed" *EZNEC* description file (included on the *ARRL Antenna Book* CD) is loaded. When analyzing a "signed" file, *EZNEC-ARRL* becomes a fully

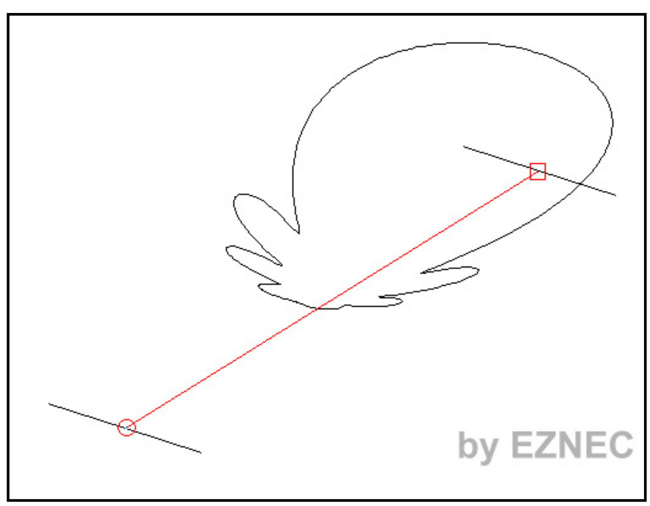


Fig 4-1 — "View Antenna" screen in the *EZNEC 3* program of a model representing a 300-meter long Beverage for 160 meters, using two quarter-wave in-line terminations at each end.

functional standard type *EZNEC* program. Both *EZNEC-ARRL* V3.0 and *EZNEC-ARRL* V4.0 will function with "signed" files from any *ARRL Antenna Book* edition. For more details go to **www.eznec.com/eznec_arrl.htm**.

NEC-Win Plus (now at V 1.2) by Nittany Scientific (**www.nittany-scientific.com**) is another popular *Windows* version of *NEC-2* that features spreadsheet-type input pages with design-by-equation capabilities. The program also offers stepped-diameter corrections. It provides 2D and 3D plots and antenna views and graphical outputs. *NEC-Win Pro* is the high-end version of *NEC-Win Plus*. (Prices at time of writing: *NEC-Win Plus*, $150; *NEC-Win Pro*, $425.)

4nec2 by Arie (**home.ict.nl/~arivoors**) is another user-friendly shell wrapped around the standard *NEC-2* computing engine. It can also be used with the *NEC-4* engine. *4nec2* is freeware that is in a continuous development phase. The *4nec2* package contains all the software to specify, calculate, evaluate and optimize your antenna system(for a single frequency or a band of frequencies). It is capable of modeling *NEC-2 or NEC-4* files up to 11,000 segments. It produces some of the most spectacular 3D radiation patterns (both near field and far field) and is used in different places to create 3D radiation patterns throughout this book. **Fig 4-2** shows an example. A powerful geometry builder tool is also available to automatically create complex geometry structures. Frequency sweep tables can be generated (both linear as well as logarithmic) for gain, SWR, efficiency and F/B or F/R ratio. *4Nec2* also includes a graphical editor. Finally the software is also equipped with a gradient style as well as a generic algorithm based optimizer to optimize an antenna design. Let me also mention the Smith Chart display with integrated line-length calculator. Furthermore *4nec2* can calculate the series- or stub-matching or Pi, L or Tee network required to get an optimal match for your antenna to the line. This is without any doubt one of the most powerful packages around, and the price is very reasonable (freeware).

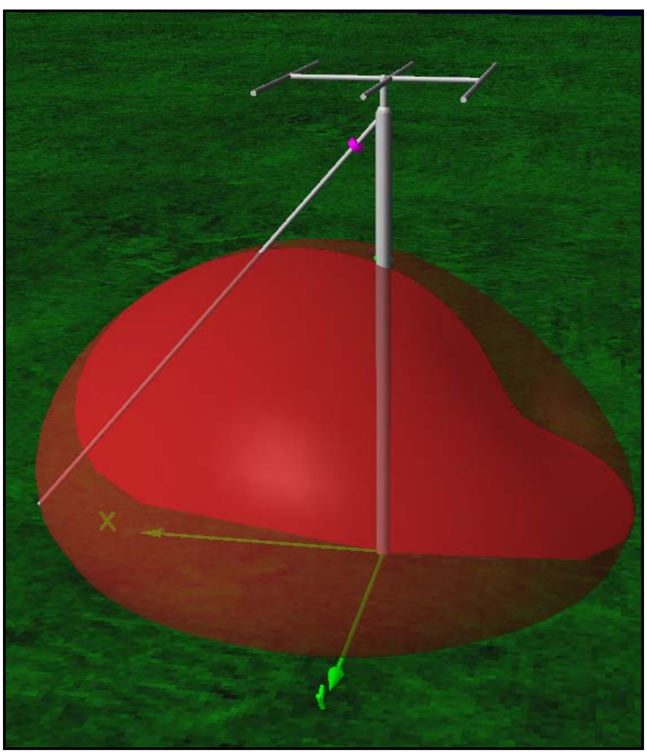


Fig 4-2 — The freeware program *4nec2* makes it possible to create awesome looking 3D radiation plots In this example we see a "sloper" antenna, the 3-D radiation body and a vertical cut of the body.

NEC-2 for MMANA by Dimitry Fedorov, UA3AVR, (**www.qsl.net/ua3avr**) is a freeware utility allowing you to enjoy all the benefits of the *NEC-2* core while using modeling files created for *MMANA* (described above). The *NEC-2* core was improved in different areas and also includes the Sommerfeld-Norton ground model.

Nec2Go by Nova Plus software. A free limited performance demo model is available on **www.nec2go.com**. A special feature is that it uses ant file input definitions similar to the format for AO (*Antenna Optimizer* by K6STI). (Price at time of writing: $39.95.)

Antenna Solver (by Grating Solver, **www.gsolver.com**) is a freeware program written in C++ and developed by KJ5AT that uses a graphical user interface that provides much greater flexibility in the construction, analysis and interpretation of radiating structures than *NEC* alone. The Help files that are distributed with *Antenna Solver* include several step-by-step (click-by-click) examples that explore many *Antenna Solver* features.

NEC-Win Synth (NWS) by Nittany Scientific (**www.nittany-scientific.com**) is a wire grid model generating software. It is not an antenna modeling program in itself but rather a program to synthesize wire-grid structures for use in any *NEC* (-2/-4) program. The user may select a preset shape and enter critical dimensions or synthesize a structure with the spreadsheet entry facility. *NEC-Win Synth* can be directly linked to *NEC-Win Plus* or save its output in a standard *NEC* file or *EZNEC* geometry files to be imported into other *NEC* based analysis programs such as *EZNEC* or *NEC-Win Plus*. (Price at time of writing: $99.)

1.4. The *NEC-4* Core

The latest version of *NEC* is *NEC-4* (1992), which overcomes most of the shortcomings with earlier *NEC-2* codes. Using *NEC-4* one can now do accurate modeling of:

- close to the ground as well as buried radial systems
- elements of varying diameter sections
- close-spaced parallel wires
- insulated wires

NEC-2 does not allow modeling antennas with tapered diameter elements. Therefore such a tapered element had to be substituted with a constant diameter element (using the K2BT or the Leeson algorithm) prior to running *NEC-2*. This however can only be done with elements having a sinusoidal current distribution, which is not the case of elements having loading coils or traps. In *NEC-4* you can model such antennas directly however.

NEC-4 is also said to be more tolerant of certain geometry configurations, which means that you would get meaningful and correct results when *NEC-2* would fail to do that.

Being able to model close to ground and buried radials makes *NEC-4* in principle the best software for modeling low band antennas. Roy Lewallen, W7EL, author of *EZNEC* however finds that the overly simplistic ground model really limits its usefulness. At best you can do some rough comparison of various types of ground systems. This is because in reality the ground properties are very changeable from one spot to another, even if those spots are only separated by fractions of wavelengths. The issue of modeling radial systems will be covered in detail in Chapter 9 (vertical antennas) and Chapter 11 (vertical arrays).

While *NEC-2* is public-domain software, the copyright for *NEC-4* is held by Lawrence Livermore National Labs and you must obtain a license to use either *NEC-4* or any of the other software packages that use the *NEC-4* core. For US citizens the license fee at the time of writing is $300 for academic and noncommercial applications and $500 for customers outside the US. For more information, see **ipo.llnl.gov**. Click Technologies, then Software then Browse Software to locate the *NEC-4* licensing information.

1.4.1. *NEC-4* Based Programs

EZNEC Pro (now at version 5), by Roy Lewallen, W7EL, (**www.eznec.com**) is normally equipped with the *NEC-2* engine. You can buy *EZNEC Pro* with an option for *NEC-4*, if you can show a license for *NEC-4* (obtained from Livermore National Labs, see Section 1.4). *EZNEC Pro* imports and exports files in generic *.NEC format as well as *.EZ format.

4nec2 also works with the NEC-4 engine, although the name may be a little confusing,. All you need to do is to obtain a *NEC-4* license (see Section 1.4.).

GNEC (**www.nittany-scientific.com**) is the *NEC-4* version of *NEC-Win Plus*. This program implements all or nearly all of the input "cards" of the complete *NEC-4* input deck. Output capabilities include 3D, polar plots and many rectangular (X-Y) graphs, as well as a large array of tabular reports. The spreadsheet and dialogue box interface is similar to *NEC-Win Pro*. Here too a license for the *NEC-4* core must be purchased separately.

Considering the price you need to pay for the license and for the software that runs *NEC-4*, I would like to say that there are very few circumstances that a ham would really need it to do his antenna modeling. As a professional antenna designer, especially when confronted with low band antennas and radials, it is, however, the way to go. After all the *NEC-4* license fee has come down to a more or less acceptable level.

1.5. Optimizing Programs

With a regular *MININEC* or *NEC*-based program, you will have to spend quite some time if you want to optimize a design for a given parameter (whether that is gain, F/B or maybe impedance or SWR bandwidth). This is exactly what I used to do almost 30 years ago with my Apple 2e and the original *MININEC*, where I would use batch files, that would do repetitive modeling of an antenna using slight changes in dimension. In those days, one run of a 5 element Yagi took about 30 minutes. Now it is 0.3 second.

Optimizing programs can be quite helpful. *YO* (*Yagi Optimizer*) by K6STI and *YagiMax* by K4VX were the first popular optimizing programs, but they works only for monoband Yagi antennas. *AO* (*Antenna Optimizer*) is a similar program, but it works for any type of antenna. Both are based on *MININEC* and are no longer available nor supported by the author.

MultiNEC by AC6LA was based on an *Excel* spreadsheet program and was a very powerful program that could steer a number of modeling programs such as *NEC-Win Plus*, *EZNEC*, *4nec2* and *Antenna Model* doing intelligent user defined "batch processing." Unfortunately Dan, AC6LA, has stopped all further development and support for the time being.

At this time of writing, the only programs that provide optimization routines are:

4nec2 is a *NEC-2* based program (freeware) that can be used to calculate all parameters for a frequency range you specify (sweep function). It also includes a weighted optimization system for both a single frequency and a band of frequencies.

NEC-2 for MMANA has a weighted optimization routine built-in.

1.6. Other Antenna Programs

Arrayfeed designs feed systems for phased arrays (Four Square, 2-element end-fire and rectangular 4). It is offered as freeware on the CD coming with the 21st edition of the *ARRL* A*ntenna Book.*. For more details go to **www.eznec.com/eznec_arrl.htm** where *Arrayfeed1.exe* can also be downloaded. This programs is also used in Chapter 11 to calculate feed systems of different types of arrays.

Moxon Rectangle Generator by AC6LA (**www.ac6la.com/moxgen.html**). In Chapter 13 I will cover the Moxon antenna in detail. The *Moxon Rectangle Generator* (*MoxGen*) is a standalone program written by AC6LA after a lot of research and modeling work done by Cebik. It calculates the dimensions of a Moxon rectangle for a near 50 Ω feed point impedance. The program will also create a model in .EZ format for use with *EZNEC* or in .NEC format for use with *NEC*-based software

1.7. Antenna Modeling, The One and Only Truth?

Modeling programs have been around for more than 30 years. Performance has gradually improved all along: more friendly interfaces, more accurate, faster, with all bells and whistles, "sexy" diagrams and patterns, you name it. I find them very useful tools for testing a "basic" or a "generic" design. However, it is a good thing to also understand the physics that are involved.

Nowadays I sometimes have the impression that modeling programs are used to test (to question) the laws of physics, and also the laws of simple logical reasoning. I see this especially happening in the field of radials for vertical systems. Now that *NEC-4* is said to be able to model buried wires "accurately," it seems to me that modeling ground systems has become a hobby in itself. Rarely have I seen any such articles where the modeling is complemented with real life field testing (field strength measurements). After all, isn't real life performance what we are after?

When I read in such an article that model A is better than model B because the system gain is 0.1 dB higher, it makes me smile. I have one piece of advice: let's keep our two feet on the ground. Let's use the best available modeling software to guide us, but let us also be aware that we are modeling, not testing, not measuring. Let us also try to understand the physics behind it all. In the professional world, where resources are next to unlimited, mathematical modeling is always followed up by model building (full scale or not) and extensive physical testing. Radio amateurs can hardly ever do this, at least not in great detail.

It seems to me that in real life we, hams, get the best confirmation regarding a design we modeled by observing the performance of the antenna in contesting. As they say, "the proof of the pudding is in the eating." Writing a software program that tells how to make a pudding is one thing; enjoying the pudding is another. Most of us like to enjoy the pudding, I think.

2. ON4UN LOW-BAND SOFTWARE

The nice thing about personal computers is that everyone can now handle the difficult mathematics pertaining to antennas and feed lines. All you need to do is understand the question... and the answers. The programs will do the hard mathematics for you and give you answers that you can understand. The theory of antennas and feed lines is not an easy subject.

We have all been brought up to know how much is 5 times 4. But nobody can tell off the top of his head how much $5 - j3$ times $12 + j12$ is. At least I cannot. When I started studying antennas and wanted not only to understand the theory, but also to be able to calculate things, I was immediately confronted with the problem of complex mathematics. While studying the subject I wrote a number of small computer programs to do complex-number calculations. They have since evolved to quite comprehensive engineering tools that should be part of the software library of every serious antenna builder.

The *New Low Band Software* is based on the original *Low Band DXing Software* I wrote in the mid 1980s, while preparing the original *Low Band DXing* book. The latest software (version 1994) is a very much enhanced and much more user friendly. I wrote it under DOS using *Q-Basic* and it runs well in a DOS box on modern machines, also operating under *Windows* XP, but not *Windows Vista* or *Windows 7*. When you start it up in XP, you will have it in a small window. If you want to see it run full screen, just start the program *2EL and 4 EL Vertical Arrays*, and you will have that module running in full screen. Now press X for Exit and you have the menu in full screen now (compatibility is a nice thing).

Each of the modules starts with a complete on-screen introduction, telling what the software is meant to do and how to use it. All propagation-related programs are integrated into a single module. There are many help screens in each of the modules. This software package is freely available on the CD that comes with the book.

Various modules are used throughout the book for making calculations, and the readers are invited to use the software step by step as explained in the text.

2.1. Propagation Software

The propagation software module is covered in detail in Chapter 2. The module contains a low-band dedicated sunrise/sunset program. It also has a gray-line program, based on a comprehensive database containing coordinates for over 550 locations, and which can be user changed or updated. The database can contain up to 750 locations.

2.2. Mutual Impedance and Driving Impedance

From a number of impedance measurements you can calculate the mutual impedance and eventually, knowing the antenna currents (magnitude and phase), you can calculate the driving impedance of each element of an array with up to four elements.

A new spreadsheet program is now available that will make the same calculations for arrays with up to nine elements. See the file **w1mk-on4un-oh1tv-arrays.xls**, included on the CD that comes with this book.

2.3. Coax Transformer/Smith Chart

The original software covered only ideal (lossless) cables. The later versions include feed lines with loss. The real cable program will tell you everything about a feed line. You can analyze the feed line as seen from the generator (transmitter) or from the load (antenna). Impedance, voltage and currents are shown in both rectangular coordinates (real and imaginary parts) or in polar coordinates (magnitude and phase angle). You will see the Z, I and E values at the end of the line, the SWR (at the load and at the generator), as well as the loss — divided into "intrinsic" cable loss and "additional" SWR loss.

A number of "classic" coaxial feed lines with their transmission parameters (impedance, loss) are part of the program, but you can specify your own cable as well. Try a 200-foot RG-58 feed line on 28 MHz with a 2:1 SWR and compare it to a ¾-inch Hardline with the same length and SWR, and find out for yourself that a "big" coax is not necessarily there just for power reasons. It makes no sense throwing away 2 or 3 dB of signal if you have spent a lot of effort building a top performance antenna. If you are going to design your own array, you will probably use this software module more than any other.

2.4. Impedance, Current and Voltage Along Feed Lines

Again, there are two versions of each module: loss-free and "real" cable.

2.4.1. Z, I and E Listings

A coaxial cable, when not operated as a "flat" line (that is, it has an SWR greater than 1:1) acts as a transformer: The impedance, current and voltage are different at each point along the cable. You enter the feed-line data (impedance, attenuation data), the load data (impedance and current or voltage), and the program will display Z, I and E at any point of the cable.

2.4.2. Simultaneous Voltage Listing Along Feed Lines

This module was written especially as a help for designing a Christman (K3LC) feed system for driven arrays. The program lists the voltage along feed lines, allowing the user to find points on the feed lines of individual array elements where the voltages are identical. These are the points where the feed lines can be connected in parallel (see Chapter 11 on arrays). This program is also helpful to see how high the voltage really rises on your feed line with a 4.5:1 SWR, for example.

2.5. Two- and Four-Element Vertical Arrays

These modules take you step-by-step through the theory and practical realization of a 2-element (cardioid) or 4-element (Four Square) array, using the W7EL feed system. This tutorial and engineering program uses graphic displays to show the layout of the antenna with all the relevant electrical data. This unique module is extremely valuable if you want to understand arrays and if you want to build your own array with a feed system that really works.

2.6. The L Network

The L network is the most widely used matching network for matching feed lines and antennas. The module gives you all the L-network solutions for a given matching problem.

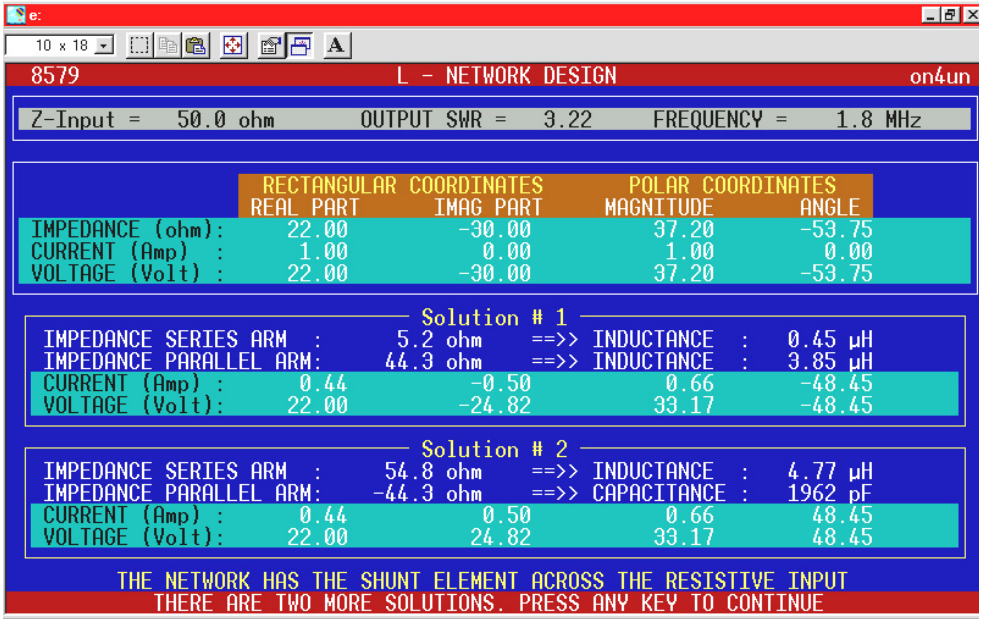


Fig 4-3 — Screen capture of the L-network module of the ON4UN *New Low Band Software*. This module is used very extensively in Chapter 11 for arrays. All relevant data (Impedance, Current and Voltage) are shown in both Cartesian (a + *j*b) as well as polar coordinates (a ∠ b°).

The software also displays voltage and current at the input and output of the network, which can be valuable to assess component ratings in the network. See **Fig 4-3**.

2.7. Series/Shunt Input L-Network Iteration

This module was written especially for use in the Gehrke array-matching system, where L networks are used to provide a desired voltage magnitude at the input of the network, given an output impedance and output voltage. **Fig 4-4** shows an example. See Chapter 11 on phased arrays for details.

2.8. Shunt/Series Impedance Network

This is a simplified form of the L network, where a perfect match can be obtained with only a series or a shunt reactive element. It is also used in the modified Lewallen phase-adjusting network with arrays that are not quadrature fed (see Chapter 11 on vertical arrays).

2.9. Line Stretcher (Pi and T)

Line stretchers are constant-impedance transformers that provide a desired voltage phase shift. These networks are used in specific array feed systems (modified Lewallen method) to provide the required phase delay. See Chapter 11 on vertical arrays for details.

2.10. Stub Matching

Stub matching is a very attractive method of feed-line matching. This module facilitates matching a feed line to a load using a single stub placed along the transmission line. It is very handy for making a stub-matching system with an open-wire line feeding a high-impedance load (2000 to 5000 Ω).

2.11. Parallel Impedances (T Junction)

This module calculates the impedance resulting from connecting in parallel a number of impedances. Do you really want to calculate on your calculator the value of 21 – *j* 34 and 78 + *j* 34 ohms in parallel?

2.12. SWR Value and SWR Iteration

2.12.1. SWR Value

This calculates the SWR (for example, the SWR for a load of 34 – *j* 12 Ω on a 75-Ω line). The mathematics are not complicated, but it's so much faster with the program (and error free!).

Fig 4-4 — John, K9DX, using the Shunt/Series impedance network module for designing the feed system of his 9-circle array (see Chapter 11).

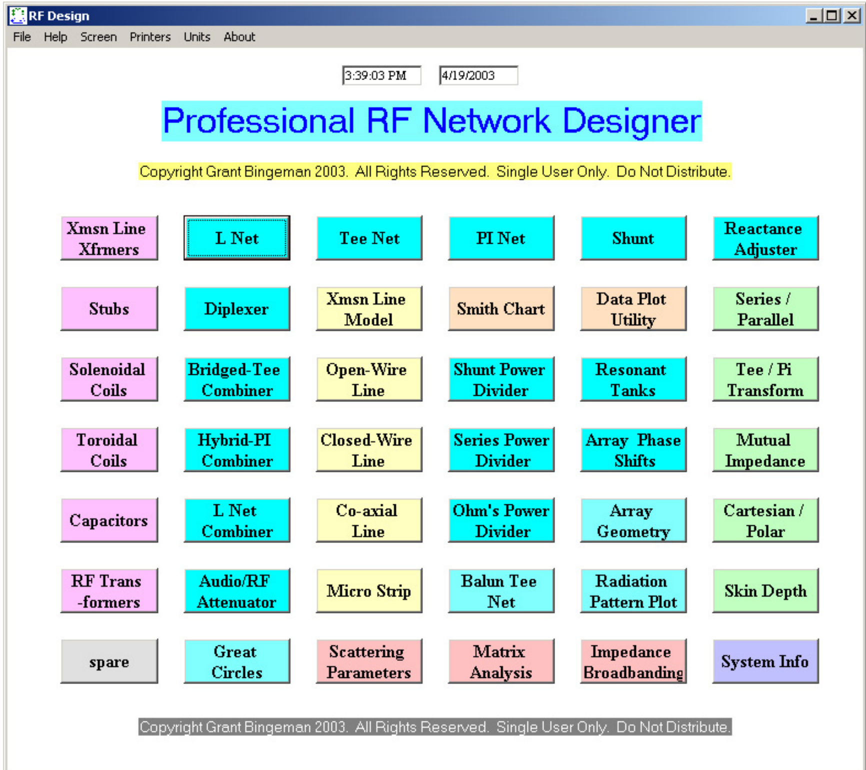


Fig 4-5 — Opening screen of the *Professional RF Network Designer* program by Grant Bingeman, KM5KG.

2.12.2. SWR Iteration

This module was especially developed for use when designing a W1FC feed system for an array (using a hybrid coupler). See Chapter 11 on arrays for details.

2.13. Radiation Angle for Horizontal Antennas

This module calculates and displays the vertical radiation pattern of single or stacked antennas (fed in phase).

2.14. Coil Calculation

With this module you can calculate single-layer coils and toroidal coils. It works in both directions (coil data from required inductance, or inductance from coil data).

2.15. Gamma-Omega and Hairpin Matching

With this module, given the impedance of a Yagi and the diameter of the driven element (in the center), you can design and prune gamma or omega or hairpin matches and see the results as if you were standing on a tower doing all the pruning and tweaking.

2.16. Element Taper

Antennas made of elements with tapering diameters show a different electrical length than if the element diameters had a constant diameter. This module calculates the electrical length of an element (quarter-wave vertical or half-wave dipole) made of sections with a tapering diameter. A modified W2PV tapering algorithm is used.

The *New Low Band Software* is available on the CD that comes with this book.

3. *PROFESSIONAL RF NETWORK DESIGNER* BY KM5KG

Grant Bingeman, KM5KG, is a professional broadcast-antenna engineer who wrote a series of what we could call *utility programs*, similar to those in my software package *New Low Band Software*. These programs can greatly ease some of the tedium of RF and antenna system design. *Professional RF Network Designer* (**Fig 4-5**) is a versatile *Windows* program.

A short description of this powerful software can be found on **www.km5kg.com/networks.htm**. At the time of writing, *Professional RF Network Designer* was also available through Array Solutions (**www.arraysolutions.com**).

With permission, I quote a short review by L.B. Cebik, W4RNL: "The buttons on the main screen are color coded by groups of related calculation sets. On the left are component calculations. The individual entries are unusually complete. For example, the capacitor entry not only provides calculations for standard two-plate capacitors, but also concentric tubing capacitors as well. If you have never explored the relative frequency sensitivity of these two capacitor types, running some values over a large frequency span can be instructive. The middle of the upper-most row covers basic networks, while column 2 (counting from the left) provides entry into combiners and diplexers. The third column permits the user to custom design or analyze most forms of common transmission line configurations. The remaining columns below the top row provide an array of useful utilities, including Smith Chart analysis, Cartesian-to-polar (and back) conversions, and series/

parallel tank circuit equivalencies.

"The individual calculation sets do not limit themselves to ideal lossless cases, but include all standard loss calculations as part of each exercise. There are a few special features worth noting in individual modules.

"In all, *Professional RF Network Designer* is a very useful tool for anyone designing or analyzing RF and antenna system circuitry, whether professional or amateur. Indeed, it is about the best of such tools that I have so far had a chance to sample or own. I highly recommend it. More importantly, I highly recommend that every purchaser spend a good bit of time with the program, sampling not only what features are available, but as well how networks operate. It only takes a systematic variation of the input variables of any module to acquire an appreciation and reasonable expectation for network variations with changing conditions. After this self-education will emerge a host of applications that we might not have previously imagined possible."

Contents

CHAPTER 5

Antennas: General, Terms, Definitions

Lew Gordon, K4VX, needs no introduction to antenna designers and builders, nor to the contest community. I first met Lew through his excellent *YagiMax* modeling software, which was very advanced 20 years ago and then already included an optimizing algorithm.

Fifteen years ago, when I evolved from an avid low-band DXer into an even more avid contester, Lew's multiop contesting station in Missouri was an outstanding example of station and antenna design. It ranked with stations like W3LPL and K3LR.

When I met Lew and his wife Terry for the first time during WRTC (World Radiosport Team Championship) in San Francisco in the summer of 1996, I met a fine gentleman and a charming lady. When I asked Lew to godfather a chapter of my new book, he immediately and enthusiastically accepted. This time again I found Lew to be willing to be my critic, my adviser and my proofreader.

Lew was first licensed as W9APY in 1947 at age 17. He graduated from Purdue University with a physics major. His professional career was as an RF systems engineer with the US government. Lew has also held the calls WA4RPK and W4ZCY. His antenna systems near Hannibal, Missouri, utilize a total of 10 towers ranging from 50 to 170 feet in height. Although he professes to be mainly a contester instead of a DXer, his DXCC total stands at 359 confirmed.

Thank you, Lew, for your help and encouragement.

Agreeing on terms and definitions is important. Too many technical discussions seem to take place in the tower of Babble. First make sure you speak the same language; then speak. Before we get involved in a debate on what's the best antenna for the low bands (that must be the key question for most), we define what we want an antenna to do for us and how we will measure its performance.

Making antennas for the low bands is one area in Amateur Radio where home building can yield results that can substantially outperform most of what can be obtained commercially. All my antennas are homemade. Visitors often ask me, "Where do you buy the parts?" Or, "Do you have a machine shop to do all the mechanical work?" Very often I don't buy parts. And no, I don't have a machine shop, just run-of-the-mill hand tools. But my friends who are antenna builders and I keep our eyes open all the time for goodies that might be useful for our next antenna project. There is a very active swap activity among us. We have access to certain facilities that make antenna building easier. It's almost like we are a team, where each one of us has his own specialty.

Don't look at low-band antenna designing and building as a "kit project." You need some know-how, a good deal of imagination and inventiveness and often some organizational talent. But unlike the area of receivers and transmitters, where we homebuilders do not usually have access to custom-designed integrated circuits and other very specialized parts, we can build antennas and antenna systems using materials found locally.

A number of successful antennas for the low bands are described in this book. These are not meant to be kits with step-by-step instructions, but are there to stimulate thinking and to put the newcomer to antenna building on the right track.

The antenna chapters of *Low-Band DXing* emphasize typical aspects of low-band antennas, and explain how and why some of the popular antennas work and what we can do to get the best results, given typical constraints. *The ARRL Antenna Book* (Ref 697) contains a wealth of excellent and accurate information on antennas.

1. THE PURPOSE OF AN ANTENNA

1.1. Transmitting Antennas

A transmitting antenna should radiate *all* the RF energy supplied to it *in the desired direction*, at the *required elevation angle* (directivity). We want to be loud; the issue is *gain*. We can do this by concentrating our RF in a given direction (in both the vertical and the horizontal planes).

1.1.1. Wanted Direction

1.1.1.1 Horizontal Directivity

We learned in Chapter 1 (Propagation) that on the low bands, paths quite frequently deviate from the theoretical great-circle direction. This is especially so for paths going through or very near the auroral oval (such as West Coast or Mid-West USA to Europe). This is a fact we have to take into consideration for a fixed-direction antenna. For paths near the antipodes, signal direction can change as much as 180° (with every direction in-between) depending on the season. All this must be taken into account when designing an antenna system. Rotary systems, of course, provide the ultimate in flexibility so far as horizontal directivity is concerned.

I want to emphasize that the term *horizontal directivity* is really meaningless without further definition. Azimuthal directivity at a takeoff angle of 0° (perfectly parallel to the horizon) is of very little use, since practical antennas produce very little signal at a 0° wave angle over real ground. This issue is important when designing or modeling an antenna. It would be ideal to design an antenna that concentrates transmitted energy at a relatively low angle, while exhibiting the highest rejection off the back at a much higher angle (to achieve maximum rejection of stronger local signals, which as a rule come in at a much higher wave angle. Horizontal directivity should always be specified at a given elevation angle. An antenna can have quite different azimuthal directional properties at different elevation angles.

We will see further that a very low dipole radiates most of its energy directly overhead at 90° (zenith angle), and shows no directivity at high wave angles (60° to 90°). The same antenna, at the same height, shows a pronounced directivity (hardly any signal off the ends of the dipole) at very low wave angles, but hardly radiates at all at very low elevation angles. These issues must be very clear in our minds if we want to understand radiation patterns of antennas.

1.1.1.2. Vertical Directivity

In the last few years a lot of modeling has been done using various propagation software packages. At ARRL HQ, D. Straw, N6BV, used *IONCAP* (Ionospheric Propagation Analysis and Prediction System) and *VOACAP* (a version of *IONCAP* upgraded by the Voice of America) to calculate elevation angles for various paths on the different MUF-controlled amateur bands (which excludes 160 meters). *IONCAP* is based on a mass of propagation data collected over more than 35 years. **Table 5-1** shows the distribution of elevation angles on 40 and 80 meters for some typical DX paths, as does **Fig 5-1** in graphical form. This elevation-angle statistical information is derived from the data on the CD-ROM included with the 20th Edition of *The ARRL Antenna Book* (Ref 697). The data obtained from *IONCAP* were modeled using isotropic antennas (no gain, no directivity) at both ends.

Because of the use of isotropic radiators in *IONCAP*, the range of elevation angles is limited only by the propagation "possibilities" and *not* by the antenna used at either the transmitting or receiving site. In other words, the charts assume a hypothetical antenna transmits and receives as well at a 1° wave angle as it does at 10°, 20° or 30° angles. An isotropic antenna, of course, does not actually exist, although a vertical over saltwater or a high horizontal antenna over a sloping terrain can approach such performance.

"No Data" in Table 5-1 means that there are no data available from the model. This does *not* mean that there is *no* possibility of propagation. On 80 and 40 meters propagation is possible from any point in the world to any other point in the world, given the right moment of the year and the right time of the day, under good propagation conditions, even though such propagation may not be statistically "significant." After all, low-band hams thrive on adversity and they love to pursue openings that are not shown in the statistics!

The elevation-angle distributions are based on statistical figures for various levels of solar activity over an entire solar cycle, and for various times and months. These distributions assume undisturbed geomagnetic conditions. There is anecdotal evidence that the prevailing elevation angles go higher during disturbed conditions. You will note that there is no statistical information for 160 meters, mainly because *IONCAP* and its derivatives are programs based on MUF (and 160 meters is *not* influenced by the MUF) and do not take into account the influence of Earth's varying magnetic field, which is crucially important on Top Band.

1.1.1.3. 40 Meters

Now that we know the range of angles we need to cover, let's have a look at how we could do this. Wave angles of 1° to 20° (except for the path from the US East Coast to Europe,

Table 5-1
Range of Radiation Angles for 40 and 80 Meters for Various Paths

The values are averages across the complete sunspot cycle and across the seasons. The value between parentheses is the most common radiation angle (peak value in the distribution).

From	*Path to*	*40 Meters*	*80 Meters*
W. Europe (Belgium)	Southern Africa	1-18 (5)	1-17 (5)
	Japan	1-19 (3)	2-17 (3)
	Oceania	1-4 (1)	No Data
	South Asia	1-17 (4)	3-5 (4)
	USA (W1-W6)	2-33 (5)	1-35 (4)
	South America	1-17 (1)	1-12 (1)
USA East Coast	Southern Africa	1-16 (3)	3-4 (4)
	Japan	1-15 (1)	1-12 (5)
	Oceania	1-9 (1)	No Data
	South Asia	1-9 (1)	No Data
	South America	1-23 (5)	1-21 (10)
	Europe	1-38 (6)	1-31 (13)
USA Midwest	Southern Africa	1-8 (4)	No Data
	Japan	1-17 (2)	1-17 (1)
	Oceania	1-12 (3)	No Data
	South Asia	No Data	No Data
	South America	2-21 (4)	1-16 (4)
	Europe	1-29 (1)	1-34 (13)
USA West Coast	Southern Africa	1-4 (1)	No Data
	Japan	1-27 (5)	2-27 (10)
	Oceania	1-17 (2)	No Data
	South Asia	1-16 (4)	No Data
	South America	1-16 (6)	1-8 (1)
	Europe	1-21 (5)	1-23 (4)

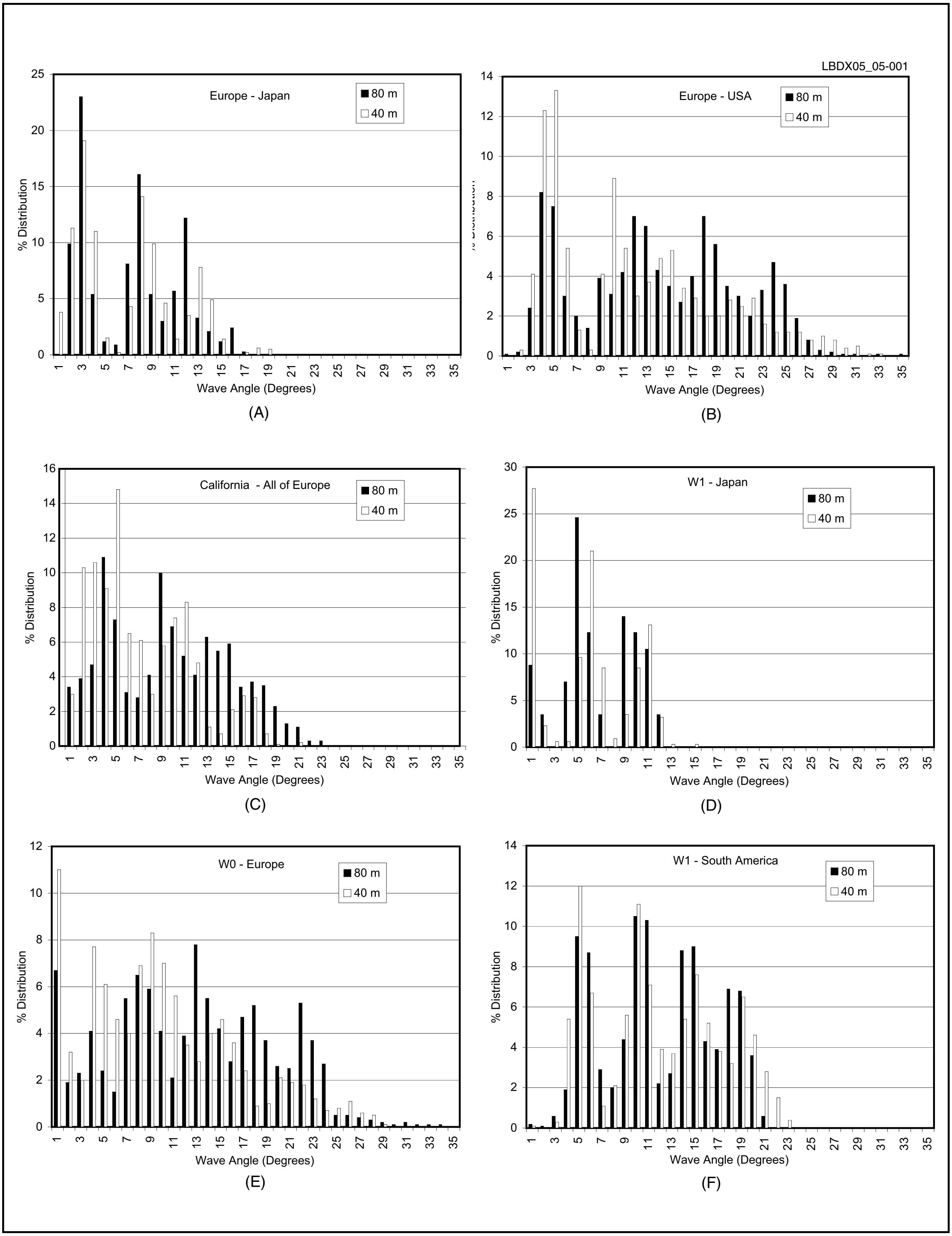


Fig 5-1 — Distribution of wave angles (elevation angles) for a few common paths on 80 and 40 meters. Notice that the distribution is not a Gaussian one. This is because many mechanisms are involved that are totally unrelated.

where the range extends to 30°) seem to be most common on 40 meters. Let's analyze how we might achieve this range over *flat terrain*. We'll take a look at three common types of antennas: A dipole, a 2-element Yagi and a λ/4 vertical over average ground.

To work at the lower angles, you need an impressively high horizontal antenna to match the wave angle distribution. In **Fig 5-2**, only the 60-meter high dipole comes relatively close to matching the statistics for the path from the US West Coast to all of Europe on 40 meters.

Fig 5-3 shows even better matches, but look at the heights involved. The stack of 2-element Yagis at 45 and 60 meters is at least 12 dB better than our 30-meter high dipole for wave angles of 5° and less!

What about verticals? **Fig 5-4** shows a single quarter-wave vertical, over very good ground with 100 λ/4 radials. This is still a poor match to the wave-angle distribution. Now,

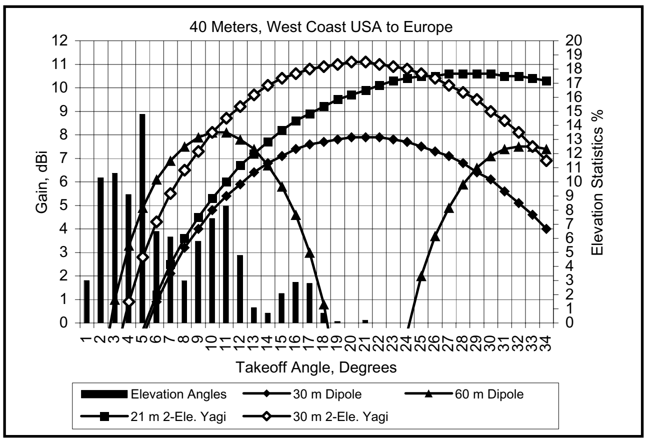


Fig 5-2 — The statistical distribution of elevation angles for the 40-meter path from Europe to the US West Coast (San Francisco), compared with the elevation responses for horizontally polarized antennas at several heights over flat ground.

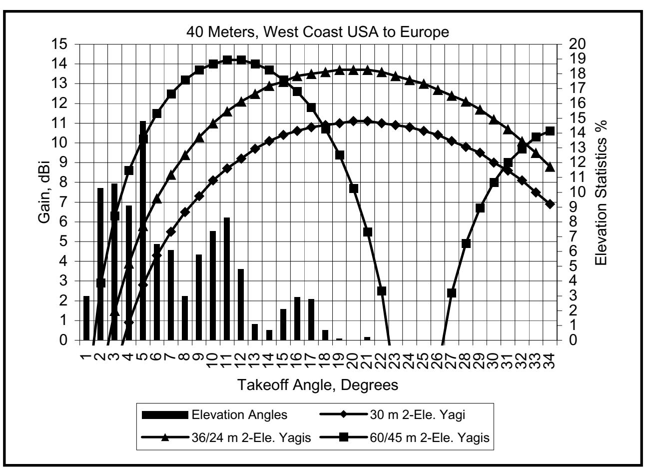


Fig 5-3 — A comparison of the elevation responses versus elevation-angle statistics for the same path as Fig 5-2, but for more ambitious antennas mounted over flat ground. At very low angles (2° to 4°), you gain approximately 8 dB going from a single 30-meter high 2-element Yagi to a very high stack of identical Yagis at 60 and 45 meters. Ambitious, indeed!

place that same vertical over saltwater and see what happens. An almost perfect match results, even better than the stack of 2-element Yagis at 45 and 60 meters! More about the magic of saltwater in Chapter 9 (verticals).

1.1.1.4. 80 Meters

Let's have a look at 80 meters. From Table 5-1 and Fig 5-1 you can see there is little difference in the overall range of elevation angles between 40 and 80 meters. Let us analyze the US East Coast to Europe path, where elevation angles extend up to approximately 35° on 80 meters.

The horizontal dipoles in **Fig 5-5** are relatively poor performers for this range of elevation angles, even for antennas at a height of 45 meters! Only a giant 2-element 80-meter Yagi at that height covers the low wave angles reasonably well down

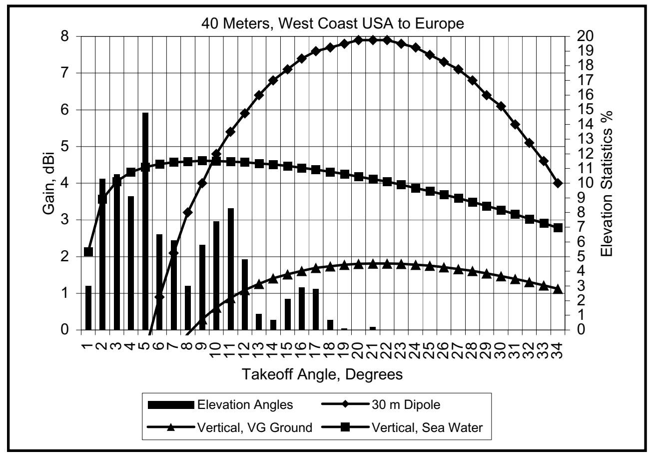


Fig 5-4 — A comparison of horizontal versus vertical antennas for the 40-meter path from Europe to the US West Coast. At very low angles (less than about 10°) a quarter-wave vertical over saltwater would have a decided advantage over a horizontal dipole that is 30 meters high over flat ground. A quarter-wave vertical mounted over "very good" ground (typical of farmland in Belgium) would be stronger than the 30-meter high dipole at elevation angles below about

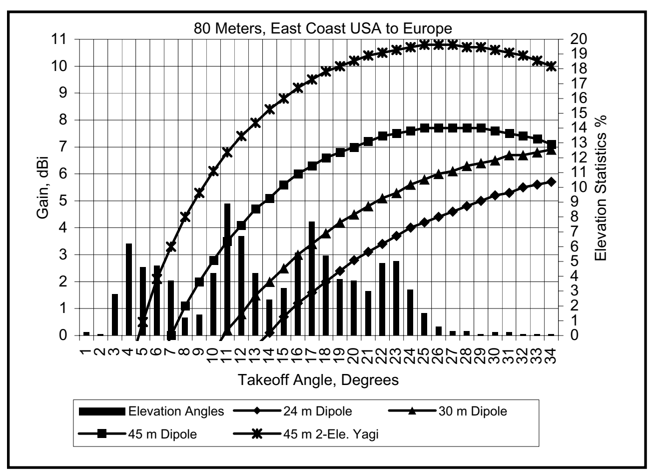


Fig 5-5 — A comparison of the elevation responses versus elevation-angle statistics for the 80-meter path from Washington, DC, on the US East Coast, to Europe. Note the response for a gigantic 2-element 80-meter Yagi at 45 meters, a truly heroic antenna!

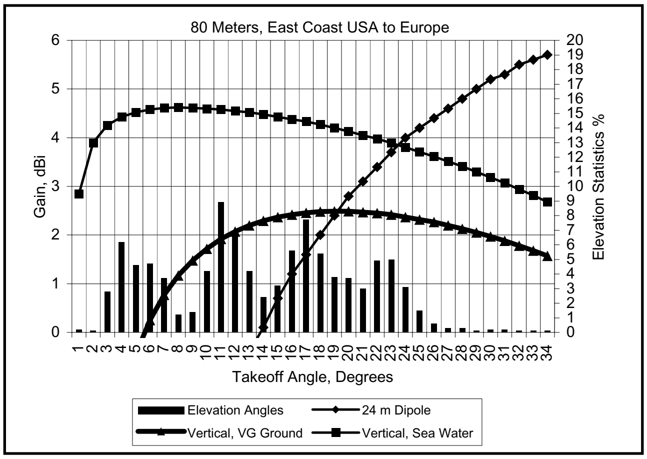


Fig 5-6 — A comparison of horizontal and vertical antennas on 80 meters from the US East Coast to Europe. A quarter-wave vertical over saltwater is virtually unbeatable for angles lower than about 20°.

to about 5°. On this band verticals perform much better than on 40 meters (**Fig 5-6**). The single vertical over very good ground is better than the horizontal dipole at 24 meters, and covers the elevation angles almost as well as the 2-element Yagi at 45 meters. The λ/4 vertical over saltwater is unbeatable!

1.1.1.5. 160 Meters

On Top Band most of us have the choice between an antenna that shoots straight up (a horizontal dipole or inverted-V dipole even at 30 meters in height will produce a 90º takeoff angle), and a vertical (it may be shortened or in the form of an inverted-L or T-antenna) that produces a good low radiation angle (20° to 40° depending on the ground quality). This means we have little chance to experience the differences in signal strength between different radiation angles. The antenna with a low wave angle would be the best in maybe 99% of the cases. Again, there are (even more than on 80 meters) exceptional cases where a high radiation angle is required to launch into a ducting mechanism, which around sunset or sunrise can produce much stronger signals than can be achieved using a low wave-angle antenna at these times.

Even if you have a very high horizontally polarized antenna (as does Tom, W8JI, with his 100-meter high inverted V), this does not mean it will perform as well as a vertical on 160 meters. The reason for that is explained in Chapter 1 (Section 3.4 and 3.5). Tom confirms that his very high dipole almost never equals his Four Square array, which uses quarter-wave verticals. The suspected mechanism only applies to 160 meters, because of the proximity of 1.8 MHz to the electron gyro frequency.

1.1.1.6. Conclusion, Elevation Angles

For 40 and 80 meters we have elevation-angle statistics generated using a mathematical model, based on long-term observed propagation data. The wave angles are averages over many sunspot cycles, throughout the different seasons of the years and throughout the night (darkness path). Looking at the California-to-Europe angle distribution on 80 meters, we see that there is 3% chance that the angle is 1° and 1% chance as well that the angle is 20°. But what will the exact wave angle be tonight? The models give us good insight on the range of what is possible. They do not tell us anything about *when* a particular angle will occur. Fortunately our real-life antennas are not radiating at just one wave angle, but rather over a range of angles. The trick is to have an antenna or antennas where the range of actual radiating angles matches the range of statistically available wave angles as closely as possible. That way you cover all the possibilities.

What we also learn from the model is that propagation angles above 35° are rarely present on 40 and 80 meters under normal geomagnetic conditions, and that there is usually some sort of mechanism that supports propagation at very low wave angles. Does this come as a surprise? No. We all have heard, time after time, that vertical antennas on the beach radiating over saltwater produce astonishingly strong signals on the low bands (Ref 183). There is also a lot of evidence about high-angle propagation near sunrise/sunset, where a high wave angle appears to be often favorable to initiate ducting (see Chapter 1). Such "anomalies," which are by definition of short duration, are not included in the statistical data on which *IONCAP* and *VOACAP* are based.

1.1.2. The Influence of Sloping Terrain

Where I live in Belgium, it's really, really flat. About 65 km from the coast, my QTH is 30 meters above sea level. It's flat as a pancake! But many low-band DXers live in hill country or even on mountaintops. It's not only saltwater locations that can do wonders — a mountaintop with the right slope and the right type of terrain pattern in the far field can also work wonders. In the mid 1980s I wrote a simple software program that could evaluate simple sloping terrains. That program is still part of the *Yagi Design* software (see Chapter 4). Years later K6STI developed *TA* (*Terrain Analysis*) and N6BV developed *YT* (*Yagi Terrain Analysis*). These programs ray-trace over complex terrain using diffraction methods. The 20th and later editions of *The ARRL Antenna Book* (Ref 697) now include a full-blown *Windows* program called *HFTA* (*High Frequency Terrain Analysis*) by N6BV. *HFTA* only models terrain for horizontally polarized antennas (for dipoles and 2 to 8-element Yagis). If you want to include a vertical over flat ground, you can model it (for example, with *EZNEC*). This is how the vertical patterns in Figs 5-4 and 5-6 were made.

Let's have a look at some 40 and 80-meter antennas on "hilly terrains." **Fig 5-7A** shows the terrain for several prominent contest and DX stations, as modeled by N6BV using *HFTA*. K1KI's QTH in Connecticut has a gentle slope that drops about 12 meters over the first 300 meters distance from the tower toward Europe. The impact of this down slope is nevertheless quite substantial and low takeoff angles are covered much better than over flat terrain, as shown in Fig 5-7B. When he was in New Hampshire, N6BV's terrain sloped down 20 meters in the first 300 meters from the tower base and this too yielded a good improvement at low angles.

The third example is the spectacular mountaintop QTH of 4O1A (located on the top of a ridge forming a peninsula), which features a fairly steep slope of almost 600 meters all the way down to the sea, some 2800 meters from his antennas (the average slope angle is approximately 12°). The low-angle fill-in looks quite spectacular, but is not the whole story, at least not on the path to the USA.

Fig 5-7C compares quarter-wave verticals with horizon-

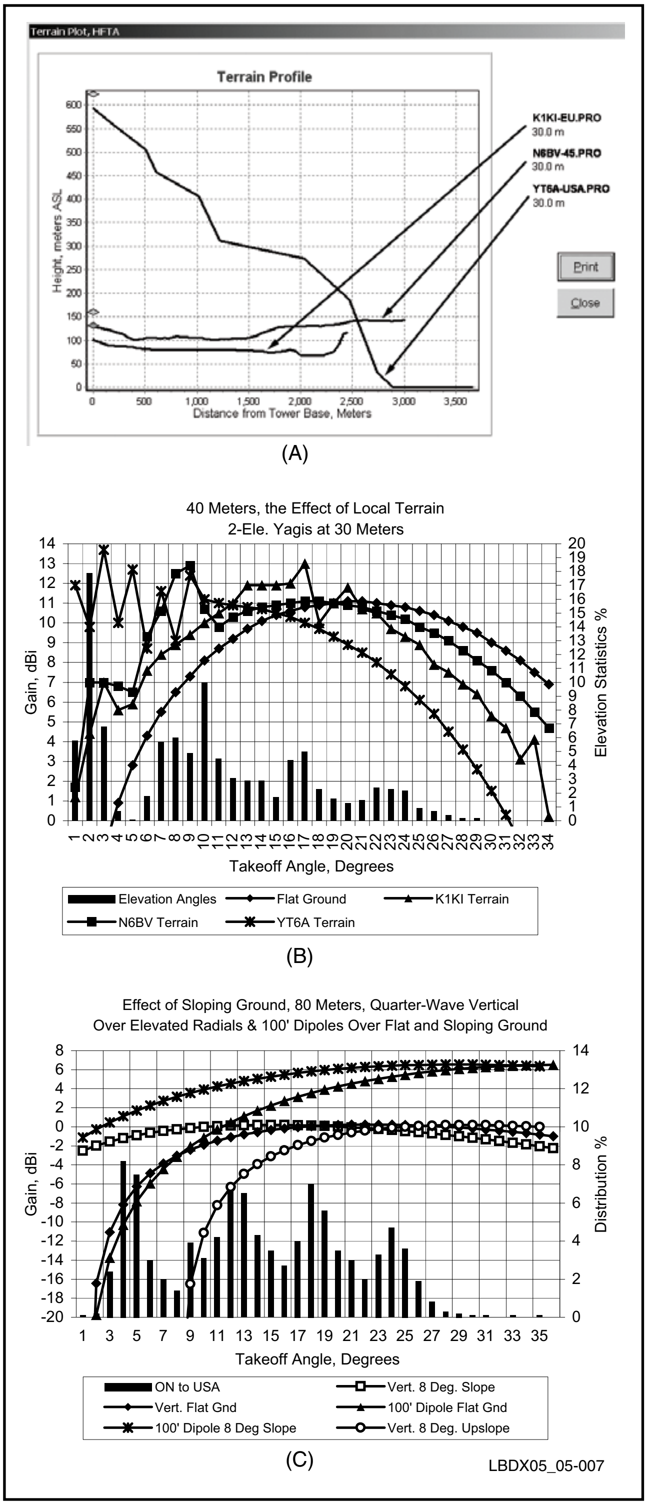


Fig 5-7 — At A, the Terrain Profiles from ARRL's *HFTA* (HF Terrain Assessment) program for several contesting stations. The drop-off from the mountaintop QTH of YT6A (now 4O1A) as modeled by N6BV may seem breathtaking but tells us only half the story, as "obstacles" at a further distance are not considered. At B, the elevation responses for each of these three QTHs for a horizontally polarized 2-element Yagi at a height of 30 meters. For reference, the response for the same antenna over flat ground is shown also. At C, elevation responses for quarter-wave verticals and horizontal dipoles over flat and sloping ground (downward and upward for a vertical antenna at 8°) on 80 meters.

tal dipoles on 80 meters. In each case the terrain is either flat ground or ground with an 8° down slope. A down slope in the direction of interest can materially aid low elevation angles for verticals as well as horizontals. One trace in Fig 5-7C is for a vertical antenna with ground sloping upward at 8°, effectively blocking really low takeoff angles.

In the 4th edition of this book, N6BV provided data for the 4O1A QTH. It appears, after my visit to this wonderful QTH, that Dean's assessment only covered half of what's going on there. The 4O1A (YT6A) terrain slope curve shown in Fig 5-7A stops where the slope is seawater. We will see that if you want to assess the situation at very low angles, you have to look much further than that.

If you do live in hilly country, you really should use terrain-modeling software to see the effects that real-world terrain has on the launch of HF signals into the ionosphere. You will have to make terrain data files for your particular QTH for all directions of interest. You can do this manually: Buy a detailed paper topographic map and note the terrain height at intervals (usually corresponding to the height contours on the topo map) along the direction of interest. If you want better accuracy use your GPS (one that displays terrain height), walk the area and annotate the map.

An easier way that should get you the same results is to visit **www.topocoding.com**. First move around on the Google Earth map (who does not know how to do that?) until you have located the area for which you want to make a profile. By clicking the map you add points to the path whose altitude profile can be displayed using the first button. See **Fig 5-8** for a profile for 4O1A.

You can also add the Altitude Resolving tool to your Google maps. For more information, visit **maps.google.com** and **www.topocoding.com**.

Caution: The profile is not everything when it comes to assessing the reflection properties of the area surrounding the antenna. The quality of the ground where the reflection takes place is also very important.

Also consider that in case of a *flat terrain*, reflection on

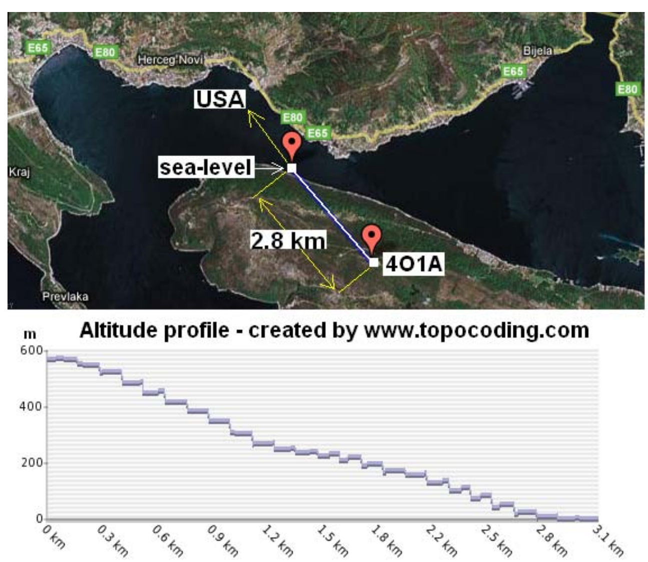


Fig 5-8 — Terrain profile toward the USA for 4O1A. Compare this profile with the profile generated with *HFTA* — they are identical. As we can see in Figs 5-9, and 5-10 this is however only half of the story.

the ground that builds up the far field pattern happens between approximately 500 meters and 3 km from the antenna (3 km distance for a 1° take off angle). In case of the antenna being at a high point above the reflection level, the reflection point for low angles can be much further away. This is what happens in the 4O1A case as shown in **Fig 5-9**. If the antennas are 600 meters above the sea, the reflection distance for a 1° wave angle is 35 km away! Thus: the higher up your antennas are, the further the far-field reflection distance for low radiation angles.

In the analysis of the 4O1A QTH case, we must take into account that our flat seawater reflection surface ends at approximately 5.8 km from the antenna (see **Fig 5-10**). A little basic trigonometry tells us that the lowest unobstructed reflection angle is ~6° [arc tan (600/5800)]. Transmitted signals arriving at a lower angle (lower than 7°) on the other side of the bay (at a distance greater than 5.8 km) will be blocked or scattered against the mountainous terrain on the north side of the bay. In essence the radiation at angles lower than ~6° will be reduced as shown for the case of a vertical antenna with a 6° upslope (Fig 5-7C). In that particular direction the gain will be substantially lower than shown in Fig 5-7A.

1.2. Receiving Antennas

For a receiving antenna, the requirements are very different on the lower bands (80 and 160 meters). We expect the antenna to receive only signals from a given direction and at a given wave angle (we want horizontal as well as vertical directivity), and that for two reasons:

1) We want to reduce QRM from signals coming from other azimuths.
2) We want to receive as little as possible of the atmospheric and man-made noise, which by definition comes in (most of the time) from all directions (see Chapter 7 on Receiving Antennas). This is the main reason why we want directivity: to eliminate noise from all other directions than the direction we are listening, and thus improve the S/N (signal to noise) ratio.

Fig 5-9 — This picture was taken from the 4O1A QTH at 600 meters ASL, looking in the direction of the USA. It is clear that there is no perfect reflecting saltwater in that direction for very low angles (below 6°, see text) as suggested by the analysis shown in Fig 5-7.

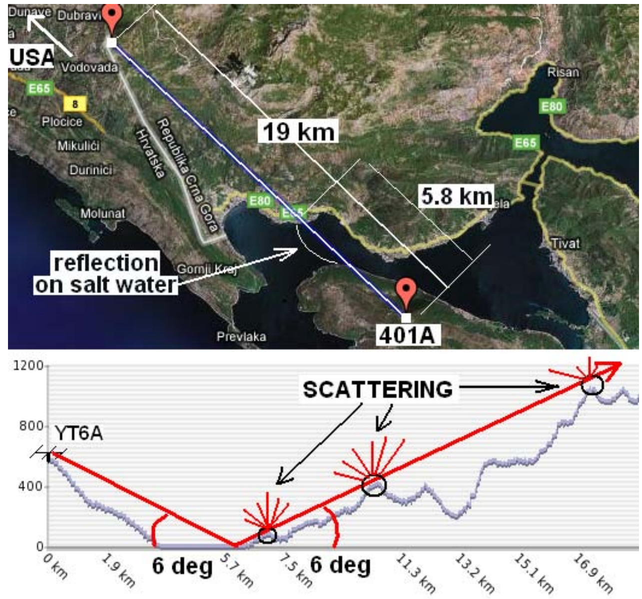


Fig 5-10 — The problem with the 4O1A (YT6A) terrain profile in Fig 5-7A is that it assumes that once the sea (bay) is reached, this saltwater surface stretches out for ever. After 5.7 km the terrain goes up again and reaches a height of over approximately 1000 meters at 19 km distance. This limits the lowest clear shot angle to approximately 7° instead of 1°.

As far as gain is concerned, all we want is the receiving antenna to produce as much signal (plus atmospheric and man-made noise) level as is necessary to override the internal noise level generated in the receiver (see Chapter 3, Section 1.2.). Having said that and knowing what the noise levels are on the low bands (see Table 3-1 in Chapter 3) and the sensitivities of our present days receivers, we can easily conclude that efficiency (see Section 2.5) of a receiving antenna is really almost never an issue.

In most amateur applications on the higher bands, the transmitting antenna is used as the receiving antenna. The transmitting requirements of the antenna outweigh typical receiving requirements. On the low bands, however, successful DXers most often use specialized receiving antennas, as we will see in Chapter 7 on Receiving Antennas. This is because most hams cannot build very directive (and efficient) transmit antennas, which are very large. It is possible, however, to build very effective directive receiving antennas that have poor efficiency, making them unsuitable for transmitting.

2. DEFINITIONS

2.1. The Isotropic Antenna

An *isotropic* antenna is a theoretical antenna of infinitely small dimensions that radiates equally well in all directions. This concept can be illustrated by a tiny light bulb placed in the center of a large sphere (see **Fig 5-11**). The lamp illuminates the interior of the sphere equally at all points. The isotropic antenna is often used as a reference antenna for gain comparison, expressed in decibels over isotropic (dBi). The radiation pattern of an isotropic antenna is a sphere, by definition. The term dBi is no more and no less than a convenient abbreviation for power per unit area over the volume of a sphere.

2.2. Antennas in Free Space

Free space is a condition where no ground or any other conductor interacts with the radiation from the antenna. In practice, such conditions are approached only at VHF and UHF, where very high antennas (in terms of wavelengths) are common. Also every real-life antenna has some degree of directivity. If it is placed in the center of a large sphere, it will illuminate certain portions better than others. In antenna terms, the antenna radiates energy better in certain directions. A half-wave dipole has maximum radiation at right angles to the wire and minimum radiation off the ends. A half-wave dipole, in free space, has a gain of 2.15 dB over isotropic (2.15 dBi).

Radiation patterns are collections of all points in a given plane, having equal field strength. **Fig 5-12** shows the radiation pattern of a dipole in free space, as seen in three dimensions and in two planes, the plane through the wire and the plane perpendicular to the wire.

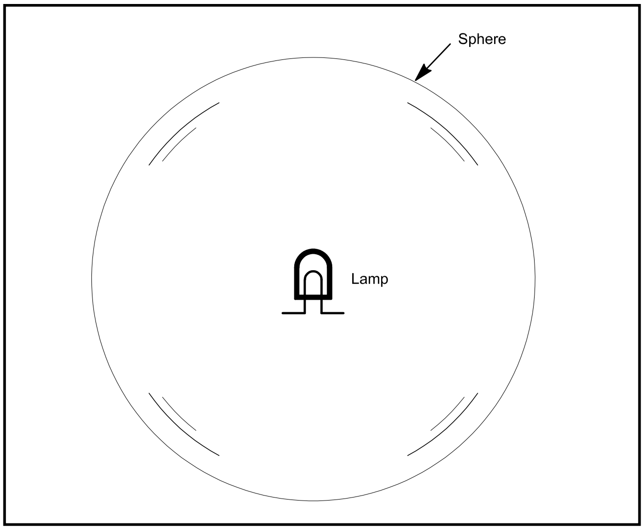


Fig 5-11 — In this drawing the isotropic antenna is simulated by a small lamp in the center of a large sphere. The lamp illuminates the sphere equally well at all points.

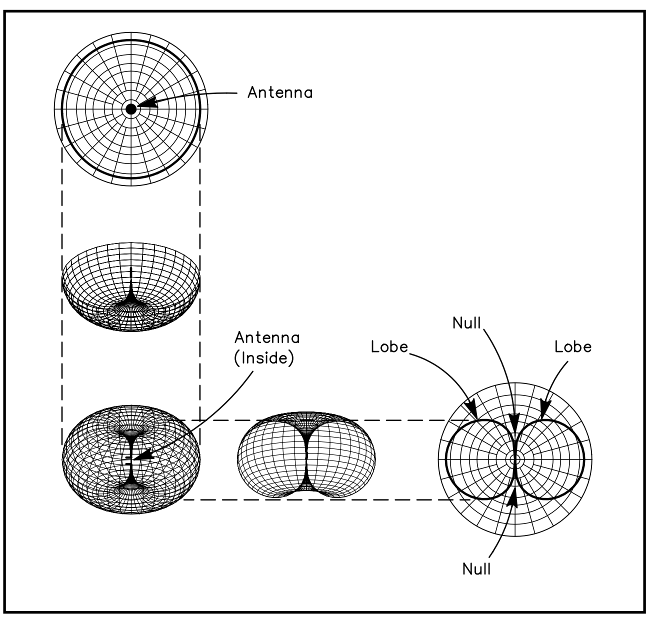


Fig 5-12 — Vertical (left) and horizontal (right) radiation patterns as developed from the three-dimensional pattern of a horizontal dipole.

2.3. Antennas over Ground

In real life, antennas are near the ground. We can best visualize this situation by cutting the sphere in Fig 5-11 in half, with a metal plate going through the center of the sphere. We are now looking at a *hemisphere* (a half sphere).

This plate represents the ground; let's assume it is a perfect electrical mirror. **Fig 5-13** shows what happens with an antenna near the ground: The antenna radiates (more or less) in all directions. Waves that are radiated "downward" (toward the reflecting "ground") are reflected on that ground. Direct waves (the waves that leave the antenna and never reach "ground") and reflected waves combine and illuminate the sphere unequally at different points at different angles. For certain angles the direct and reflected waves are in phase and reinforce one another. The field is doubled, which means a power gain of 3 dB. In addition, we have only a half sphere to illuminate with the same power, and that provides another 3 dB of gain. This means that a dipole over perfect ground will exhibit 6 dB gain over a dipole in free space.

Over ground, radiation patterns are often identified as vertical (cutting plane perpendicular to the ground) or horizontal (cutting plane parallel to the ground). The latter is of very little use, since practical antennas over real ground produce no signal at a 0° wave angle. The so-called horizontal directivity should in all practical cases be specified as directivity in a plane making a given angle with the horizon, usually at the main takeoff angle.

Low-band antennas always involve *real* ground. With real ground, the above-mentioned gain of 6 dB will be lowered, since part of the RF is dissipated in the lossy ground. For evaluation purposes, we often specify *perfect ground*, a ground consisting of an infinitely large, perfect reflector.

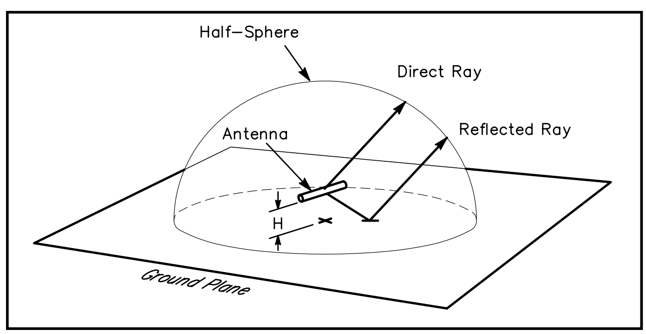


Fig 5-13 — The effect of ground is simulated in a sphere by putting a plate (the reflecting ground plane) through the center of the sphere. Since the power in the antenna is now radiated in half the sphere's volume, the total radiated field in the half sphere is doubled. The ground reflection can add up to 6 dB of signal increase compared to free space. A smaller total gain is caused in practice, since part of the RF energy is absorbed in the poorly reflecting, lossy ground.

Table 5-2
Conductivities and Dielectric Constants for Common Types of Earth

Surface Type	*Dielectric Constant*	*Conductivity (S/m)*	*Relative Quality*
Fresh water	80	0.001	
Saltwater	81	5.0	Saltwater
Pastoral, low hills, rich soil, typ Dallas, TX to Lincoln, NE areas	20	0.0303	Very Good
Pastoral, low hills, rich soil typ OH and IL	14	0.01	Good
Flat country, marshy, densely wooded, typ LA near Mississippi River	12	0.0075	
Pastoral, medium hills and forestation, typ MD, PA, NY, (exclusive of mountains and coastline)	13	0.006	
Pastoral, medium hills and forestation, heavy clay soil, typ central VA	13	0.005	Average
Rocky soil, steep hills, typ mountainous	12-14	0.002	Poor
Sandy, dry, flat, coastal	10	0.002	
Cities, industrial areas	5	0.001	Very Poor
Cities, heavy industrial areas, high buildings	3	0.001	Extremely Poor

Real grounds have varying properties, in both conductivity and dielectric constant. In this book, frequent reference will be made to different qualities of real grounds, as shown in **Table 5-2**.

2.4. Radiation Resistance

The IRE defines the radiation resistance as being equal to the power radiated as electromagnetic energy divided by the square of the net or effective current causing radiation, or

$$R_{rad} = \frac{P_{rad}}{I^2}$$

In other words: radiation resistance (referred to a certain point in an antenna system) is the resistance, which if inserted at that point, would dissipate the same energy as is actually radiated from the antenna. This definition does not state where the antenna is being fed, however. There are two common ways of specifying radiation resistance:

- The antenna is fed at the current maximum: $R_{rad\,(I)}$
- The antenna is fed at the base, between the antenna lower end and ground: $R_{rad\,(B)}$

$R_{rad\,(I)} = R_{rad\,(B)}$ for verticals of ½ wavelength or shorter. $R_{rad\,(B)}$ is the radiation resistance used in all efficiency calculations for vertical antennas. Fig 9-10 in Chapter 9 shows the radiation resistance according to both definitions for four types of vertical antennas:

- A short vertical (< 90° high)
- A quarter-wave vertical
- ⅜-wave vertical (135° high)
- ½-wave vertical

Radiation resistance is not the same as the feed-point impedance, since feed-point impedance consists of both radiation resistance and loss resistance(s), plus any reactance at the feed point.

2.5. Antenna Efficiency

The *antenna efficiency* of an antenna by itself located in free space is simply the ratio of power radiated from that antenna to the power applied to it. Any energy that is not radiated will be converted into heat in the lossy parts of the antenna. For a transmitting antenna, radiation efficiency is an important parameter. The efficiency of an antenna is expressed as follows:

$$\text{Efficiency} = \frac{R_{rad}}{R_{rad(B)} + R_{loss}} \qquad \text{(Eq 5-1)}$$

where $R_{rad\,(B)}$ is the radiation resistance of the antenna as defined in Section 2.4, and R_{loss} is the total equivalent loss resistance of all elements of the antenna (resistance losses, dielectric losses, loading coils, etc). Loss resistance is normalized to the same point where R_{rad} was defined.

The *total efficiency* of an *antenna setup* is a rather different story. While antenna efficiency only considers the lossy parts of the antenna itself, total efficiency includes losses in its environment, including the ground. In other words, total efficiency takes into account all losses in the near field as well as in the far field (see Sections 2.6, 2.7 and 2.8).

2.6. Near Field (or Radiating Near Field)

The near field is that part of the radiated field nearest to the antenna. Beyond the near field is the infinite far field. Depending on the physical dimension of the antenna the *radiating near field* (also called the *Fresnel* field) reaches out typically one or two wavelengths from simple wire antennas to many wavelengths in the case of long-boom Yagis on VHF and UHF. The relationship between magnetic and electric fields is a complex one in the near field. This is one of the reasons that we must not make antenna pattern measurements too close to the antenna. Antennas field-strength measurements should be done no less than a few wavelengths from the antenna.

With low-band antennas the ground will always be in the near field of our antennas, and losses in the near field will have to be considered. These losses will be discussed in detail in Chapter 9.

The *reactive near field* (also called *induction field* or non-radiative field) is a part of the near field, very close to the antenna where mutual coupling exists between conductors. This happens typically within a maximum of 0.5 wavelengths around the antenna.

2.7. Far Field

The *far field* (also called *radiating field* or *Fraunhöfer* field) is the area around the antenna beyond the near field. This is where ground reflections for low-angle signals occur, which greatly interest us low-band operators. In the far field the power density is inversely proportional to the square of the distance from the antenna. Total energy is equally divided between electric and magnetic fields, and the relation is defined by $E/H = Z_0 = 377\ \Omega$, the free-space impedance. See Chapter 9 for further discussion of far-field reflection losses.

2.8. Antenna Gain

The *gain* of an antenna is a measure of its ability to concentrate radiated energy in a desired direction (minus any losses in the antenna). Antenna gain is expressed in decibels, abbreviated dB. It tells us how much the antenna in question is better than a reference antenna, under defined circumstances. And that's where we enter the antenna-gain "jungle." Commonly, both the theoretical isotropic, as well as a real-world dipole, are used as reference antennas. In the former case the gain is expressed as dBi and in the latter as dBd.

But that's only part of the story. We can do a comparison in free space, or over perfect ground or over real ground. The only situation that makes a generic comparison possible is to compare antennas in free space. Gain in dBi in free space is what can always be compared; there is no inflation of gain figures by reflection. Very often manufacturers of commercial antennas will calculate gains including ground reflections — and often they will not mention this fact.

You might argue, "Why not use a real antenna, such as a dipole, as a reference, since the isotropic antenna is a theoretical antenna that does not exist, while a half-wave dipole does?" Comparing gains is really comparing the field strength of an antenna under investigation with that of our reference antenna. With an isotropic antenna the situation is clear. It radiates equally well in all directions and the three-dimensional radiation pattern is a sphere. What about the dipole as a reference? The gain of a half-wave, lossless half-wave dipole in free space over an isotropic is 2.15 dBi. But that does not mean that a real dipole has a gain of 2.15 dBi. It only means that the gain of a lossless dipole in free space (that's a theoretical condition as well, because nothing is really in free space) is 2.15 dB over an isotropic radiator. If we put the dipole over a perfect ground, it suddenly shows a gain of 8.15 dBi! You pick up 6 dB by radiating the power in half a hemisphere instead of a whole hemisphere, as in the theoretical case of free space. With less-than-perfect ground, part of the power will be absorbed in the ground and the ground-reflection gain will be less than 6 dB. It is clear that the only generic way of comparing antenna gains is in dBi, using an isotropic antenna as the only generic reference antenna not influenced by height or ground conditions. In this publication we will always quote gain figures in dBi — that is, referenced to an isotropic antenna in free space. (Ref 688).

2.9. Front-to-Back Ratio

Being a ratio (just like gain), we would expect front-to-back ratio to be expressed in decibels, which it is. The front-to-back ratio (F/B) is a measure expressing an antenna's ability to radiate a minimum of energy in the direction directly in the back of the antenna.

Free-space front-to-back ratio is always specified at a 0° wave angle, just like it is the case with the free space directivity pattern.

As over real ground there is no radiation at zero wave angle (see Section 2.3.), and as we usually specify the azimuthal radiation pattern at the main lobe wave angle, it makes sense that we specify the F/B at this same wave angle.

With the advent and the widespread use of modeling programs, especially some of the optimizer programs, the rat race started for the most ludicrous F/B figure. Let's not forget that mathematics is one thing, while antenna physics is another thing. It is possible to calculate an antenna exhibiting an F/B of 70 dB in a given direction, at a given wave angle. But that's all there is to it. One degree away the rejection may be down to maybe 25 dB. When you understand the physics behind all of this, it will be clear that F/B above a certain level (maybe 35 dB) is rather meaningless.

2.9.1 Geometric Front-to-Back Ratio

The geometric front-to-back is the ratio between the power in the forward lobe at the peak of the main wave angle to the power radiated 180° behind the front (0° lobe), at the same wave angle.

It quantifies a single point situation, and therefore is not a good way of assessing an antenna's capabilities of reducing noise and unwanted signals at the back of the antenna (see Section 1.2).

In general it is the average pattern of an antenna that determines how well an antenna discriminates against unwanted signals and noise coming from directions other than the front of the antenna. It is very unlikely that unwanted signals will be generated exactly 180° off the beam direction or at a radiation angle that is the same as the main forward lobe's radiation angle. Therefore, geometric F/B can be ruled out immediately as a meaningful way of defining the antenna's ability to discriminate against unwanted signals and noise.

2.9.2 Average Front-to-Back (Integrated Front-to-Back) Ratio

The *average front-to-back ratio* can be defined as the average value of the front-to-back as measured (or computed) over a given back angle (both in the horizontal as well as the vertical plane). In the chapter on receiving antennas (Chapter 7, Section 1.8 and 1.9) I use this concept for evaluating different antennas. This integrated F/B ratio is the only meaningful one when evaluating antennas for the low bands.

2.9.3. Worst-Case Front-to-Rear Ratio (F/R)

The *front to rear ratio* expresses the ratio of the forward power to the power in the "worst" lobe in the entire back of the antenna (the rear, from 90° to 270° azimuth and from 0° to 90° elevation). This one may be meaningful for weighing Yagis and quads on the higher bands, but is useless for the low bands, as it does not say how much better everything in the back is than in that worst spot!

2.9.4. Front-to-Back Ratio and Gain

Is there a link between gain and the front-to-back ratio of an antenna? Let's visualize a three-dimensional radiation pattern of a simple Yagi. The front lobe resembles a long stretched pear, while the back lobe (let's assume for the time we have a single back lobe) is a much smaller pear. The antenna sits where the stems of the two pears touch. The volume of the two pears (the total volume of the three-dimensional radiation pattern) is determined only by the power fed to the antenna. If you increase the power, the volume of the large as well as the small pear will increase in the same proportion. Let's look at front-to-back ratio, the ratio of the power radiated in the back versus the power radiated in the front. This means that the F/B ratio is proportional to the ratio of the volume of the two pears.

By changing the design of the Yagi (by changing element lengths or element positions), we change the size and the shape of the two pears. But so long as we feed the same power to it, the sum of the volumes of the two pears remains unchanged. It's as if the two pear-shaped bodies are connected with a tube, and are filled with a liquid. By changing the design of the antenna, we merely push liquid from one pear into the other. If the antenna were isotropic, the radiation body would be a sphere having the volume of the sum of the two pears.

Assume we have 100 W of power with 10% of this power applied to the antenna in the back lobe. The F/B will be 10 × log (10 / 1) = 10 dB. Ninety percent of the applied power is available to produce the forward lobe.

Let's take a second case, where only 0.1% of the applied power is in the back lobe. The F/B ratio will be 10 × log (100 / 0.1) = 30 dB. Now we have 99.9% of the power available in the front lobe.

The antenna gain realized by having 99.9 W instead of 90 W in the forward lobe is 10 × log (99.9 / 90) = 0.45 dB. Pruning an antenna with a modest F/B pattern (10 dB) to an exceptional 30-dB value, gives us 0.45 dB more forward gain, provided that the extra liquid is used to lengthen the cone of the big pear.

The mechanism for obtaining gain and F/B is much more complicated than that described above. I am only trying to explain that optimizing an antenna for F/B does not necessarily mean that it will be optimized for gain. What is always true, however, is that a high-gain antenna will have a narrow forward lobe. You cannot concentrate energy in one direction without taking it away from other directions! We will see later that maximum-gain Yagis show a narrow forward lobe, but often a poor front-to-back ratio. This is the case with very high-Q, gain-optimized 3-element Yagis, for example.

Conclusion: There is no simple relationship between front-to-back ratio and gain of an antenna.

2.9.5. The Importance of Directivity

Directivity can be important for two very different reasons: With *transmit antennas* we want to have directivity because directivity is invariably linked to gain. What you take away in certain directions is added in other directions. We want gain because we want to be heard (to be strong), and that also implies the notion of efficiency.

With *receiving antennas* the story is different. We want to hear well above the noise (manmade, atmospherics, QRM, etc). The issue is one of signal-to-noise ratio, not signal strength. While antenna efficiency is a secondary issue with receiving antennas for the low bands, directivity is primary. That's why the concept of quantifying the directivity of an antenna was developed (see Chapter 7, Section 1.2.).

2.10. Directivity Merit Figure and Directivity Factor

For a receiving antenna on the low bands directivity is the main concern. There are currently two methods to quantify this directivity:

2.10.1 Directivity Merit Figure

The average front-to-back (the peak forward lobe versus what happens in the back 180° over the entire elevation angle range) gives a good indication of directivity. This method was for the first time described in the third edition of this book to quantify some of the special receiving antennas in Chapter 7. The DMF (directivity merit figure) is the difference between the forward gain (at the desired wave angle, such as 20°) and the average gain over the entire back (the back quadrisphere) of the antenna (see Chapter 7, Section 1.10).

2.10.2. Receiving Directivity Factor

Tom, W8JI, (**www.w8ji.com**) goes a step further and compares the forward-lobe gain to the average gain of the antenna in all directions (both azimuth and elevation). This figure tells you not only how good the average front-to-back ratio is, but also how narrow your forward (wanted) lobe is. This merit figure is called *RDF* (receiving directivity factor).

Both the RDF and the DMF are ratios, which means that they are expressed in dB. They are extensively used in Chapter 7 to measure the performance of receiving antennas for the low bands.

2.11. Standing-Wave Ratio

SWR is *not* a performance measure of an antenna! SWR is only a measure of how well the feed-point impedance of the antenna is matched to the characteristic impedance of the feed line. The characteristic impedance of a cable does not change as a function of the load values; it depends only on the dimensions of the cable and the characteristics of the insulating material.

If a 50 Ω feed line is terminated in a 50 Ω load, then the impedance at any point on any length of the cable is 50 Ω. If the same feed line is terminated in an impedance different from 50 Ω, the voltage and the current will vary along the line, hence also the impedance in those points along the line. The SWR is defined as the V_{max} / V_{min}. The SWR is a measure of the match between the line and the load. Changing the length of a feed line does not change the SWR on the line (apart from minute changes due to feed-line loss with longer lengths of line). What changes is the impedance at the input end of the line.

If changing the line length slightly changes the SWR reading on your SWR meter, then your SWR meter is not measuring correctly (many SWR meters fall into this category) or else you have stray common-mode current flowing on the shield of your feed line. A good test for an SWR meter is to insert short cable lengths between the end of the antenna feed line and the SWR meter (a few feet at a time). If the SWR reading changes significantly, don't expect correct SWR values.

If there are stray currents on the outside of the coaxial cable shield, a change in position on the line can indeed change the SWR reading (see Chapter 6). That's why we use a *balun* (*bal*anced-to-*un*balanced transformer) when feeding balanced feed points with a coaxial cable. In fact, current baluns (choke baluns) are a good idea to install on any coaxial feed line. You can insert a current balun (eg, a short length of coax equipped with a stack of 50 to 100 ferrite cores) at the SWR meter. If this balun changes the SWR value, RF currents are flowing on the outside of the coaxial cable.

Changing the feed-line length doesn't change the performance of the antenna. A feed line is an element that is *not* supposed to radiate. SWR on a feed line has no relation whatsoever to the radiation characteristics of an antenna. A perfect match between the line and the antenna results in a 1:1 SWR.

What then are the reasons we like a 1:1 SWR or the lowest possible SWR value?

- Showing convenient 50 Ω impedance: Unless we want to use a transmission line as an impedance transformer, we would like all feed lines to show a 1:1 SWR. This would present the design load impedance of 50 Ω for solid-state transceivers.
- Minimizing losses: All feed lines have inherent losses. This loss is minimal when the feed line is operated as a flat line (SWR = 1:1) and increases when the SWR rises. On the low bands this will seldom be important, because the nominal losses on the low frequencies are quite negligible, unless very long lengths are used.

For most hams, SWR is the only property they can more or less accurately measure. Measuring gain and F/B with any degree of accuracy is beyond the capability of most. That is why most hams pay attention only to SWR properties. The amount of SWR that can be tolerated on a line depends on:

- Additional loss caused by SWR, and this is only determined by the quality of the feed line. A high quality feed line can tolerate more SWR from an additional-loss point of view than a mediocre quality line. Bigger coax means better coax: lower initial losses and mechanically stronger.
- How much SWR the transceiver or linear amplifier can live with.
- How much power we will run into a line of given physical dimensions (for a given power, a larger coax will withstand a higher SWR without damage than a smaller one).

It must be said that a poor-quality line (a small-diameter cable with high intrinsic losses), when terminated in a load different from its characteristic impedance, will show at its input end a lower SWR value than if a good (low-loss, large-diameter) cable is used. Remember that a very long, poor (having high losses) coaxial cable (whether terminated, open or shorted at the end) will exhibit a 1:1 SWR at the input (a perfect dummy load) because of those losses.

From a practical point of view an SWR limit of 2:1 is usually sought after. From a loss point of view, it is clear that higher values can easily be tolerated on the low bands. Coaxial feed lines used in the feed systems of multi-element low-band arrays sometimes work with an SWR of 10:1!

You can always use an antenna tuner if the SWR is higher than the transceiver or the amplifier will tolerate (usually less than 2:1). Remember that the antenna tuner will not change the SWR on the line itself; it will merely transform the impedance existing at the line input and present the transceiver or linear amplifier with a more compatible SWR value. While this approach is valid on the low bands, it is not recommended for frequencies above 7 MHz, since the additional line losses caused by the SWR can become quite significant.

2.12. Bandwidth

The *bandwidth* of an antenna is the difference between the highest and the lowest frequency on which a given property exceeds or meets a given performance mark. Many amateurs only think of SWR bandwidth when the term bandwidth is used. In actual practice, the bandwidth can refer to other properties at least as important, if not more important (gain, F/B etc). Consider a dummy load, which has an excellent SWR bandwidth, but a very poor gain figure, since it does not radiate at all!

In this book, "bandwidth" (referring to an antenna) is SWR bandwidth, unless otherwise specified. In most cases the SWR bandwidth is determined by the 2:1 SWR points on the SWR curve. In this text the SWR limits will be specified when dealing with antenna bandwidths.

SWR bandwidth is an important performance criterion on the low bands. The relative frequency spread (percentage wise) of 80 meters and 160 meters is quite large as compared to the higher HF bands. For example, the center of the 80 meter band is 3.750 MHz, and the total band is almost ±6.6% of 3.750 MHz. Special attention must be given to all bandwidth aspects, not only SWR bandwidth.

2.13. Q-Factor

2.13.1. The Tuned Circuit Equivalent

An antenna can be compared to a tuned LCR circuit. The *Q factor* of an antenna is a measure of the SWR bandwidth of an antenna. The Q factor is directly proportional to the difference in reactance on two frequencies around the frequency of analysis, and inversely proportional to the radiation resistance and relative frequency change.

$$Q = \frac{F_0 \times (X1 - X2)}{2 \times R \times \Delta F} \qquad \text{(Eq 5-2)}$$

where

X1 = reactance at the lower frequency

X2 = reactance at the higher frequency

R = average value of resistive part of feed-point impedance at frequencies of analysis ($R_{rad} + R_{losses}$)

ΔF = relative frequency change between the higher and the lower frequency of analysis

Example:

F_{low} = 3.5 MHz

F_{high} = 3.6 MHz

F_0 = 3.55

ΔF = 3.6 – 3.5 = 0.1

$R_{feed\ (Avg)}$ = 50 Ω

X1 = –20 Ω

X2 = +20 Ω

It is clear that a low Q can be obtained through:

- A high value of radiation resistance.
- High loss resistance.
- A flat reactance curve.

An antenna with a low Q will have a large SWR bandwidth, and an antenna with a high Q will have a narrow SWR bandwidth. Antenna Q factors are used mainly to compare the (SWR) bandwidth characteristics of antennas.

2.13.2. The Transmission-Line Equivalent

A single-conductor antenna (vertical or dipole) with sinusoidal current distribution can be considered as a single-wire transmission line for which a number of calculations can be done, just as for a transmission line.

2.13.2.1 Surge Impedance

The characteristic impedance of the antenna seen as a transmission line is called the *surge impedance* of the antenna.

The surge impedance of a vertical is given by:

$$Z_{surge} = 60 \times \ln \left[\frac{4h}{d} - 1 \right] \qquad \text{(Eq 5-3)}$$

where

h = antenna height (length of equivalent transmission line)

d = antenna diameter (same units).

The surge impedance of a dipole is:

$$Z_{surge} = 276 \times \log \left[\frac{S}{d \times \sqrt{1 + \frac{S}{4h}}} \right] \qquad \text{(Eq 5-4)}$$

where

S = length of antenna

d = diameter of antenna

h = height of antenna above ground.

2.13.2.2 Q-factor

The Q-factor of the transmission-line equivalent of the antenna is given by:

$$Q = \frac{Z_{surge}}{R_{rad} + R_{loss}} \qquad \text{(Eq 5-5)}$$

Example 1:

A 20-meter (66-foot) vertical with OD = 5 cm (1.6 inches), and $R_{rad} + R_{loss} = 45\ \Omega$.

$$Z_{surge} = 60 \times \ln \left[\frac{4 \times 2000}{5} - 1 \right] = 443\ \Omega$$

$$Q = \frac{443}{45} = 9.8$$

Example 2:

A 40-meter (131-foot) long dipole, at 20 meters (66 feet) height is made of 2 mm OD wire (AWG 12). The feed-point impedance is 75 Ω.

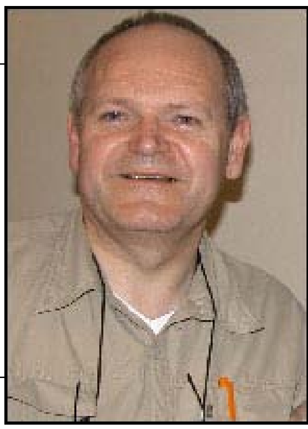
$$Z_{surge} = 276 \times \log \left[\frac{4000}{0.2 \times \sqrt{1 + \frac{4000}{4 \times 2000}}} \right] = 1163\ \Omega$$

$$Q = \frac{1163}{75} = 16$$

Contents

CHAPTER 6

The Feed Line and the Antenna



Probably my closest friend (both in and outside of ham radio circles) is Roger Vermet, ON6WU. We met some 25 years ago, and since then Roger has been a major part in my Amateur Radio life and in all my antenna projects. Roger was also very helpful when I wrote the Low Band and Yagi Design software, and entire sections were written by him.

Roger has the best from two different worlds: his theoretical knowledge in antenna matters is outstanding, his mathematics are perfect and he has hands-on experience as well. Here too the proof of the pudding is in the eating. Once a project is worked out on paper and in his lab, he wants to build it and test it.

Roger's professional career was with a major CATV cable company and later with the largest Belgian electric power company, where he retired a few years ago. The nice thing about that is that we now have even more time for common projects.

The feed line is the necessary link between the antenna and the transmitter/receiver. It may seem odd that I cover feed lines and antenna matching before discussing any type of antenna. I want to make it clear that antenna matching and feeding has no influence on the characteristics or the performance of the antenna itself (unless the matching system and/or feed lines also radiate, which is normally not what we want). Antenna matching is generic, which means that any matching system can, in theory, be used with any antenna. Antenna matching must therefore be treated as a separate subject.

The following topics are covered:

- Coaxial lines; open-wire lines
- Loss mechanisms
- Real need for low SWR
- Quarter-wave transformers
- L networks
- Stub matching
- Wide-band transformers
- 75-Ω feed lines in 50-Ω systems
- Baluns
- Connectors

Before we discuss antennas from a theoretical point of view and describe practical antenna installations, let us analyze what matching the antenna to the feed line really means and how we can do it.

1. PURPOSE OF THE FEED LINE

The feed line transports RF from a source to a load. The most common example is from a transmitter to an antenna. When terminated in a resistor having the same value as its own characteristic impedance, a transmission line operates under ideal circumstances. The line will be *flat* — meaning that there are no standing waves on the line. The value of the impedance will be the same in each point of the line. If the feed line were lossless, the magnitude of the voltage and the current would also be the same along the line. The only thing that would change is the phase angle and that would be directly proportional to the line length. All practical feed lines have losses, however, and the values of current and voltage decrease along the line.

In the real world the feed line will rarely if ever be terminated in a load giving a 1:1 SWR. Since the line is most frequently terminated in a load with a complex impedance, in addition to acting as a transport vehicle for RF, the feed line also acts as a transformer. The impedance (also the voltage and current) will be different at each point along a mismatched line.

Besides transporting energy from the source to the load, feed lines are also used to feed the elements of an antenna array, whereby the characteristics of the feed lines (with SWR) are used to supply current at each element with the required relative magnitude and phase angle. This application is covered in detail in Chapter 11 (Vertical Arrays) and Chapter 7 (Receiving Antennas).

2. FEED LINES WITH SWR

A feed line that is not terminated in a load having exactly the same impedance as the characteristic impedance of the feed line has standing waves, and is said to be a feed line with SWR. The typical characteristics of a line with SWR are:

- The characteristic impedance of the line remains unaltered and is the same in every point of the line.
- The impedance in every point of the line is different, which means that the line acts as an impedance transformer.
- While the impedances in a lossless line repeat themselves every half wavelength, the impedances in a real-world lossy line do not repeat exactly. The longer the line and the higher the nominal losses, the more the impedances along the line will converge toward the value of characteristic impedance of the line.
- The voltage and the current at every point on the feed line are different.
- The losses of the line are higher than for a flat line.
- The phase shift in current and voltage is not linearly proportional to the line length. Line length in degrees does not equal phase shift in degrees, except in very special cases such as for 90° long lines.

Most transmitters, amplifiers and transceivers are designed to work into a nominal impedance of 50 Ω. Although they will provide a *reasonable* match to a range of impedances that are not too far from the 50 Ω value (eg, within the 2:1 SWR circle on the Smith Chart), it is generally a proof of good engineering and workmanship that an antenna on its design frequency, shows a 1:1 SWR on the feed line. This means that the feed-point impedance of the antenna must be *matched* to the characteristic impedance of the line at the design frequency. The SWR bandwidth of the antenna will be determined by the Q factor of the antenna, but the bandwidth will be largest if the antenna has been matched to the feed line (1:1 SWR) at a design frequency within that passband, unless special broadband matching techniques are employed. This means we want a low SWR for reasons of convenience: We don't want to be forced to use an antenna tuner between the transmitter and the feed line in order to obtain a match.

2.1. Conjugate Match

A conjugate match is a situation where all the available power is coupled from the transmitter into the line. In a conjugate match with lossless line, the impedance seen looking toward the load (a + *j*b) at a point in the transmission line is the complex conjugate of that seen looking toward the source (a – *j*b). A conjugate match is automatically achieved when we adjust the transmitter for maximum power transfer into the line. In transmitters or amplifiers using vacuum tubes, this is done by properly adjusting the common pi or pi-L network. Modern transceivers with fixed-impedance solid-state amplifiers do not have this flexibility, and an external antenna tuner will be required in most cases if the SWR is higher than 1.5:1 or 2:1. Many present-day transceivers have built-in antenna tuners that automatically take care of this situation.

But this is not the main reason for low SWR. The above reason is one of "convenience." The real reason is one of losses or attenuation. A feed line is usually made of two conductors with an insulating material (dielectric) in between. Open-wire feeders and coaxial feed lines are the two most commonly used types of feed lines.

2.2. Coaxial Cable

Coaxial feed lines are by far the most popular type of feed lines in amateur use, for one very specific reason. Due to their coaxial (unbalanced) structure, all magnetic fields caused by RF current in the feed line are kept inside the coaxial structure. This means that a coaxial feed line is totally *inert* from the outside, when terminated in an unbalanced load. The fact that whether or not the load impedance is a perfect match to the feed line characteristic impedance is irrelevant. In other words, feed lines with SWR do not radiate if terminated in an unbalanced load, whatever its impedance. An unbalanced load is a load where one of the terminals is grounded.

This means you can bury the coax, affix it to the wall or under the carpet, tape it to a steel post or to the tower without in any way upsetting the electrical properties of the feed line. Sharp bending of coax should be avoided, however, to prevent impedance irregularities and permanent displacement of the center conductor caused by cable dielectric heating and induced stresses. A minimum bending radius of five times the cable outside diameter is a good rule of thumb for coaxial cables with a braided shield.

Like anything exposed to the elements, coaxial cables deteriorate with age. Under the influence of heat and ultraviolet light, some of the components of the outer sheath of the coaxial cable can decompose and migrate down through the copper braid into the dielectric material, causing degradation of the cable. Ordinary PVC jackets used on older coaxial cables (RG-8, RG-11) showed migration of the plasticizer into the polyethylene dielectric. Newer types of cable (RG-8A, RG-11A, RG-213 and so on) use non-contaminating sheaths that greatly extend the life of the cable. CATV and hardline cable normally use high density polyethylene outer jackets.

Also, coaxial cables love to drink water! Make sure the end connections and the connectors are well sealed. Because of the structure of the braided shield, the interstices between the inner conductor insulation and the outer sheath will literally suck up liters of water, even if only a pin hole is present. Once water has penetrated cable with a woven copper shield, it is ruined. Here is one of the big advantages of the larger coaxial cables using expanded polyethylene and a corrugated solid copper outer conductor: Since the polyethylene sticks (bonds) to the copper, water penetration is impossible even if the outer jacket is damaged. Also, migration of contaminants though the conductor shield into the inside of the coax is impossible with solid (welded) corrugated copper shields.

You should check the attenuation of your feed lines at regular intervals. You can easily do this by opening the feed line at the far end. Then feed some power into the line through an accurate SWR meter and measure the SWR at the input end of the line. A lossless line will show infinite SWR (Ref 1321).

The loss in the cable at the frequency you do the measurement is given by:

$$\text{Loss (dB)} = 10 \log \left[\frac{\text{SWR} + 1}{\text{SWR} - 1} \right] \qquad \text{(Eq 6-1)}$$

where SWR stands for the measured value with the line open or shorted at the far end.

The attenuation can also be computed using the graph in

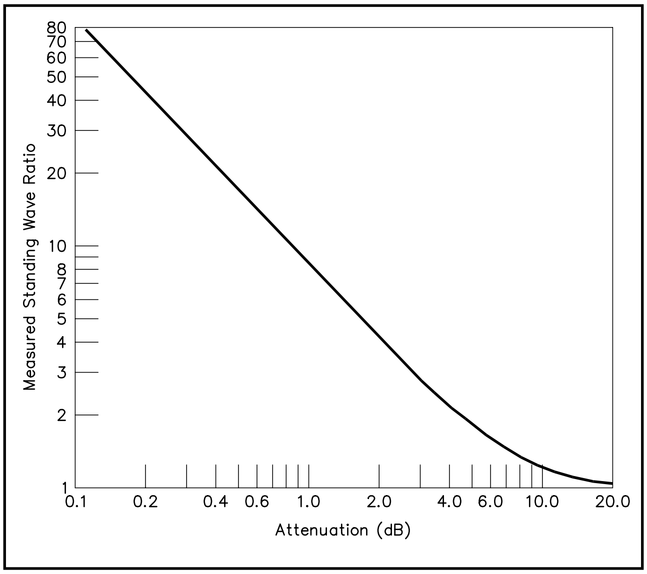


Fig 6-1 — Cable loss as a function of SWR measured at the input end of an open or short-circuited feed line. For best accuracy, the SWR should be in the 1:1 to 4:1 range.

Fig 6-1. It is difficult to do this test at low frequencies because the low attenuation is such that accurate measurements are difficult. For best measurement accuracy the loss of the cable to be measured should be on the order of 2 to 4 dB (SWR between 2:1 and 4:1). The test frequency can be chosen accordingly. Use a professional type SWR meter such as a Bird wattmeter, or a more modern one like the N8LP LP-100 (**www.telepostinc.com**) or the Array Solutions PowerMaster (**www.arraysolutions.com**). Many low cost SWR meters are inadequate.

This measurement can also be done using one of the popular antenna analyzers (eg MFJ, Autek, AEA, RigExpert). High performance network analyzers such as the AIM 4170 or the VNA 2180 vector network analyzer (see Chapter 11, Section 3.5.2.4), developed by W5BIG (**w5big.com**) and available from Array Solutions (**www.arraysolutions.com**), are ideally suited for such measurements.

It is recommended that you do an attenuation check on all your feed lines on a regular basis. Once a year is a good idea. Keep a record of the results (both the cable loss and the test frequency) and compare results of new measurements with the previous results. If you see any sudden changes, you might as well investigate that cable. A network analyzer will indicate the return loss (in dB) in addition to SWR. If you connect an open ended or shorted cable, the loss in the cable is always half of the indicated return loss. If you use the AIM 4170 analyzer you can directly read the losses of the cable on your PC screen (see **Fig 6-2**). To obtain reasonable measurement accuracy, make sure that the SWR measured at the test frequency is at least 15:1 (equivalent to a return loss of 1.2 dB or a line loss of 0.6 dB).

If it is difficult to decouple the feed line at the far end, you can still do a reasonably accurate measurement as described by K6LL: "Plug your antenna into the feed line in the shack and tune it to a frequency where it shows a peak SWR.

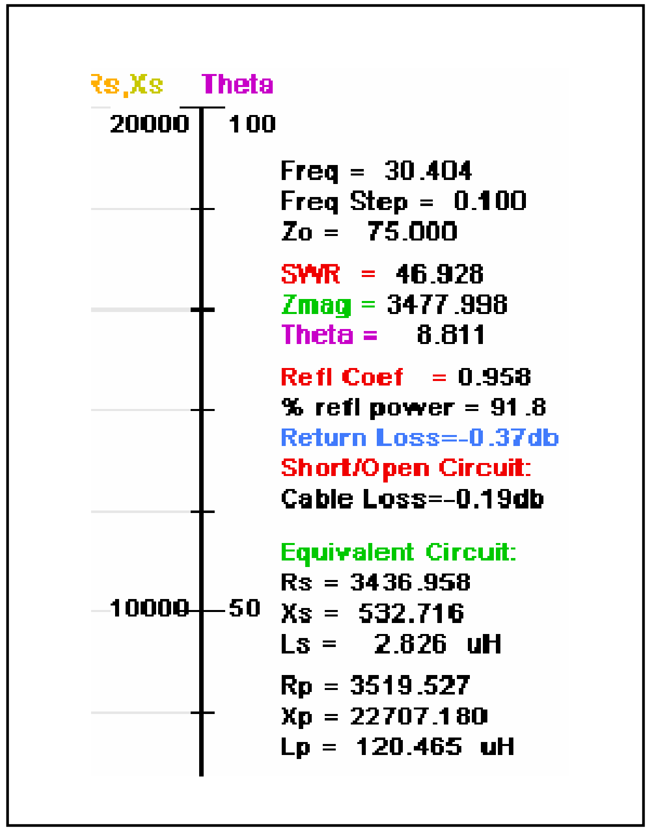


Fig 6-2 — The AIM4170 network analyzer displays on the screen the return loss (in this example 0.37 dB), as well as the cable loss (0.19 dB) for the case that the far end of the cable is open or shorted.

At this frequency, the antenna, whatever it is, will be a good approximation of an open or short circuit. The frequency will probably *not* be in the ham bands. Start at 30 MHz and work down." The same formula as above (Eq 1) and the graph in Fig 6-1 apply to this technique too. According to K6LL the impedance of most nonresonant antennas is several thousand ohms (SWR >40). If the presence of an antenna does degrade the measurement at all, it will be in a direction to make the feed line loss appear higher than it really is. If you have any concerns about whether this method is making your feed line appear too lossy, you will have to disconnect the antenna. At that time you can do the measurement at any frequency, as long as you see at least a few dB of return loss

2.3. Open-Wire Transmission Line

Even when properly terminated in a balanced load, an open-wire feeder will exhibit a strong RF field in the immediate vicinity of the feed line (try a neon bulb close to an open-wire feeder with RF on it!). This means you cannot "fool around" with open-wire feeders as you can with coax. During installation all necessary precautions should be taken to preserve the balance of the line: The line should be kept away from conductive materials. In one word, generally it's a nuisance to work with open-wire feeders!

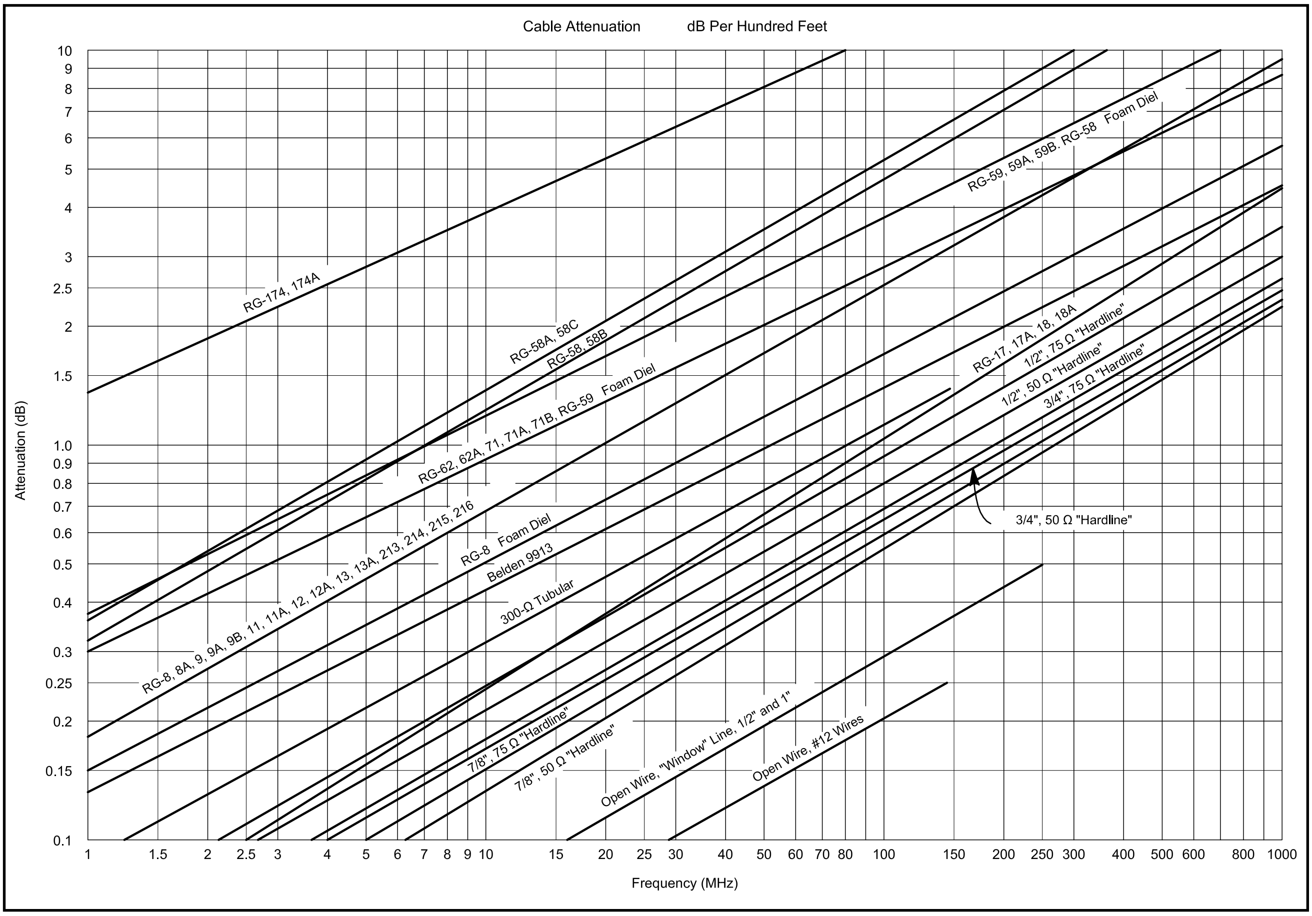


Fig 6-3 — Nominal attenuation characteristics in dB per 100 feet (30.48 meters) for commonly used transmission lines. (Courtesy of *The ARRL Antenna Book*.) If you need to know precise attenuation, use *TLW*, the transmission line program by N6BV from *The ARRL Antenna Book* (see Section 2.5).

Apart from this mechanical problem, open-wire feeders outperform coaxial feed lines in all respects on HF (VHF/UHF can be another matter).

2.4. The Loss Mechanisms

The intrinsic losses of a feed line (coaxial or open-wire) are caused by two mechanisms:

- Conductor losses (conductivity losses in the copper conductors — core and shield).
- Dielectric losses (losses in the dielectric material).

Dry air is an excellent insulator. From that point of view, an open-wire line is unbeatable. Coaxial feed lines generally use polyethylene (PE) as a dielectric, or polyethylene mixed with air or nitrogen (cellular PE or foam PE). Cables with foam or cellular PE have lower losses than cables with solid PE. They have the disadvantage of potentially having less mechanical (impact and pressure) resistance. When the cable does not need to be flexed (as for a rotary antenna), a cable with foam-PE insulation and a solid copper shield (a welded corrugated copper or aluminum tube) is the best. Such cable, commonly called hardline is perfectly water impermeable, has very high mechanical strength and lowest loss.

Flexible cables need to be used if the cable is frequently flexed. Such cables use a braided copper shield. Other cables use a non-welded metal foil with a sparsely braided copper shield

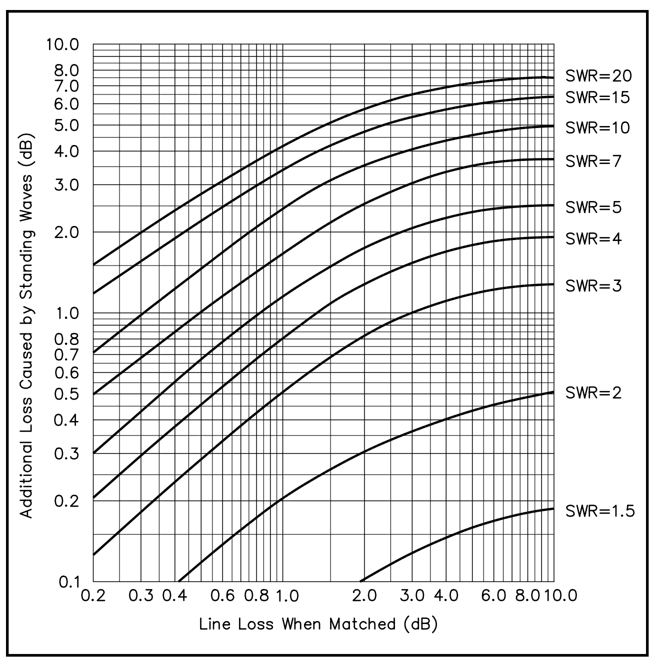


Fig 6-4 — This graph shows how much additional loss occurs for a given SWR on a line with a known (nominal) flat-line attenuation. (Courtesy of *The ARRL Antenna Book*.)

on top. The shielding effect of such cables is excellent but they should not be used for frequent flexing because the metal foil will eventually crack and rupture. Sometimes Teflon is used as dielectric material. This material is mechanically stable and electrically superior, but very expensive. Teflon-insulated coaxial cables are often used in baluns (See Section 7).

Coaxial cables commonly come in two impedances: 50 Ω and 75 Ω. For a given cable outer diameter, 75-Ω cable will show the lowest losses. That's why 75 Ω is always used in systems where losses are of primary importance, such as CATV. If power handling is the major concern, a much lower impedance is optimum (35 Ω). The standard of 50 Ω has been created as a good compromise between power handling and attenuation.

Fig 6-3 shows typical matched-line attenuation characteristics for many common transmission lines. Note how the open-wire line outperforms even its biggest coaxial brother by a

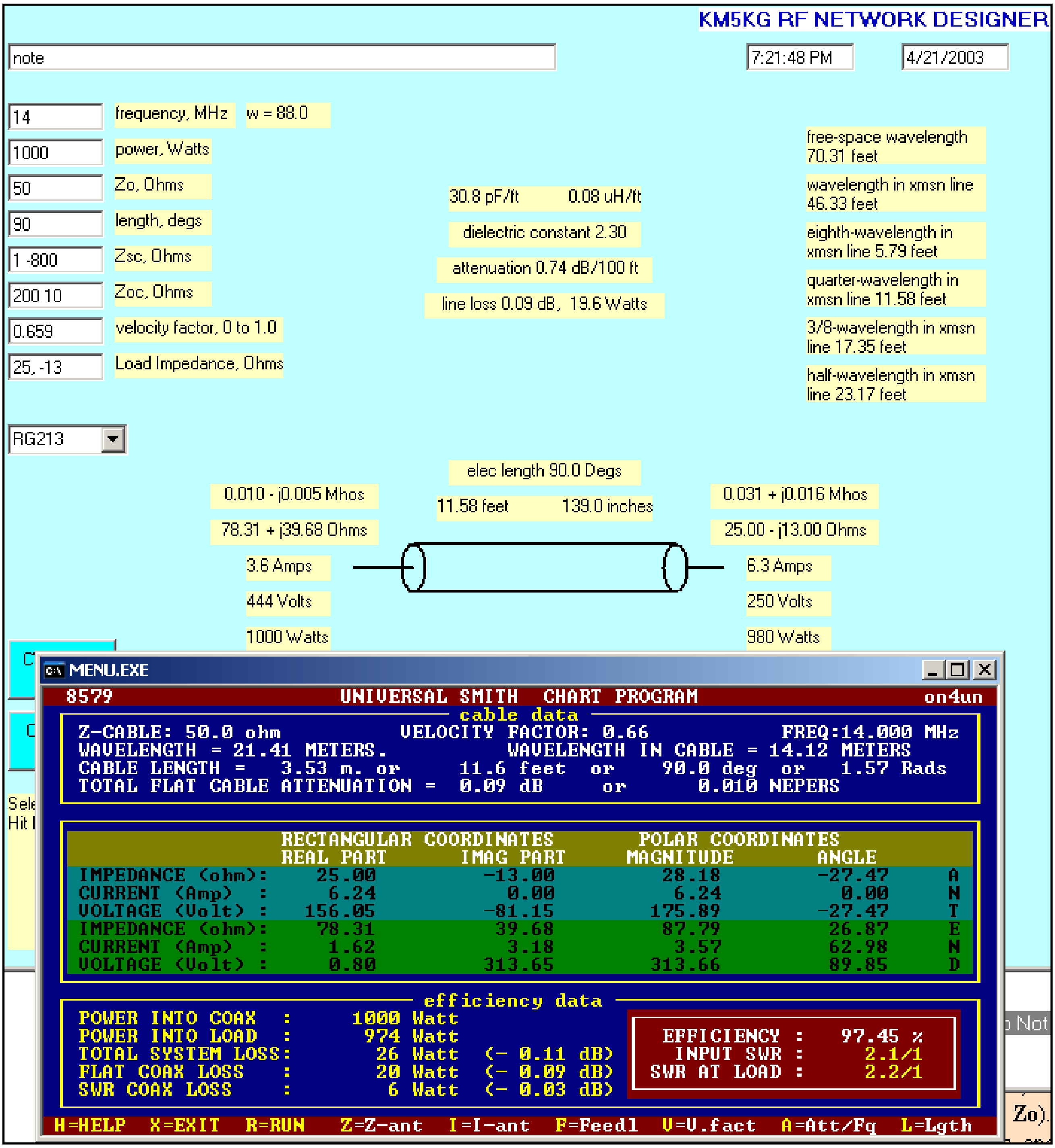


Fig 6-5 — An overlay of two transmission-line programs. At the top, the KM5KG *RF Network Designer* program shows that how a load impedance of 25 – *j* 13 Ω is transformed through a 90°-long 50-Ω feed line (having a loss of 0.74 dB/100 feet at 14 MHz). At the bottom is shown ON4UN's Universal Smith Chart, a module of the *New Low Band Software*. See text for details. Both programs calculate the impedance at the end of this (lossy) line as 78.31 + *j* 39.68 Ω.

large margin. But these attenuation figures are only the "nominal" attenuation figures for lines operating with a 1:1 SWR.

When there are standing waves on a feed line, the voltage and the current will be different at every point on the line. Current and voltage will change periodically along the line and can reach very high values at certain points. The feed line uses dielectric (insulating) and conductor (mostly copper) materials with certain physical properties and limitations. The very high currents at peaks along the line are responsible for extra conductivity-related losses. The voltages associated with the voltage peaks will be responsible for increased dielectric losses. These are the mechanisms that make a line with a high SWR have more losses than the same line when matched. **Fig 6-4** shows additional losses caused by SWR. By the way, the losses of the line are the reason why the SWR we measure at the input end of the feed line (in the shack) is always lower than the SWR at the load. An extreme example is that of a very long cable, having a loss of at least 20 dB, where you can either short or open the end and in both cases measure a 1:1 SWR at the input. Such a cable is a perfect dummy load!

For a transmission line to operate successfully under high SWR, we need a low-loss feed line with good dielectric properties and high current-handling capabilities. The feeder with such properties is the open-wire line. Air makes an excellent dielectric, and the conductivity can be made as good as required by using heavy gauge conductors. Good-quality open-wire feeders have always proved to be excellent as feed-line transformers. Elwell, N4UH, has described the use and construction of homemade, low-loss open-wire transmission lines for long-distance transmission (Ref 1320). In many cases, the open-wire feeders are used under high SWR conditions (where the feeders do not introduce large additional losses) and are terminated in an antenna tuner. On the low bands the extra losses caused by SWR are usually negligible (Ref 1319, 322), even for coaxial cables.

2.5. The Universal Transmission-Line Program

The Coax Transformer/Smith Chart computer program, which is part of the *New Low-Band Software*, is a good tool for evaluating the behavior of feed lines. You can use the module in two modes — the lossless cable mode and the real cable mode. You can analyze from the load end or from the far end. The program calculates impedance, voltage and current in both rectangular and polar coordinates as well as SWR, power loss and efficiency. See **Fig 6-5** for an example.

Fig 6-5 shows a screen print obtained from the Coax Transformer/Smith Chart module (using the program *with* cable losses). All the operating parameters are listed on the screen: impedance, voltage and current at both ends of the line, as well as the attenuation data split into nominal coax losses (0.61 dB) and losses due to SWR (0.03 dB). We also see the real powers involved. In our case we need to put 1734 W into the 100-meter long RG-213 cable to obtain 1500 W at the load, which represents a total efficiency of 86%. Note also the difference in SWR at the load (1.4:1) and at the feed line end (1.3:1). For higher frequencies, longer cables or higher SWR values, this software module is a real eye-opener.

You can do the same feed line analysis using the *Transmission Line Transformer* or the *Transmission Line Model* module from Grant Bingeman's (KM5KG) software *Professional RF Network Designer* (see Chapter 4). Fig 6-5B shows the screen result of Bingeman's program and the same using the author's program, both yielding exactly the same results.

One more program that allows you to do a thorough analysis of a feed line is *TLW* (*Transmission Line for Windows*), by N6BV. *TLW* is available from the ARRL as part of the CD that came with the 20th (and later) editions of *The ARRL Antenna Book*. *TLW* is a full-featured transmission line analysis program with beautiful graphic capabilities. The database contains transmission line characteristics of over 30 current types of lines, and the user can, in addition, enter the specs of more types of cable. The evaluation of voltage and current along the cable is displayed graphically (see **Fig 6-6**).

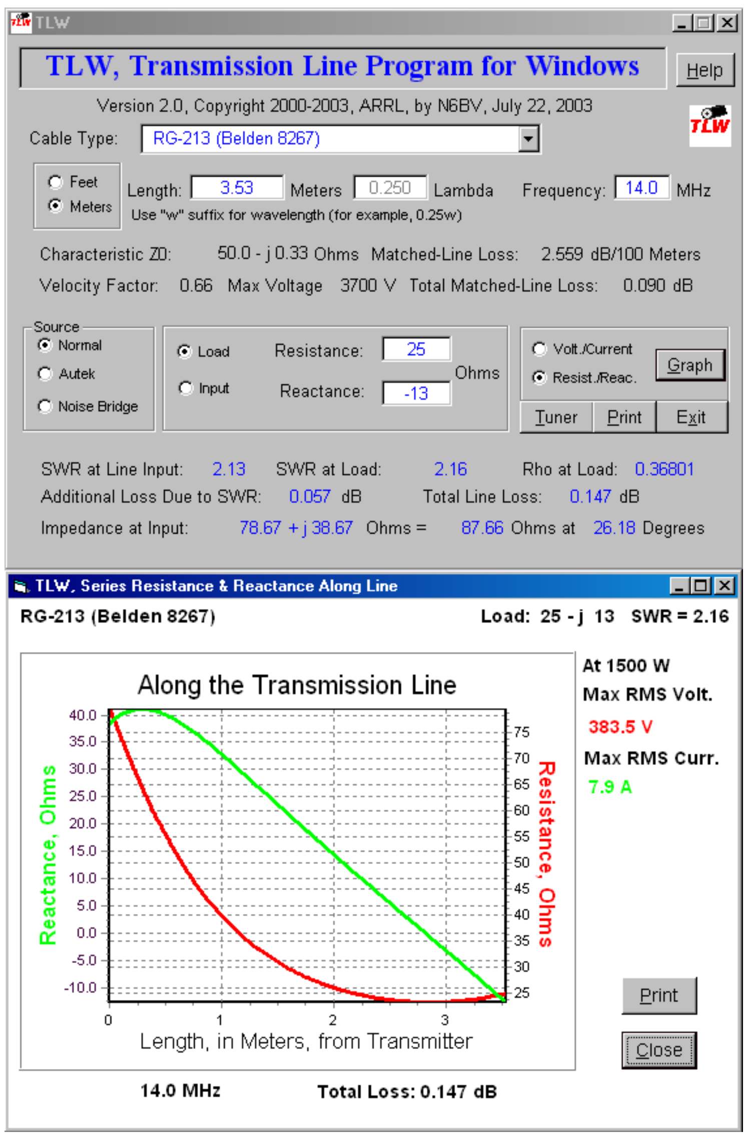


Fig 6-6 — *TLW*, the ARRL transmission line program by N6BV, is all one normally needs when analyzing the behavior of a coax line.

2.6. Which Size of Coaxial Cable?

RG-213 will easily handle powers up to 2 kW on the low bands, even with moderate SWR. Is there any point in using "heavier" coax? RG-213, when perfectly matched (SWR of 1:1) gives a loss of approximately 2 dB/100 meters on 7 MHz, 1.2 dB/100 meters on 3.5 MHz and 0.8 dB/100 meters on 1.8 MHz. What's 0.8 dB? Do you have to worry about 0.8 dB? The answer is: You need not worry about 0.8 dB on 40 meters, and even 80 meters. Top band is a different story. I have often worked a new country where 1 dB made the difference between making a QSO or not. In addition, we often see that general attitude of "don't worry, it's less than a dB" applied in various circumstances: 0.8 dB here, 0.5 dB there and again 0.3 dB somewhere else. It's the sum of all these fractions of a dB you need to worry about!

I use ⅞-inch hardline on all my antennas, even on 160 meters (loss is approximately 0.25 dB/100 meters at 1.8 MHz). If your run is 100 meters long, you "gain" 0.55 dB over the same length of RG-213, which is a gain of 13% in power. I have often said to myself, while trying to copy the weak Pacific station through the noise "I wish he was just 1 dB louder…".

An additional reason for using hardline is that it is practically indestructible. With a solid copper shield, water ingress is impossible, and the black PE jacket used on these types of cables is perfectly UV resistant for a lifetime! In addition this cable can often be obtained for less money than new RG-213 from cell phone companies renewing their sites.

2.7. Conclusions

Coaxial lines are generally used when the SWR is less than 3:1. Higher SWR values can result in excessive losses when long runs are involved, and also in reduced power-handling capability. Many popular low-band antennas have feed-point impedances that are reasonably low, and can result in an acceptable match to either a 50-Ω or a 75-Ω coaxial cable.

In some cases we will intentionally use feed lines with high SWR as part of a matching system (eg, stub matching) or as a part of a feed system for a multi-element phased array. It is good engineering practice to use a feed line with the lowest possible attenuation. This employs the concept of cost versus performance, called in the USA getting the most "bang for the buck." We would like that cable to operate at a 1:1 SWR at the design frequency of our antenna system.

3. THE ANTENNA AS A LOAD

A very small antenna can radiate the power supplied to it almost as efficiently as much larger ones (see Chapter 9 on vertical antennas), but small antennas have two disadvantages. Since their radiation resistance is very low, antenna efficiency will be lower than it would be if the radiation resistance was much higher. Further, if short antennas use loading devices, the losses of these loading devices have to be taken into account when calculating antenna efficiency. On the other hand, if the short antenna (dipole or monopole) is not loaded, the feed-point impedance will exhibit a large capacitive reactance in addition to the resistive component.

You could install some sort of remote tuner at the antenna feed point to match the complex antenna impedance to the feed-line impedance. Then the matched feed line will no longer act as a transformer itself. Matching done with such a remote tuner results in a certain sacrifice in efficiency, especially for extreme impedance ratios. Transforming a very short vertical with a feed-point impedance of, say, $0.5 - j\,3000\ \Omega$ to a $50 + j\,0\ \Omega$ transmission line is a very difficult task, one that can't be done without a great deal of loss.

You can also supply power to an antenna point without inserting a tuner at the antenna's feed point. In this case the feed line itself acts as a transformer. In the above example of $0.5 - j\,3000\ \Omega$, an extremely high SWR would be present on the feed line. The losses in the transmission line itself will be determined by the quality of the materials used to make the feed line. In pre-WWII days, when coaxial cables were still unknown, everybody used 600 Ω open-wire lines, and nobody knew (or cared) about SWR. The transmission line is fed with a low-loss antenna tuner in the shack. What is a quality antenna tuner? The same qualifications for feed lines apply here: One that can transform the impedances involved, at the required power levels and with minimal losses.

Many modern unbalanced to unbalanced antenna tuners use a toroidal transformer/balun to achieve a relatively high-impedance balanced output. This principle is cost effective, but has its limitations where extreme transformations are required. The "old" tuners (for example, Johnson Matchboxes, now museum pieces or boat anchors) are well suited for matching a wide range of impedances. Unfortunately these Matchboxes are no longer available commercially and are not designed to cover 160 meters.

4. A MATCHING NETWORK AT THE ANTENNA

Let's analyze a few of the most commonly used matching systems.

4.1. Quarter-Wave Matching Sections

For a given design frequency you can transform impedance A to impedance B by inserting a quarter-wave long coaxial cable between A and B having a characteristic impedance equal to the square root of the product A × B.

$$Z_{\lambda/4} = \sqrt{A \times B} \qquad \text{(Eq 6-2)}$$

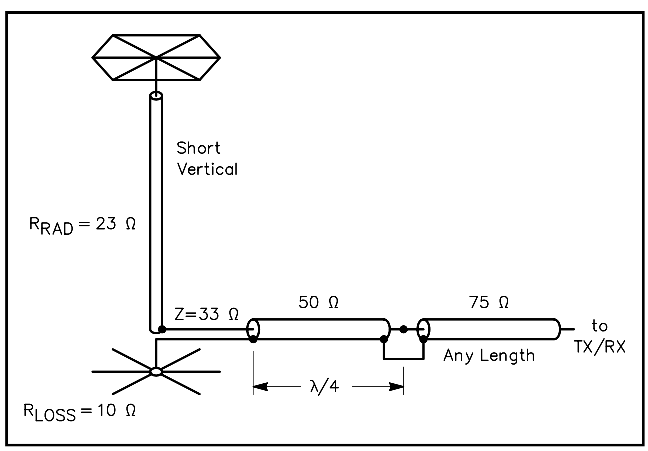


Fig 6-7 — Example of a quarter-wave transformer, used to match a short vertical antenna (R_{rad} = 23 Ω, R_{ground} = 10 Ω, Z_{feed} = 33 Ω) to a 75-Ω feed line. In this case a perfect match can be obtained with a 50-Ω quarter-wave section.

Example:

Assume we have a short vertical antenna that we wish to feed with 75-Ω coax. We have determined that the radiation resistance of the vertical is 23 Ω, and the resistance from earth losses is 10 Ω (making the feed-point resistance 33 Ω). We can use a ¼-wave section of line to provide a match, as shown in **Fig 6-7**. The impedance of this line is determined to be

$$\sqrt{33 \times 75} = 50\ \Omega$$

Coaxial cables can also be paralleled to obtain half the nominal impedance. A coaxial feed line of 35 Ω can be made by using two parallel 70-Ω cables. The Wireman (**www.thewireman.com**) sells RG-83, which is 35 Ω coax with the same OD as RG-213.

You can parallel coaxial cable of different impedances to obtain odd impedances, which may be required for specific matching or feeding purposes. See **Table 6-1**. Make sure you use cable of exactly the same electrical length! Don't fool yourself — just because you parallel three identical cables the attenuation will not be one-third the attenuation of one cable. There is no change: currents are now divided by the three cables, so all remains the same. Three cables in parallel will increase the power handling capability though.

One way to adjust ¼- or ½-wavelength cables exactly for a given frequency is shown in **Fig 6-8**. Connect the transmitter through a precision SWR meter to a 50-Ω dummy load. (If you happen to use a LP-100 digital wattmeter/SWR meter/impedance meter, you will be able to cut the quarter-wave cable very precisely.) Insert a coaxial-T connector at the output of the SWR bridge. Connect the length of coax to be adjusted at this point and use the reading of the SWR bridge to indicate where the length is resonant. Quarter-wave lines should be short-circuited at the far end, and half-wave lines left open. At the resonant frequency, a cable of the proper length represents an infinite impedance (assuming lossless cable) to the T-junction. At the resonant frequency, the SWR will not change when the quarter-wave shorted line (or half-wave open line) is connected in parallel with the dummy load. At slightly different frequencies, the line will present small values of inductance or capacitance across the dummy load, and these will influence

Table 6-1
Net characteristic impedance resulting from paralleling different coaxial cables.

Cables in Parallel	*Net Impedance*
75-Ω + 75-Ω	37.5 Ω
75-Ω + 50-Ω	30 Ω
50-Ω + 50-Ω	25Ω
75-Ω + 75-Ω + 50-Ω	21.5 Ω
75-Ω + 50-Ω + 50-Ω	18.8 Ω
50-Ω + 50-Ω + 50-Ω	16.7 Ω

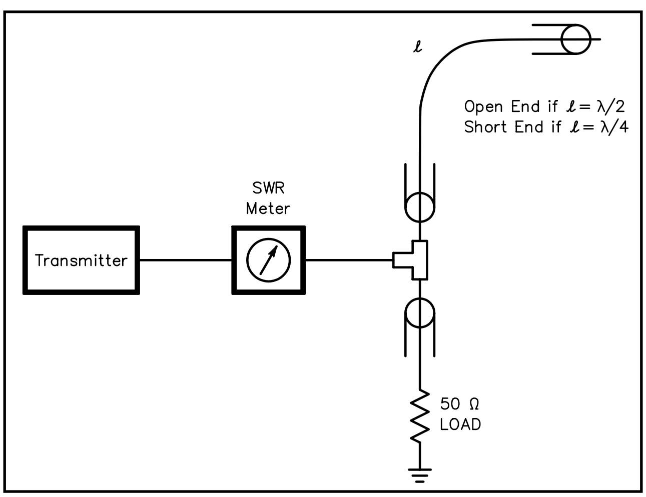


Fig 6-8 — Precise trimming of ¼ λ and ½ λ lines can be done by connecting the line under test in parallel with a 50-Ω dummy load and watching the SWR meter while the feed line length or the transmit frequency is changed. See text for details.

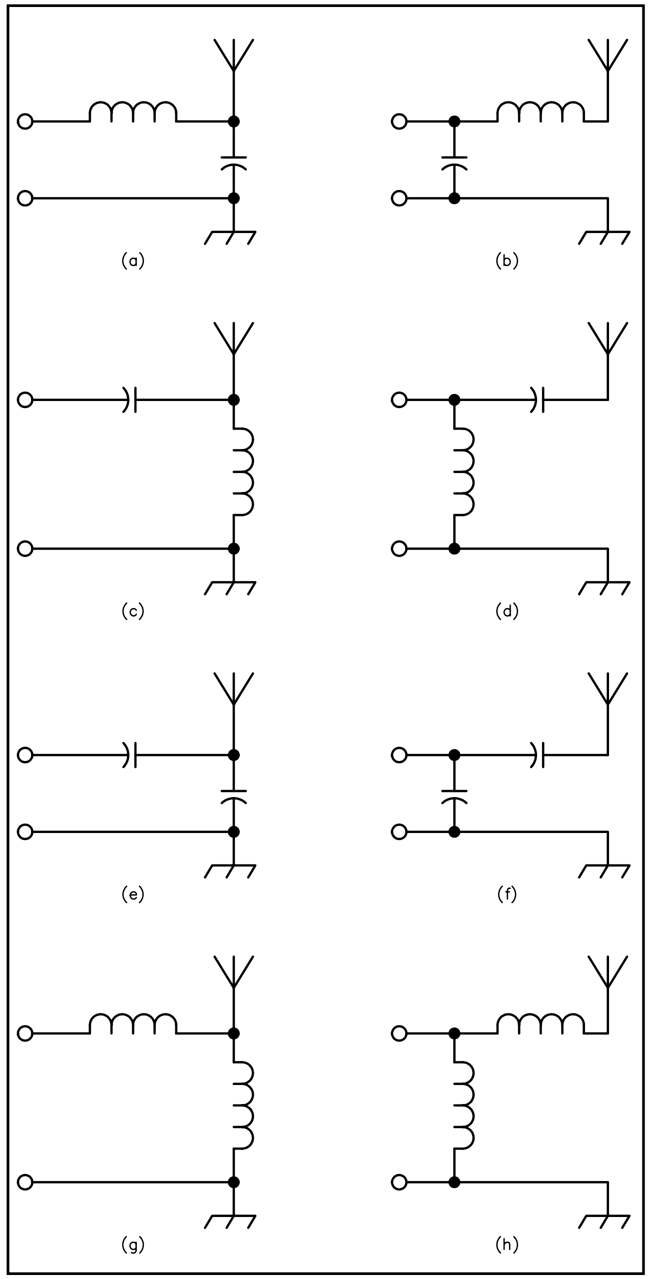


Fig 6-9 — Eight possible L-network configurations. (After W.N. Caron, *Antenna Impedance Matching*, published by ARRL but out of print.)

the SWR reading accordingly. I have found this method very accurate, and the lengths can be trimmed precisely, to within a few kHz. Alternative methods using an antenna analyzer are described in Chapter 11, Section 3.5.2.

Odd lengths, other than ¼- or ½ wavelength, can also be trimmed this way. First calculate the required length difference between a quarter (or half) wavelength on the desired frequency and the actual length of the line on the desired frequency. For example, if you need a 73° length of feed line on 3.8 MHz, that cable would be 90° long on (3.8 × 90°/73°) = 4.685 MHz. The cable can now be cut to a quarter wavelength on 4.685 MHz using the method described above.

Some people use a dip oscillator, but this method isn't the most accurate way to cut a 90° length of feed line, and it often accounts for length variations of 2° or 3° (due to the inductance of the link use to couple to the dip oscillator). You can also use a noise bridge and use the line under test to effectively short-circuit the output of the noise bridge to the receiver.

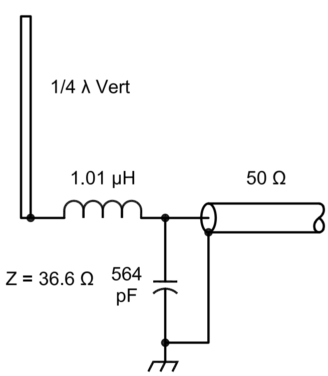
If you have an antenna analyzer or a vector network analyzer, these are very well suited for adjusting the electrical length of coaxial cables. More on these network analyzers and antenna analyzers can be found in Chapter 11.

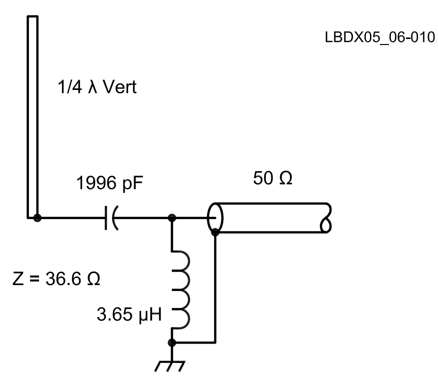
4.2. The L Network

The L network is probably the most commonly used network for matching antennas to coaxial transmission lines. In special cases the L network is reduced to a single-element network, being a series or a parallel impedance network (just an L or C in series or in parallel with the load).

The L network is treated in great detail by W.N. Caron in his excellent book *Antenna Impedance Matching* (an ARRL publication, now out of print). Caron exclusively used the graphical Smith Chart technique to design antenna-matching networks. The book also contains an excellent general treatment of the Smith Chart and other basics of feed lines, SWR and matching techniques. Graphic solutions of impedance-matching networks have been treated by I. L. McNally, W1NCK (Ref 1446), R. E. Leo, W7LR (Ref 1404) and B. Baird, W7CSD (Ref 1402).

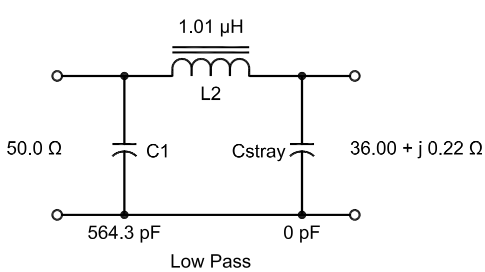
Fig 6-9 shows the eight possible L-network configurations. So-called shunt-input L networks are used when the resistive part of the output impedance is lower than the required input impedance of the network. The series-input L network is used when the opposite condition exists. In some cases, a series-input L network can also be used when the output resistance is smaller than the input resistance (in this case we have four solutions).





f = 3.5 MHz		
At 1500 W:	C1	L2
Unloaded Q	1000	200
Reactance	-80.579	22.201
Peak Voltage	387 V	202 V
RMS Current	3.4 A	6.4 A
Est. Pwr. Diss.	1 W	5 W
RMS Vin: 273.86 V at 31.49 deg		RMS Vout: 231.95 V at 0 deg.

f = 3.5 MHz		
At 1500 W:	C1	L2
Unloaded Q	200	1000
Reactance	80.193	−22.784
Peak Voltage	387 V	208 V
RMS Current	3.4 A	6.4 A
Est. Pwr. Diss.	5 W	1 W
RMS Vin: 273.86 V at −32.41 deg		RMS Vout: 231.94 V at 0 deg.



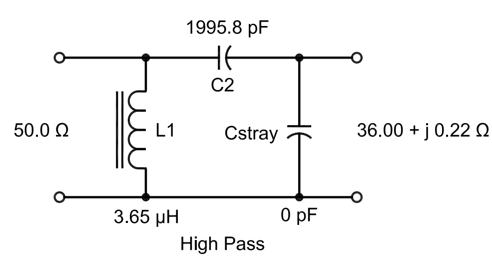


Fig 6-10 — Design of an L-network to match a resonant quarter-wave vertical with a feed-point impedance of 36.6 Ω to a 50-Ω line. Note that in practice we must add the ground resistance to the radiation resistance to obtain the feed-point impedance. Therefore, in most cases the impedance of a quarter-wave vertical will be fairly close to 50 Ω.

Table 6-2
Toroid Cores Suitable for Matching Networks

Supplier	*Code*	*Permeability*	*OD (inches)*	*ID (inches)*	*Height (inches)*	*A_L*
Amidon	T-400-A2	10	4.00	2.25	1.30	360
Amidon	T-400-2	10	4.00	2.25	0.65	185
Amidon	T-300-2	10	3.05	1.92	0.50	115
Amidon	T-225-A2	10	2.25	1.41	1.00	215

The choice of the exact type of L network to be used (low pass, high pass) will be up to the user, but in many cases, component values will determine which choice is more practical. In other instances, performance may be the most important consideration: Low-pass networks will give some additional harmonic suppression of the radiated signal, while a high-pass filter may help to reduce the strength of strong medium-wave broadcast signals from local stations.

Some solutions provide a direct dc ground path for the antenna through the coil. If dc grounding is required, such as in areas with frequent thunderstorms, this can also be achieved by placing an appropriate RF choke at the base of the antenna (between the driven element and ground).

There are several computer programs that will perform all the L-network calculations you imagine. The ARRL program *TLW* (see also Fig 6-6) can design L-networks that take into account component losses. The same is true for the L-network module from Grant Bingeman's (KM5KG) software *Professional RF Network Designer* (see Chapter 4). Both programs also calculate voltage across and current through the components. These can be used to determine the required component ratings. Capacitor current ratings are especially important when the capacitor is the series element in a network. The voltage rating is most important when the capacitor is the shunt element in the network. Consideration regarding component ratings and the construction of toroidal coils are covered in Section 4.2.1.2.

If you want to do an L-network design (without transmission line attached) using *TLW*, just enter "0" for like length. **Fig 6-10** shows both the high pass and the low pass solutions for matching an antenna with a 36 Ω impedance to a 50-Ω line.

4.2.1. Component Ratings

What kind of capacitors and inductors do we need for building the L networks?

4.2.1.1. Capacitors

The transmitter power as well as the position of the component in the L network will determine the voltage and current ratings that are required for the capacitor.

Both ARRL's *TLW* by N6BV as well as KM5KG's *Professional RF Network Designer* calculate the voltage and the current through each of the elements of the L-network. In KM5KG's program you can specify the power. In *TLW* the power is always assumed to be 1500 W. You can of course convert to other powers taking into account that $P = I^2R$ and $P = E^2/R$. This means that if you have 4 times less power (375-W), currents and voltages will be half the values shown in *TLW*.

For component rating we must always take into account the voltage peak value, while for currents we can use the RMS value. *TLW* lists these values. This is because the current failure mechanism is a thermal mechanism. In practice we should always use at least a 100% safety factor on these components. For the capacitors across low-impedance points, transmitting type mica capacitors can be used, as well as BC-type variables such as those normally used as the loading capacitor in the pi network of a linear amplifier.

For series capacitors, only transmitting type ceramic capacitors (eg, doorknob capacitors) should be used because of the high RF current. For fine tuning, high-voltage variables or preferably vacuum variables can be used. I normally use parallel-connected transmitting-type ceramics across a low-value vacuum variable (these can usually be obtained at real bargain prices at flea markets).

4.2.1.2. Coils

Up to inductor values of approximately 5 µH, air-wound coils are usually the best choice. A roller inductor comes in handy when trying out a new network. Once the computed values have been verified by experimentation, the variable inductor can be replaced with a fixed inductor. Large-diameter, heavy-gauge Air Dux coils are well suited for the application.

Above approximately 5 µH, powdered-iron toroidal cores can be used. Ferrite cores are not suitable for this application, since these cores are much less stable and are easily saturated. The larger size powdered-iron toroidal cores, which can be used for such applications, are listed in **Table 6-2**.

The required number of turns for a certain coil can be determined as follows:

$$N = 100 \times \sqrt{\frac{L}{A_L}} \qquad \text{(Eq 6-3)}$$

where L is the required inductance in µH. The A_L value is taken from Table 6-2. The transmitter power determines the required core size. It is a good idea to choose a core somewhat on the large side for a margin of safety. You may also stack two identical cores to increase power-handling capability, as well as the A_L factor. The power limitations of powdered-iron cores are usually determined by the temperature increase of the core. Use large-gauge enameled copper wire for minimum resistive loss, and wrap the core with glass-cloth electrical tape before winding the inductor. This will prevent arcing at high power levels.

Consider this example: A 14.4-µH coil requires 20 turns on a T-400-A2 core. AWG #4 or AWG #6 wire can be used with equally-spaced turns around the core. This core will easily handle well over 1500 W.

In all cases you must measure the inductance. A_L values can easily vary 10%. It appears that several distributors (such as Amidon) sell cores under the same type number coming from various manufacturers and this accounts for the spread in characteristics.

It is important to measure the inductance of a toroidal core on the operating frequency, especially when dealing with ferrite material. The impedance versus frequency ratio is far from linear for this type of material. Be careful when using a simple digital L-C meter, which usually uses one fixed frequency for all measurements (eg 1 MHz). Accurate methods of measuring impedances on specific frequencies are covered in Chapter 11 (Arrays).

4.2.1.3. The Smoke Test

Two things can go wrong with the matching network:

- Capacitors and coils can flash over (short circuit, explode, vaporize, catch fire, burn up, etc) if their voltage rating is too low.
- Capacitors or coils will heat up (and eventually be destroyed after a certain time), if the current through the component is too high or the component's current carrying capability is too low.

In the second case excessive current will heat up either the conductor in a coil or the dielectric in capacitor. One way to find out if there are any losses in the capacitor, resulting from large RF currents, is to measure or feel the temperature of the components in question (not with power applied!) after having stressed them with a solid carrier for a few minutes. This is a valid test for both coils and capacitors in a network. If excessive heating is apparent, consider using heavier-duty components. This procedure also applies to toroidal cores.

4.3. Stub Matching

Stub matching can be used to match resistive or complex impedances to a given transmission-line impedance. The Stub Matching software module, a part of the *New Low Band Software*, allows you to calculate *the position* of the stub on the line and *the length of the stub*, and whether the stub must be open or shorted at the end. This method of matching a (complex) impedance to a line can replace an L network. The approach saves the two L-network components, but necessitates extra cable to make the stub. The stub may also be located at a point along the feed line that is difficult to reach.

Fig 6-11 shows the screen of the computer program where we are matching an impedance of 36.6 Ω to a 50-Ω feed line. Note that between the load and the stub the line is not flat, but once beyond the stub the line is now matched. The computer program gives line position and line length in electrical degrees.

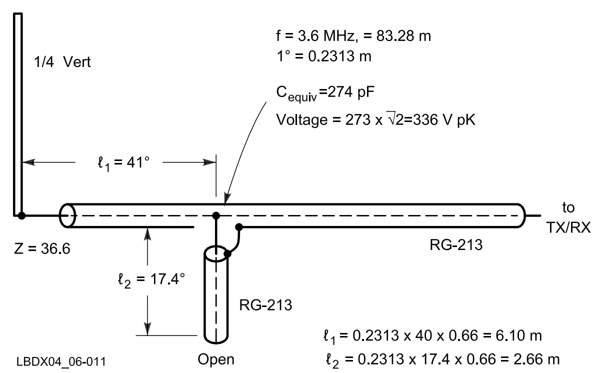


Fig 6-11—A 36.6-Ω resistive load is matched to a 50-Ω feed line using stub matching.

```
7921                          STUB MATCHING PROGRAM                        on4un

Freq:  3.6  Mhz                  Z-line: 50.0 ohm                      SWR:  1.1

                    RECTANGULAR COORDINATES      POLAR COORDINATES
                   REAL PART      IMAG PART     MAGNITUDE        ANGLE
IMPEDANCE (ohm) =   36.60          0.00          36.60           0.00      A
CURRENT (Amp)   =    6.40          0.00           6.40           0.00      N
VOLTAGE (Volt)  =  234.35          0.00         234.35           0.00      T
Posit.   IMPEDANCE        VOLTAGE          ------------STUB------------   IMP
Stub    Resis  React    Magnit   Angl     Imped    Value   Length   Type   ohm
35      43.2   12.9     265.7    43.7    -157.9    280 pF   17.6    OPEN   45.8
36      43.6   13.1     267.1    44.8    -157.7    280 pF   17.6    OPEN   46.5
37      44.0   13.4     268.6    45.8    -157.8    280 pF   17.6    OPEN   47.2
38      44.4   13.7     270.1    46.9    -158.0    280 pF   17.6    OPEN   48.0
39      44.8   13.9     271.6    47.9    -158.5    279 pF   17.5    OPEN   48.7
40      45.3   14.1     273.1    48.9    -159.2    278 pF   17.4    OPEN   49.4
41      45.7   14.4     274.6    49.9    -160.0    276 pF   17.4    OPEN   50.1
42      46.2   14.6     276.1    50.9    -161.1    274 pF   17.2    OPEN   50.9
43      46.7   14.8     277.6    51.9    -162.3    272 pF   17.1    OPEN   51.6
44      47.2   14.9     279.1    52.8    -163.8    270 pF   17.0    OPEN   52.3
45      47.7   15.1     280.5    53.8    -165.4    267 pF   16.8    OPEN   53.0
46      48.2   15.3     282.0    54.7    -167.3    264 pF   16.6    OPEN   53.7

H:HELP  X:EXIT  R:RUN  Z:Z-cable  F:Freq  I:Imp.load  C:Curr.load  V:Volt.load
```

To convert this to cable length you must take into account the velocity factor of the feed line being used.

4.3.1. Replacing the Stub with a Discrete Component

Stub matching is often unattractive on the lower bands because of the lengths of cable required to make the stub. The module Stub Matching also displays the equivalent component value of the stub (in either μH or pF). You can replace the stub with an equivalent capacitor or inductor, which is then connected in parallel with the feed line at the point where the stub would have been placed. The same program shows the voltage where the stub or discrete element is placed. To determine the voltage requirement for a parallel capacitor, you must know the voltage at the load.

Consider the following example: The load is 50 Ω (resistive), the line impedance is 75 Ω, and the power at the antenna is 1500 W. Therefore, the RMS voltage at the antenna is:

$$E = \sqrt{P \times R} = \sqrt{1500 \times 50} = 274 \text{ V}$$

Running the Stub Matching software module, we find that a 75-Ω impedance point is located at a distance of 39° from the load. See **Fig 6-12** for details of this example. The required 75-Ω stub length, open-circuited at the far end, to achieve this resistive impedance is 22.2° (equivalent to 241 pF for a design frequency of 3.6 MHz). The voltage at that point on the line is 334 V RMS (472 V peak). Note that the length of a stub will never be longer than ¼ wavelength (either open-circuited or short-circuited).

4.3.2. Matching with Series-Connected Discrete Components

In stub matching in a 50-Ω system, we look on a line with SWR for a point where the impedance on the line, together with the impedance of the stub (in parallel) will produce a 50-Ω impedance.

A variation consists of looking along the line for a point where the insertion of a series impedance will yield 50 Ω. At that point the impedance will look like 50 + j X Ω or 50 – j Y Ω. All we need to do is to put a capacitor or inductor in series with the cable at that point. A capacitor will have a reactance of X Ω or an inductor will have a reactance of Y Ω.

Example: Match a 50-Ω load to a 75-Ω line (same example as above). The software module Impedances, Cur-

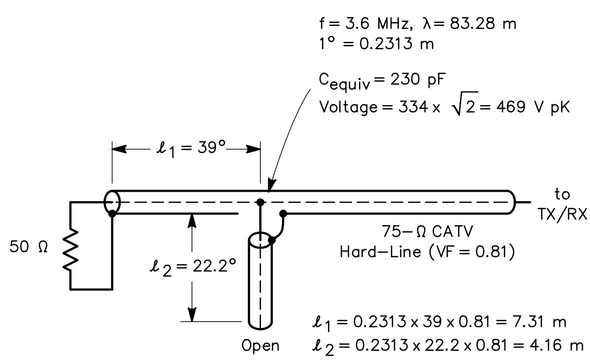


Fig 6-12—Example of how a simple stub can match a 50-Ω load to a 75-Ω transmission line. Note that between the load and the stub the SWR on the line is 1.5:1. Beyond the stub the SWR is 1:1.

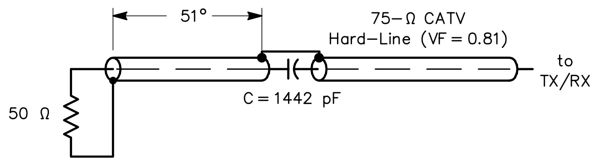
7921 STUB MATCHING PROGRAM on4un

Freq: 3.6 Mhz Z-line: 75.0 ohm SWR: 1.1

	RECTANGULAR COORDINATES		POLAR COORDINATES		
	REAL PART	IMAG PART	MAGNITUDE	ANGLE	
IMPEDANCE (ohm) =	50.00	0.00	50.00	0.00	A
CURRENT (Amp) =	5.47	0.00	5.47	0.00	N
VOLTAGE (Volt) =	273.50	0.00	273.50	0.00	T

Posit. Stub	IMPEDANCE Resis	React	VOLTAGE Magnit	Angl	-------------STUB------------ Imped	Value	Length	Type	IMP ohm
34	60.5	23.4	322.6	45.3	-180.0	246 pF	22.6	OPEN	67.8
35	61.2	24.0	324.9	46.4	-180.2	245 pF	22.6	OPEN	69.2
36	61.9	24.5	327.3	47.5	-180.7	245 pF	22.5	OPEN	70.6
37	62.6	25.1	329.6	48.5	-181.4	244 pF	22.5	OPEN	71.9
38	63.3	25.6	332.0	49.5	-182.3	243 pF	22.4	OPEN	73.3
39	64.1	26.1	334.4	50.5	-183.4	241 pF	22.2	OPEN	74.7
40	64.9	26.6	336.8	51.5	-184.8	239 pF	22.1	OPEN	76.0
41	65.7	27.1	339.2	52.5	-186.4	237 pF	21.9	OPEN	77.4
42	66.6	27.6	341.6	53.5	-188.2	235 pF	21.7	OPEN	78.7
43	67.4	28.0	343.9	54.4	-190.2	232 pF	21.5	OPEN	80.0
44	68.3	28.4	346.3	55.4	-192.5	230 pF	21.3	OPEN	81.3
45	69.2	28.8	348.6	56.3	-195.0	227 pF	21.0	OPEN	82.5

H:HELP X:EXIT R:RUN Z:Z-cable F:Freq I:Imp.load C:Curr.load V:Volt.load



$$C = \frac{10^6}{(2 \times \pi \times f \times X_c)} = 1442\ \text{pF}$$

$X_c = 30.7\ \Omega$

$F_d = 3.6,\ \lambda = 83.28\ \text{m},\ 1° = 0.2313\ \text{m}$

$\ell = 51 \times 0.2313 \times 0.81 = 9.55\ \text{m}$

Fig 6-13—Example of how a series element can match a 50-Ω load to a 75-Ω transmission line. See text for details.

8579		LOSS FREE CABLE Z / I / E LISTING PROGRAM				on4un
Z-CABLE: 75.0 ohm			STEP = 1.00 deg.		SWR = 1.50	
Length	Z-real	Z-imag	I-magnitude	I-angle	E-magnitude	E-angle
0.00	50.0	0.0	1.0	0.0	50.0	0.0
39.00	64.1	26.1	0.88	28.4	61.1	50.5
40.00	64.9	26.6	0.88	29.2	61.6	51.5
41.00	65.7	27.1	0.87	30.1	62.0	52.5
42.00	66.6	27.6	0.87	31.0	62.4	53.5
43.00	67.4	28.0	0.86	31.9	62.9	54.4
44.00	68.3	28.4	0.86	32.8	63.3	55.4
45.00	69.2	28.8	0.85	33.7	63.7	56.3
46.00	70.2	29.2	0.84	34.6	64.2	57.2
47.00	71.1	29.6	0.84	35.6	64.6	58.1
48.00	72.1	29.9	0.83	36.5	65.0	59.0
49.00	73.1	30.2	0.83	37.5	65.4	59.9
50.00	74.2	30.4	0.82	38.5	65.8	60.8
51.00	75.2	30.7	0.82	39.5	66.2	61.6
52.00	76.3	30.9	0.81	40.5	66.6	62.5
53.00	77.4	31.0	0.80	41.5	67.0	63.3
54.00	78.6	31.1	0.80	42.5	67.4	64.2
55.00	79.7	31.2	0.79	43.6	67.8	65.0
56.00	80.9	31.2	0.79	44.7	68.2	65.8
57.00	82.1	31.2	0.78	45.8	68.5	66.6
	ESC: STOP LISTING			ENTER KEY: RESUME LISTING		

7921	SERIES IMPEDANCE NETWORK (L OR C)				on4un
	RECTANGULAR COORDINATES		POLAR COORDINATES		
	REAL PART	IMAG PART	MAGNITUDE	ANGLE	
IMPEDANCE (ohm) =	75.25	30.67	81.26	22.17	O
CURRENT (Amp) =	3.44	2.83	4.46	39.46	L
VOLTAGE (Volt) =	172.12	318.82	362.32	61.64	D
IMPEDANCE (ohm) =	75.25	0.00	75.25	0.00	N
CURRENT (Amp) =	3.44	2.83	4.46	39.46	E
VOLTAGE (Volt) =	259.04	213.25	335.52	39.46	W

CAPACITANCE = 1442 pF FREQUENCY = 3.60 MHz

X = EXIT R = RUN Z = Z-load E = E-load I = I-load F = Freq

Fig 6-14—Calculations of the value of the series element required to tune out the reactance of the load 75.248 + *j* 30.668 Ω. See text for details.

rents and Voltages Along Feedlines from the *New Low Band Software* lists the impedance along the line in 1° increments, starting at 1° from the load. Somewhere along the line we will find an impedance where the real part is 75 Ω (see **Fig 6-13**). Note the distance from the load. In our example this is 51° from the 50-Ω load. The impedance at that point is 75.2 + *j* 30.7 Ω.

If we want to assess the current through the series element (which is especially important if the series element is a capacitor), we must enter actual values for either current or voltage at the load when running the program. Assuming an antenna power of 1500 W, the current at the antenna is:

$$I = \sqrt{\frac{P}{R}} = \sqrt{\frac{1500}{50}} = 5.47 \text{ A}$$

All we need to do now is connect an impedance of –30.7 Ω (capacitive reactance) in series with the line at that point. Also note that at this point the current is:

$$I = \sqrt{\frac{1500}{75.2}} = 4.46 \text{ A}$$

The software module Series Impedance Network can be used to calculate the required component value. In this example, the required capacitor has a value of 1442 pF for a frequency of 3.6 MHz (see **Fig 6-14**). The required voltage rating (RMS) is calculated by multiplying the current through the capacitor times the capacitive reactance, which yields a value of E = I × Z = 4.46 × 30.7 = 136.9 V RMS = 193.6 V peak at 1500 W. As outlined above you need to take the peak value into consideration for a capacitor, and apply a safety factor of approximately two. The most important property of this capacitor is its current-handling capability, and we should use a capacitor that is rated approximately 10 A for the job.

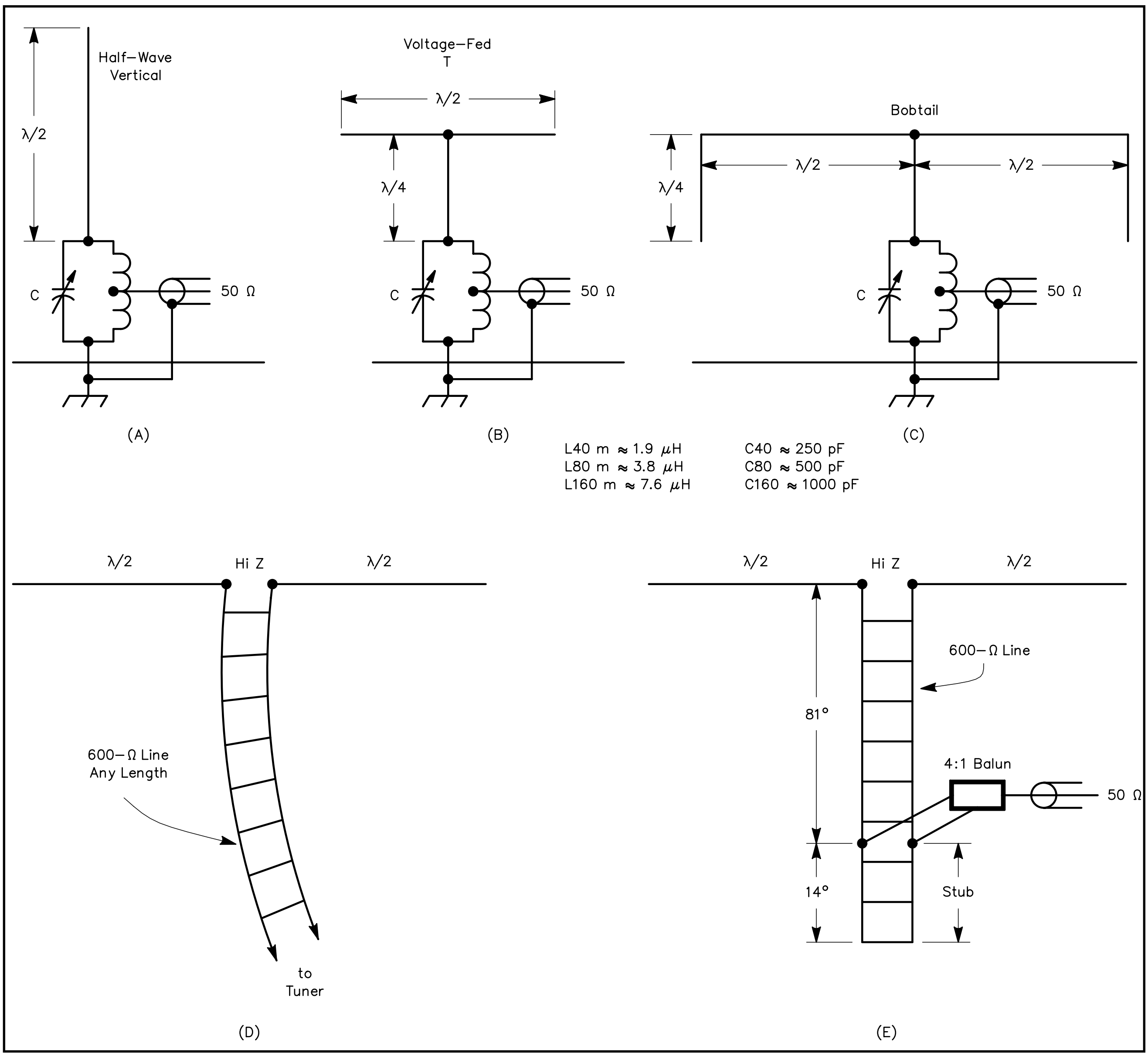


Fig 6-15 — Recommended feed methods for high-impedance (2000 to 5000-Ω) feed points. Asymmetrical feed points can be fed via a tuned circuit. The symmetrical feed points can be fed via an open-wire line to a tuner, or via a stub-matching arrangement to a 4:1 (200 to 50-Ω) balun and a 50-Ω feed line.

In the case of a complex load impedance, the procedure is identical, but instead of entering the resistive load impedance (50 Ω in the above example), we must enter the complex impedance.

4.4. High-Impedance Matching Systems

Unbalanced high-impedance feed points, such as a half-wave vertical fed against ground, a voltage-fed T-antenna, the Bobtail antenna, etc, can best be fed using a parallel-tuned circuit on which the 50-Ω cable is tapped for the lowest SWR value. See **Fig 6-15**. Symmetrical high-impedance feed points, such as for two half-wave (collinear) dipoles in phase, the bi-square, etc, can be fed directly with a 600 Ω open-wire feeder into a quality antenna tuner (see Fig 6-15D).

Another attractive solution is to use a 600 Ω line and stub matching, as shown in Fig 6-15E. Assume the feed-point impedance is 5000 Ω.Using the Stub Matching software module, we find that a 200 Ω impedance point is located at a distance of 81° from the load. The required 600 Ω stub to be connected in parallel at that point is 14° long (X = 154 Ω). The impedance is now a balanced 200 Ω. Using a 4:1 balun, this point can now be connected to a 50-Ω feed line.

Let me sum up some of the advantages and disadvantages of both feed systems.

Tuned open-wire feeders:
- Fewest components, which means the least chance of something going wrong.
- Least likely loss.
- Very flexible (can be tuned from the shack).
- Open-wire lines are mechanically less attractive.

Stub matching plus balun and coax line:
- Coaxial cables are much easier to handle.

4.5. Wideband Transformers

4.5.1. Low-Impedance Wideband Transformers

Broadband transformers exist in two varieties: The classic autotransformer and the transmission-line transformer. The first is a variant of the Variac, a genuine autotransformer. The second makes use of transmission-line principles. What they have in common is that they are often wound on toroidal cores. It is beyond the scope of this chapter to go into details on this subject. More details can be found in Chapter 7 (Receiving Antennas, where such broadband transformers are commonly used to feed receiving antennas such as Beverages), and in Chapter 11, section 3.4.6.3. *Transmission Line Transformers* by J. Sevick, W2FMI, is an excellent textbook on the subject of transmission-line transformers. It covers all you might need in the field of wide-band RF transformers.

4.5.2. High-Impedance Wideband Transformer

If the antenna load impedance is both high and almost perfectly resistive (such as for a half wavelength vertical fed at the bottom), you may also use a broadband transformer such as is used in transistor power amplifier output stages. **Fig 6-16** shows the transformer design used by F. Collins, W1FC. Two turns of AWG #12 Teflon-insulated wire are fed through two stacks of 15½-inch (OD) powdered-iron toroidal cores (Amidon T-50-2) as the primary low impedance winding. The secondary consists of 8 turns. The turns ratio is 4:1, the impedance ratio 16:1.

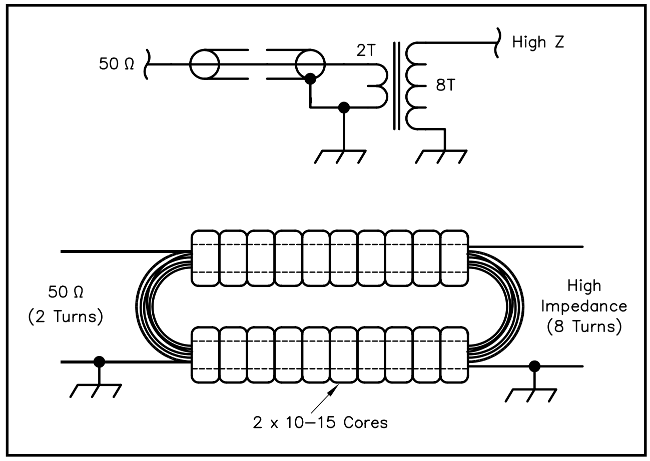


Fig 6-16 — A wideband high-power transformer for large transformation ratios, such as for feeding a half-wave vertical at its base (600 to 10,000 Ω), uses two stacks of 10 to 15 half-inch-OD powdered-iron cores (eg, Amidon T502-2). The primary consists of 2 turns and the secondary has 8 turns (for a 50- to 800-Ω ratio).

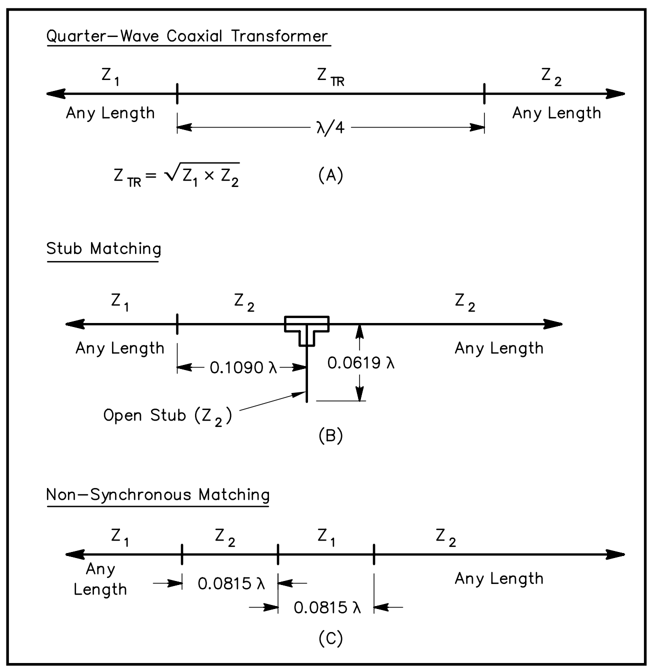


Fig 6-17 — Methods of matching 75-Ω cables in 50-Ω systems. The quarter-wave transformer at A requires a cable having an impedance that is the geometric mean of the values being matched. The stub matching system at B and the non-synchronous matching system at C require only cables of the impedances being matched. The stub can be replaced with a capacitor or an inductor. All these matching systems are frequency sensitive.
Z_{TR} — 60-Ω line.
Z_1 — 50-Ω line (or load).
Z_0 — 75-Ω line.

The efficiency of the transformer can be checked by terminating it with a high-power 800 Ω dummy load (or with the antenna, if no suitable load is available), and running full power to the transformer for a couple of minutes. Start with low power. Better safe than sorry. If there are signs of heating in the cores, add more cores to the stack. Such a transformer has the advantage of introducing no phase shift between input and output, and therefore can easily be incorporated into phased arrays.

5. 75-Ω CABLES IN 50-Ω SYSTEMS

Lengths of 75-Ω hardline coaxial cable can often be obtained from local TV cable companies. If very long runs to low-band antennas are involved, the low attenuation of hardline is an attractive asset. If you are concerned with providing a 50-Ω impedance, you need to use a transformer system. Transformers using toroidal cores (so called *ununs*) have been described (Ref 1307, 1517, 1518, 1521, 1522, 1523, 1524, 1525, 1526, 1527, 1528, 1829, 1830).

Ununs (unbalanced-to-unbalanced transformers) are really *autotransformers* and have been described for a wide range of impedance ratios. One application is as a matching system for a short, loaded vertical. If the short, loaded vertical is used over a good ground radial system, its impedance will be lower than 50 Ω. Ununs have been described that will match 25 Ω to 50 Ω, or 37.5 Ω to 50 Ω.

Ununs can also be used in array-matching systems to provide proper drive for various elements (see Chapters 7 and 11). Transformer systems can also be made using only coaxial cable, without any discrete components. If 60-Ω coaxial cable is available (as in many European countries), a quarter-wave transformer will readily transform 75 Ω to 50 Ω at the end of the hardline.

Carroll, K1XX, described the non-synchronous matching transformer and compared it to a stub-matching system (Ref 1318). While the toroidal transformer is broadbanded, the stub and non-synchronous transformers are single-band devices.

Compared to quarter-wave transformers, which need coaxial cable having an impedance equal to the geometric mean of the two impedances to be matched, the non-synchronous transformer requires only cables of the same impedances as the values to be matched (see **Fig 6-17**).

On the low bands (and even up to 30 MHz) the losses caused by using 75-Ω hard line in a 50-Ω system (50-Ω antenna and 50-Ω transceiver/amplifier) are generally negligible. A real problem is that 75-Ω feed line itself works as a transformer, and even when terminated with a perfect 50-Ω load, will show 100 Ω at the end of the line if the line is an odd multiple of ¼ λ long. This may cause problems for your linear amplifier.

There is an easy solution to that problem, which is using ½-λ (or multiples of) lines. If you use a multiband antenna, make sure that the line is a number of half waves on all the frequencies used. For an antenna that works on 80 and 160 meters, make the coaxial line a multiple of half waves on 160 meters. Assuming a 75-Ω hardline with a velocity factor (VF) of 0.8, then the line should be $0.8 \times (300/1.83)/2 = 65.6$ meters, or any multiple thereof. You can trim the length by terminating the line with a 50-Ω load, and adjusting the length for minimum SWR on the highest frequency (in the above case, 3.66 MHz).

Don't fool yourself though. In this case the SWR on the 75-Ω line is still 1.5:1, but the consequences are minimal so far as additional losses are concerned (because we use a feed line with intrinsic low losses) and are compensated for as far as the transformation effect is concerned, by using ½-λ lengths. To be fully correct the transformation is not a perfect 1:1 transformation with a real line, but close (1:1 is only with a lossless line).

Greg Ordy, W8WWV developed *SMC* (*Series Matching Calculator* software) that calculates non-synchronous

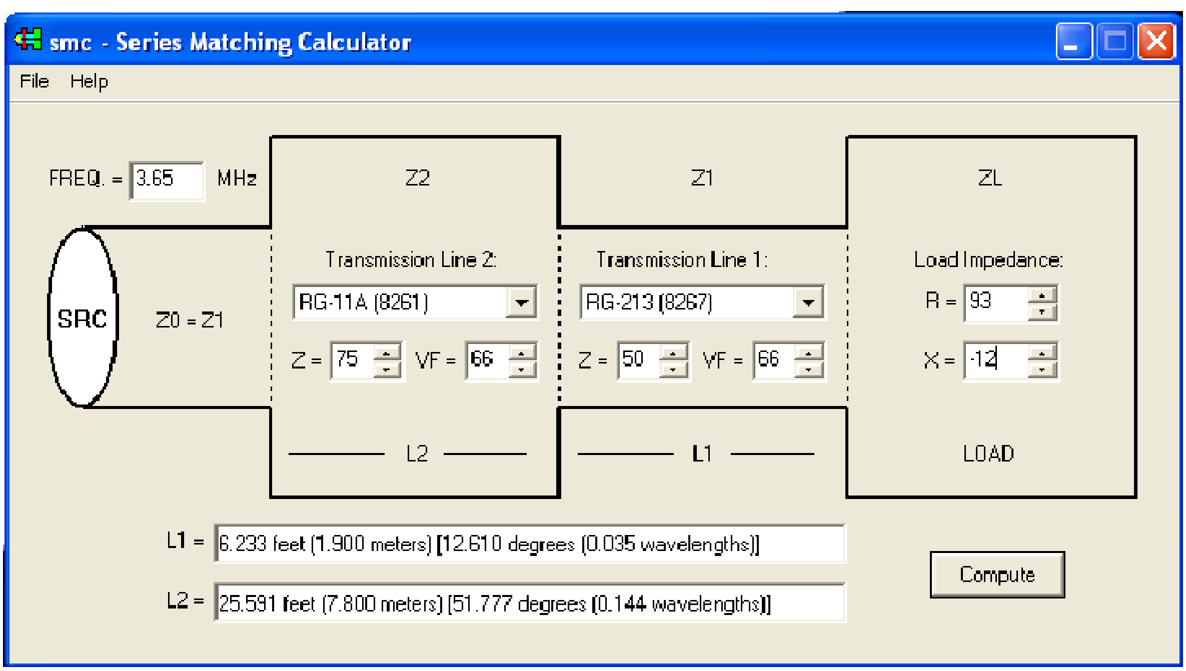


Fig 6-18 — Two short pieces of transmission lines with different characteristic impedances act as two impedance transformers is series. Their lengths have been chosen so that at the end of the second transmission line the impedance equals the characteristic impedance of the first transmission line.

transformers that can be used in a relatively wide range of impedance matching situations, and also if the load has a complex load impedance (**www.seed-solutions.com/gregordy/Software/SMC.htm**). This technique will not work for all load impedances. There is a finite range of impedance values which can be matched by this technique. The range is related to the impedance values of the two transmission lines. See **Fig 6-18**.

6. THE NEED FOR LOW SWR

In the past many radio amateurs did not understand SWR. Unfortunately, many still don't understand SWR. Reasons for low SWR are often false and SWR is often cited as the single parameter telling us all about the performance of an antenna.

Maxwell, W2DU, published a series of articles on the subject of transmission lines. They are excellent reading material for anyone who has more than just a casual interest in antennas and transmission lines (Refs 1308-1311, 1325-1330 and 1332). These articles have been combined and, with new information added, published as a book. (The most recent edition as of this writing is *Reflections III*, available from *CQ* Communications.) J. Battle, N4OE, wrote a very instructive article "What is your Real Standing Wave Ratio" (Ref 1319), treating in detail the influence of line loss on the SWR (difference between apparent SWR and real SWR).

Everyone has heard comments like: "My antenna really gets out because the SWR does not rise above 1.5:1 at the band edges." Low SWR is no indication at all of good antenna performance! It is often the contrary. The "antenna" with the best SWR is a quality dummy load. Antennas using dummy resistors as part of loading devices come next (Ref 663). The TTFD (Tilted Terminated Folded Dipole) and the B&W broadband folded dipole model BWD-18-30 are such examples. You should conclude from this that low SWR is no guarantee of radiation efficiency. The reason that SWR has been wrongly used as an important evaluation criterion for antennas is that it can be easily measured, while important parameters such as efficiency and radiation characteristics are more difficult to measure.

Antennas with lossy loading devices, poor ground systems, high-resistance conductors and the like, will show flat SWR curves. Electrically short antennas should always have narrow bandwidths. If they do not, it means that they are inefficient. The two real reasons are covered in detail in Chapter 5, Section 2.11.

7. THE BALUN OR COMMON MODE CHOKE

Balun is a term coming from the words *balanced* and *unbalanced*. It is a device we must insert between a symmetrical feed line (such as an open-wire feeder) and an asymmetric load (such as a ground-mounted vertical monopole) or an asymmetric feed line (for example, coax) and a symmetric load (such as a center-fed half-wave dipole). If we feed a balanced feed point with a coaxial feed line, currents will flow on both the outside of the coaxial braid (where we don't want them) and on the inside (where we do want them). Currents on the outside will cause radiation from the line.

Unbalanced loads can be recognized by the fact that one of the terminals is at ground potential. Examples: the base of a monopole vertical (the feed point of any antenna fed against real ground), the feed point of an antenna fed against radials (that's an artificial ground), the terminals of a gamma match or omega match, etc.

Balanced loads are presented by dipoles, sloping dipoles, delta loops fed at a corner, quad loops, collinear antennas, bi-square, cubical quad antennas, split-element Yagis, the feed points of a T match, a delta match, and so on.

Many years ago I had an inverted-V dipole on my 25-meter tower and the feed line was just hanging unsupported alongside the tower, swinging nicely in the wind. When I took down the antenna some time later, I noticed that in several places where the coax had touched the tower in the breeze, holes were burned through the outer jacket of the RG-213. Further, water had penetrated the coax, rendering it worthless. The phenomena of burning holes illustrates that currents (thus also voltages) are present on the coax if no balun is used. Such currents also create radiated fields, and fields from the feed line upset the field pattern from the antenna.

How much radiation there is from such a feed line depends on several factors, the main one being its length. In most cases the feed-line outer conductor will be (RF) grounded at the station. Assume the feed line (seen as a long wire, which means with a velocity factor of 98%) is an odd number of quarter-waves long. In that case the impedance of the long wire (which is the outer shield of the feed line) will be very high at the antenna feed point, and hence the currents will be minimal, resulting in low unwanted radiation from it. If, however, the feed line is a number of half-waves long (and the outer shield grounded at the end), then we have a low-impedance point at the antenna end and consequently a large current can flow.

In actual practice, unless the feed lines are a multiple of half-wavelengths long, the impedance of the "long-wire" will be reactive. (Watch out: it's not wavelengths of the transmission line but wavelengths of the coax shield acting as a "fat antenna wire".) In parallel with the resistive and low impedance of the real antenna (at resonance), that will result in a relatively small current flowing on the outer shield of the coaxial feed line. The best answer is "take no chances" and always use a current balun.

Baluns have been described in abundance in the amateur literature (Refs 1504, 1505, 1502, 1503, 1515, 1519, and 1520 through 1530).

The traditional balun (for example, the well-known W6TC balun, see **Fig 6-19**) is a *voltage balun*, which produces equal, opposite-phase voltages into the two resistances. With the two resistances we mean the two "halves" of the load, which are "symmetrical" with respect to ground (not necessarily in value!). If the load is perfect in common-mode balance and of a controlled impedance, a voltage-type balun is as good as a choke-type balun. But the choke-type balun is almost always much better in the real world. The choke type balun is commonly called a current balun. The toroidal-core type voltage and current baluns are covered in *Transmission Line Transformers* by J. Sevick, W2FMI.

The current type balun is based on the principle of making the shield of the coaxial feed line, seen from the outside, to represent a high total impedance so that RF current *on* the shield is minimized. This total impedance consists of two parts: a resistive (ohms) part and a reactive part (usually inductive). In order to make the choke-type balun as broadbanded as possible, we want to have as a high a resistive part as possible, which makes the choke a very low-Q choke and avoids resonance issues.

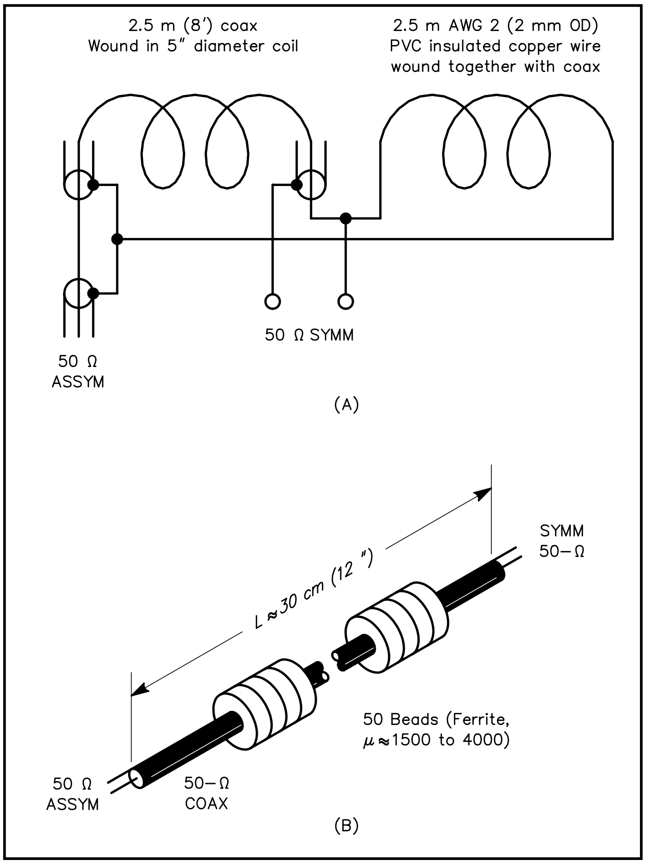


Fig 6-19 — At A, details of a W6TC voltage-type balun for 160-40 meters, and at B, a current transformer for 160-10 meters. See text for details.

A Coil of Coaxial Cable

In the simplest form a common-mode choke (current balun) consists of a number of turns of coaxial cable wound into a close coil. In order to present enough reactance at the low-band frequencies, a fairly large coil is required. This is a (very) high Q current balun, as the total impedance is formed almost exclusively by the inductance of the coil of coax.

Using Ferrite Beads on the Coaxial Cable

Another approach was introduced by Maxwell, W2DU. This involves slipping a stack of high-permeability ferrite cores over the outer shield of the coaxial cable at the load terminals. Cores made of the proper ferrite material increase the inductance of the coaxial line as well as introduce a fair to high degree of loss resistance for the coax shield (seen from the outside). Such current chokes, when properly designed can be very broadbanded, which is not the case with a simple coil of coax (see above).

In order to reduce the common-mode currents flowing on the feed line, one must achieve a total choking impedance of at least 1000 Ω (preferably more). In a 50-Ω system a 1000 Ω choke will reduce the current on the feed line shield by 20 × log (50/1050) = 26 dB.

In order to reduce the required ID of the toroids or beads, you can use a short piece of Teflon-insulated coaxial cable such as RG-141, RG-142 or RG-303. These have ODs of approximately 5 mm. A balun covering 3.5 to 30 MHz uses 50 no. 73 beads (Amidon no. FB-73-2401 or Fair-Rite no. 2673002401-0) to cover a length of approximately 30 cm (12 inches) of coaxial cable. On 160 meters one must use 100 such beads to reach the required 1000 Ω choking impedance.

Amidon beads type 43-1024 can be used on RG-213 cable. About 10 to 30 beads will be required, depending on the lowest operating frequency.

The two above approaches are called *current baluns, choke baluns* or *common mode baluns*. They are called current-type baluns because even when the balun is terminated in unequal resistances, it will still force equal, opposite-in-phase currents into each resistance.

Differential Mode and Common Mode

Differential mode signals are signals that are present between the conductors of a cable, for example between the wires of a twisted pair, between the conductors of a parallel open wire line, or between the inner conductor and the inside of the shield of a coaxial line. At any place in the cable the current in the two conductors will be out of phase (180º phase shift). With *common mode* signals all conductors of a cable carry the same signals (in phase). In the case of a coaxial cable common mode signals are present on the outside of the cable shield (which acts as a single "fat" conductor). RF signals picked up by an antenna wire travel on the wire as common mode signals. RF signals picked up by unshielded cables in the various conductors cause common mode currents to flow in these conductors. RF signals picked up by the shield of a coaxial cable cause common mode currents to flow on the outside of the shield.

I have stated on several occasions that if the reading of an SWR meter changes with its position on the line (small changes in position, not affected by attenuation) this means the SWR meter is not functioning properly. The only other possible reason for a different SWR reading with position on the line is the presence of RF currents on the outside of the coax. For that reason it is common practice in professional SWR-measuring setups to put a number of ferrite cores on the coaxial cable on both sides of the measuring equipment.

We've touched upon three good reasons for using a balun with a symmetrical feed point:

- We don't want to distort the radiation pattern of the antenna.
- We don't want to burn holes in our coax.
- We want our SWR readings to be correct.

Are there good reasons to put a so-called current balun on a feed-line attached to an asymmetrical feed point? Yes, there are. Assume a vertical antenna using two elevated radials. The feed point is an asymmetric one, but the ends of the two radials are *not* the real ground nor a perfect ground, If we do not connect a current balun on the feed line at the antenna feed point, antenna return currents will flow on the outside of the coaxial feed line in addition to flowing in the elevated radials, which is not what we want with elevated radial (see also Chapter 8, Section 1.5 and Chapter 9, Section 2.2.14). The feed point of a vertical antenna using an elevated radial systems is never "perfectly asymmetrical." In other words the common point of the radials are never really at ground potential. Therefore, when you use a vertical with an elevated radial system always put a current balun at the feed point.

Is it harmful to put a current balun on all the coaxial antenna feed lines for all your antennas? Not at all. If the feed point is asymmetric, there will be no current flowing and the beads will do no harm. As a matter of fact they may help reduce unwanted coupling from antennas into feed lines of other nearby antennas. A good thing is to use an RF-current meter (see Chapter 11) and check currents on the outside of any feed line while transmitting on any nearby (within ½ wavelength) antenna. These currents should be zero; if not, they act as parasitically excited elements, which will influence the radiation pattern of your antenna.

The ferrite cores used as beads in a common-mode choke are not lossless, and depending on the mix used, they can be quite lossy. The losses make the beads heat up when the power handled is high.

Where no or very little power is involved (such as for solving EMC problems or on receiving antennas) this is never a problem. As explained above it is the total loss of the RF choke (made up by the impedance of the inductance in series with the loss resistance) that chokes off the unwanted currents. In a low Q situation (R >>) where the ferrite cores are used to choke off potentially high RF currents (this is mostly the case with current baluns on transmitter feed lines), the resistive losses of the ferrites may actually heat those up to the point where they either become totally ineffective (permanently destroyed) or actually crack or explode! This problem can be avoided by using ferrite material that is not very lossy on the transmit frequency. In that case the choke will require more beads and the choke will be less broadbanded.

In actual practice you can successfully combine two sorts of ferrite cores in a current choke balun: low resistive (high Q) cores at the "hot side" of the balun and lower-Q beads at the "cold side). In practice the touch-and-feel method is an adequate test method. First run reduced power. If some of the cores get warm at 100 W, chances are you will destroy them with a kW.

When running high power, *very high*-µ beads made of low-Q core material can get quite warm, especially those nearest to the current source. A remedy is to use 50 ferrite beads of #43 material (µ ~ 850) installed near the current source (antenna), followed by 50 beads of #73 material (µ ~2500), slipped over a 0.6 meter long piece of Teflon coax (RG-303/U). Make sure the beads can cool to ambient air. I have been using this type of current choke quite successfully on my 80 meter Four Square, running high power (2 kW).

Commercially-made current baluns are available from different sources. W8JI tested a series of baluns for use on 160 meters. His conclusion is that the DX Engineering balun (model DXE-BAL050-H05-P) really performs. Beside the issue of the impedance offered by the balun to currents flowing on the outside of the coaxial feed line, there also is an issue of power handling capability of the balun, which is a major concern with most baluns (see **www.w8ji.com/Baluns/balun_test.htm**).

The Wireman (**www.thewireman.com**) sells a kit (#833) at a very attractive price. The kit consists of a length of Teflon coax (RG-141 or RG-303) plus 50 ferrite beads (#73 material) to be slipped over the Teflon coax. Array Solutions (**www.arraysolutions.com**) also has a common-mode choke (model AS-50-L1), which they call a "50-Ω line insulator."

Current chokes are also extensively used with special receiving antennas (Chapter 7, Section 2.7.2.9).

8. CONNECTORS

A good coaxial cable connector, such as a PL-259 connector, has a loss of less than 0.01 dB, even at 30 MHz, and typically 0.005 dB or less on the low bands. This means that for 1 kW of power you will have a heat loss of about 1 W per connector. Given the mass of a connector, and the heat-dissipating capacity of the cable, this will produce a hardly noticeable temperature increase. If you feel a connector getting hot (with "reasonable" power) on the low bands, then there is something wrong with that connector. You needn't avoid connectors for their high intrinsic losses, as claimed by some.

When using connectors make sure they are well installed, and properly waterproofed. Despite what some may claim, N connectors will easily take 5 kW on the low bands, and over 2 kW on 30 MHz. N connectors are waterproof and the newer models are extremely easy to assemble (much faster than a PL-259). A PL-259 connector is not a constant-impedance connector, but that is not relevant on the low bands. It is, however, a connector that is difficult to waterproof without external means. I always use a generous amount of medical-grade petroleum jelly (Vaseline) inside the connector to keep moisture out. Some cheaper coax, as well as semi-air-insulated coax, may see the inner conductor retract or protrude after time. Such coaxial cables are best used with PL-259 connectors, where you can mechanically anchor the inner conductor in the connector by soldering. In an N connector, the retracting inner conductor sometimes will retract the connector pin to the point of breaking the contact.

9. BROADBAND MATCHING

A steep SWR curve is due to the rapid change in reactance in the antenna feed-point impedance as the frequency is moved away from the resonant frequency. There are a few ways to try to broadband an antenna:

- Employ elements in the antenna that counteract the effect of the rapid change in reactance. The so-called "Double Bazooka" dipole is a well-known (and controversial) example. This solution is dealt with in more detail in the chapter on dipoles.
- Instead of using a simple L network, use a multiple-pole matching network that can flatten the SWR curve. This solution is covered in great detail in *Antenna Impedance Matching*, by W.N. Caron, published by the ARRL but out of print.

ANTMAT is a computer program described in technical Document 1148 (Sep 1987) of the NOSC (Naval Ocean Systems Center). The document describing the matching methodology as well as the software is called "The Design of Impedance Matching Networks for Broadband Antennas." The computer program assists in designing matching networks to match antennas (such as small whip antennas) over a wide frequency range (an abstract is available at **adsabs.harvard.edu/abs/1987nosc.reptQ....L**).

Contents

CHAPTER 7

Receiving Antennas

Six years ago **Robye Lahlum, W1MK**, sent me a lot of good information on a new feed system for arrays, information which I could use for the 4th edition of this book. I met Robye the first time in Dayton in 2005, where together we did a presentation at the Hamvention antenna forum. Over the years we have kept in touch via a QSO now and then, mainly on the 160 meter band. His station excels in signal strength, proof that what he's doing with his antennas is right (and it is not done from a 50 or 100 acre lot way out in the country). Indeed, once more the proof of the pudding is in the eating. During the preparation phase of this 5th edition, we inevitably we got in touch again, and Robye had a good deal of new material to propose to me.

Robye grew up in North Dakota. He went to North Dakota State University and earned his BSEE degree in 1963. He moved to the Boston area in the mid 1960s, where he earned an MSEE from Northeastern University in 1965. After graduate school Robye worked for Bell Labs for 37 years in the Boston area.

He has been a ham since 1955, and consecutively held the calls WØGBQ, K4JEP, W1EEF and now W1MK. At Bell Labs he worked for Frank Witt, AI1H, the author of a whole series of articles on 80-meter broadband antennas. He also knew Jerry Sevick, W2FMI (now a Silent Key) at Bell Labs, and saw a lot of his transmission line transformers when he worked there.

After working together with Robye on the previous edition of this book, I knew I wanted to do more joint efforts, and that's why I asked Robye to be the godfather of this chapter. I feel I could not find a better man to help me in making this chapter even better than in the last edition. It was great working on a project like this with Robye. His technical knowledge is profound.

Thank you Robye for your encouragement, support and suggestions. A perfect godfather I could always turn to for help.

Roger Vermet, ON6WU, the man who for nearly 30 years now has been helping me with all my antenna projects, has been very helpful during the rewriting phase of this and many other chapters of this book. He did a lot of experiments and lab tests to confirm what I wanted to tell my readers. During the early months of 2009 we spent whole days together, week after week, to discuss important technical subjects as well as the results of the many lab tests he did for this book. Thank you Roger for your friendship and help.

I also would like to thank **Brian Mattson, K8BHZ**, the "inventor" of the Six Circle, who suggested some changes to the section covering his "baby."

Some of the questions I will try to answer in this chapter:

- Why do we need separate receiving antennas?
- What is noise? How can I eliminate noise?
- Are Beverages so superior?
- Is a longer Beverage better?
- How about vertical receiving arrays?
- What's the correct way of feeding special receiving antennas?
- Can I do as well from my city lot as the big guns from their rural farms?
- Are Flags, Pennants and K9AY loops an alternative to Beverages?
- Why not receiving arrays with parasitic elements?

In an e-mail K9RJ wrote, "The challenge of 160-meter (low band) DXing is receiving. It should be no surprise that the highest DXCC totals on this band are achieved only by those who have the space for good receiving antennas, or who live in a location where much of the DX is close by. I'm not aware of any exceptions to this. Thus, the greatest need is for creative development of low-noise directional receiving antennas or techniques such as active noise canceling that can be used to improve receiving capability."

Not so long ago, any mention of "receiving antenna" usually invoked thoughts of "Beverage antennas." The evolution in all technical fields is staggering, and it includes receiving antennas. Not that something spectacularly new has been invented, but our ability to communicate worldwide on a regular basis has improved drastically thanks to the Internet. Technical knowledge is spread more easily, and technical discussions have become accessible to nearly everyone interested.

1. INTRODUCTION TO RECEIVING ANTENNAS

This chapter is not a Beverage-only chapter. It will not even start with Beverages. Readers have asked for more receiving antennas, so here they are! However, before we get into describing receiving antennas and antenna projects in detail, it is important to understand a few basics.

1.1. Why Separate Receiving Antennas?

Separate antennas are necessary because optimum receiving and transmitting have different requirements. For a transmit antenna, we want maximum possible field strength in a given direction (or directions) at the most useful elevation (wave) angles. We cannot tolerate unnecessary power loss in a transmit antenna, because any amount of transmitting loss decreases signal-to-noise ratio at the distant receiver. Antenna efficiency is an important issue for transmitting. It is obvious that for a given elevation angle and direction, the highest gain antenna will deliver the strongest signal to the target area. We really do not care if we are being heard in other directions (areas) or not; we are only interested in the target direction.

Choosing a transmit antenna is a matter of properly positioned gain. Transmitting antennas require high directivity to achieve high gain, not directivity just for the sake of reducing the strength of the signal transmitted in unwanted directions. Tom, W8JI, at **www.w8ji.com** adds to that: "Takeoff angle is not important. What we actually need is maximum possible gain at the desired angle and direction. After all, we don't care where the peak is as long as the antenna we pick has more signal (gain) at the desired spot than other antenna choices!"

A receiving antenna on the other hand has a different design priority. The goal is obtaining a signal that can be read comfortably, which means having the greatest possible signal-to-noise (S/N) and signal-to-QRM ratio. Receiving antennas providing the best performance can and will be different under different circumstances, even at the same or similar locations. There is no such thing as a universal "best low-band receiving antenna."

Why doesn't the reciprocity law apply to signal-to-noise ratio as it applies to signal level? It's easiest to explain this with an example: Consider a high-band Yagi with 7-dBd gain, including ground-reflection gain. This antenna will improve the transmitted signal by 7 dB over a dipole, provided both have peak gain oriented to the target area. Does a 3-element Yagi with the same efficiency as a dipole improve reception S/N by the same amount as it improves transmission?

The answer is simple: Probably not! S/N will improve much more than the 7-dBd gain when very strong noise sources are located in a pattern null. If the null is –25 dBd, S/N can increase as much as 32 dB (+7 dBd signal to –25 dBd noise). Of course, the improvement will normally be less than 32 dB, since that is extreme.

If the noise comes from exactly the same direction as the desired signal, the Yagi's 7-dBd gain will not improve S/N at all. The Yagi will deliver equally increased signal and noise power, both being 7 dB stronger than the dipole.

The Yagi also might have decreased efficiency. This is actually very common, because of losses caused by increased element current. In reality, a 7-dBd Yagi often has more than 7-dBd directivity. If Yagi efficiency were only 50%, 7-dBd gain would require having 10-dB directivity increase. These are all reasons why gain does not determine receiving S/N improvement, and why the higher-gain antenna very often does not provide the best reception.

There is one predictable effect of gain. In the forward direction of the antenna signal levels will be increased by the amount of gain, both in transmitting and receiving (keep in mind though that signal level is not the same as signal-to-noise ratio). Continuing the example above and assuming perfect lobe alignment with the path, the Yagi's signal level will be 7 dB above the dipole in the target area. The distant receiver will always have 7 dB more S/N when the Yagi is used. This is true regardless of any S/N improvement we might or might not observe when receiving with the same Yagi. What counts for improving communications is the ratio of signal-to-noise on both receiving ends of the circuit. In practice this means there is no reciprocity in readability— reciprocity only applies to signal level. This is *not* "one-way" propagation, although it sometimes may cause people to think this is happening.

1.2. Gain Versus Directivity

Gain is a function of efficiency and directivity. High gain means an antenna has high directivity and "high" efficiency. The increased field strength comes with a price. The extra energy found in the main lobe is energy that was removed from other directions (also see Chapter 5, Section 1.2).

The answer to improved receiving can be the same as for transmitting. Installing a highly directive transmit antenna results in high-performance receiving, so long as the antenna is not aligned with or installed near noise sources. Unfortunately, the physical size and height of efficient antennas — especially on 160 and 80 meters — often makes high-gain transmitting antennas prohibitively expensive.

Fortunately, high or even modest efficiency is not a direct requirement for directivity and receiving. This chapter will show it is possible to build relatively small receiving antennas that exhibit excellent directivity and greatly improve reception, even though the antennas are useless for transmitting because of high losses and low gain.

Directivity is not the same as gain. It is possible to construct very directive antennas that actually have negative gain but that provide phenomenal receiving improvements. It is worth repeating: We need directivity — not gain — for a good receiving system.

The next question is: How much negative gain can we live with? The answer is fairly simple once an antenna is installed. If you can easily detect a background noise increase when a dummy load is removed (from the receiver input) and the antenna connected under the quietest operating conditions (usually winter daytime within a few hours of sunrise or sunset) with the narrowest IF filter selected, gain is OK! As Tom, W8JI, puts it with regard to preamps and matching devices in particular: "Once you clearly hear external noise, amplifiers or impedance matching won't help. Just be sure you can hear noise at the quietest time you expect to operate."

We learned in Chapter 3 that our present-day receivers have a large sensitivity margin when used with reasonably efficient antennas, especially considering the large amount of noise on the low bands (unless you live on a desert island or in the wilderness). Most receivers are sensitive enough to use with antennas having –10 to –20 dBi gain, depending on various factors (see Chapter 3). For the rest, we can always use a preamplifier to boost the signal to a more comfortable level.

In very quiet locations, with 250-Hz selectivity, a minimum discernable signal sensitivity of –140 to –145 dBm might be required while using narrow-pattern, low-efficiency receiving antennas. In suburban locations, –125 to –135 dBm sensitivity is often adequate (see also Chapter 3).

Very directional antennas and narrow receiver selectivity reduce noise power, requiring less receiving system sensitivity to yield a satisfactory output S/N. Since noise power is proportional to bandwidth, a 250-Hz filter requires 10 dB less receiver sensitivity compared to the same system using a 2.5-kHz filter. Directivity has the same effect when noise is evenly distributed. A 3 dB increase in directivity for a given amount of antenna gain will provide 3 dB less noise power, and required receiver sensitivity decreases by 3 dB. Sensitivity is adequate when external noise from outside the antenna system clearly dominates the receiver noise at the narrowest selectivity being used.

1.3. Noise

We have covered the nature of noise and its intensity in different environments (urban, suburban, rural) in detail in Chapter 3. What is noise? Noise is the sum of many signals, with most sources unintentional. We can distinguish three sorts:

- Noise generated by nature: noise from thunderstorms (static, QRN) and precipitation static.
- Noise generated by man: mostly from arcs or rapidly switched sources (spikes or square wave signals), such as power lines, switching power supplies, digital systems, electronic voltage controls such as dimmers or motor speed controls, defective doorbell transformers, lighting systems, electric fences, thermostats and so on.
- Noise generated by poorly designed, operated or maintained transmitters: CW clicks, sideband splatter, noise sidebands, spurious oscillations and other transmitter defects.

When we consider how noise propagates or travels to our location, we can distinguish:

- Near-field noise generated in the antenna system or coupled directly to the antenna though induction or electric fields from nearby wiring. This near-field noise includes precipitation static, but is mostly man-made switching or sparking noise.
- Fresnel region noise generated outside the induction field area but before the antenna pattern is completely formed. This noise includes man-made noises, such as those from arcing high-voltage wiring or strong local static discharges.
- Noise propagated from the far field by ground wave or ionospheric propagation. This noise includes CW clicks, sideband splatter, noise sidebands, lightning noise and other natural and man-made sources. It includes the sum of many hundreds of thousands of low-level noise sources, such as the accumulated noise from entire cities.

We usually refer to the sum of all *unidentifiable* noises as *band noise* or even *background noise*. We generally classify identifiable noise generated by intentional transmitters as QRM, although the end effect is largely the same as any other noise.

In quiet rural locations (away from polar regions) lower-frequency band noise is evenly distributed at all wave angles and directions whenever darkness surrounds the receiving location. Noise is only lacking in directions where propagation is very poor, or directions with a total lack of noise sources. We do not consider the sky as "quiet" on the lower bands because the ionosphere reflects all types of very small noise sources from both nearby and distant sources. While the amount of noise from each source might be very small, the accumulated effect of innumerable noise sources is a smooth broadband hissing noise.

In some locations, QRM consistently arrives from well-defined directions. If you live in Western Europe, almost all QRM (transmitter generated noise) arrives from the East. In the very northeast coast of Canada, QRM generally arrives from the mainland USA to the southwest. In many locations, QRM comes from many (if not all) directions, with nearly random distribution. You may want a different antenna pattern when DXing on relatively clear bands compared to patterns used during crowded contests.

1.4. Reducing Various Noise Types

Noise has exactly the same characteristics, so far as an antenna is concerned, as signals from *intentional* transmitters. There is no way to sort "good signals" from "bad noise" except through directional characteristics or *directivity* of the receiving antenna.

External noises can be eliminated or reduced only by the principle of phase opposition: Receive the noise with at least two different antennas (elements) and add the signals received from the elements in such a way that the sum is zero (equal amplitude and 180° out-of-phase). We can do this using arrays (groups of antennas) or using a special configuration where one antenna is usually a small noise pick-up antenna and the second one is the regular receiving antenna. In this case a so-called noise-canceller will combine the two signals to cancel a given noise signal (see Section 1.5).

Different types of noise are controlled through different methods. There are three primary sources of noise:

- Noise from thunderstorms.
- Precipitation static.
- Man-made noise.

1.4.1. Noise From Thunderstorms

If a very active thunderstorm is local (directly overhead), noise is the least of our worries. We really should disconnect lightning-sensitive or inadequately protected equipment (before the storm) and stay away from the radios! If the storm is somewhere in the distance (usually covering a wide azimuth), an antenna with a very broad pattern null and extremely good front-to-storm direction ratio will help (see Chapter 5, Section 2.10).

1.4.2. Precipitation Static

While often attributed to charged particles (such as water droplets) hitting an antenna, most precipitation static is actually caused by intense electric field gradients in the area surrounding the antenna. Such conditions commonly appear during inclement weather, when movement of particles or moisture causes concentrated areas of charges. The strong electric fields are responsible for noise-producing *corona discharges*. The noise comes from low-current corona discharges from sharp or protruding objects.

Sailors saw this effect on tall-masted ships, calling it *St Elmo's fire*. This noise generally builds slowly from a sizzle to a high-pitched whine and disappears with nearby lightning flashes. Lightning "equalizes" the potential difference between earth and nearby clouds, reducing the charge gradient and corona. Since this noise is generated in or very near the antenna, directivity is of no help.

Using an antenna at a lower height reduces corona current as the electric-field gradient is smaller close to the wide smooth surface of the earth. This is especially true when the low antenna is surrounded by taller structures. Round, smooth and insulated conductors are helpful because they reduce voltage gradient and resulting corona discharges. Vertical antennas are particularly sensitive to precipitation static; they have pointed ends protruding upward toward the oppositely charged sky. The corona also comes from the very high-impedance antenna end, which aids in coupling power into the receive system.

Beverages on the other hand, being near earth, will have fewer corona discharge problems. They also have low surge impedances. This means the low-current high-voltage arcs transfer very little noise power into the antenna. Beverages are thus quite resistant to precipitation static.

Quads are more resistant than Yagis because quads have long flat sides with blunt lower-impedance high-current areas toward the sky. Yagis have protruding high-impedance pointed ends. Low-current arcs are not only more likely to happen in Yagis; they are also better impedance matched to the antenna! Quads have a reputation for being "quiet antennas," but this only applies to corona. For all other noises quads are no better than any other antenna.

1.4.3. Man-Made Noise

Local man-made noise is received several ways. When the source is a modest distance (1 to 10 km) away, noise arrives by ground wave propagation. If noise comes from just outside or nearly outside the antenna's Fresnel zone, it can be eliminated with pattern nulls. The Fresnel-zone area is where the pattern is not fully formed. The zone is related to array size. It can extend a few kilometers with a very large array, particularly one using broadside elements on low frequencies. If the noise source cannot be eliminated using a directive antenna, we often make use of so-called *noise-cancellers* to solve the problem. If the noise source is from a single source we can define a few solutions.

1.4.3.1. Single-Point Radiation Far Source

Local noise arriving from one clear radiation point, even if multiple sources, can easily be nulled out. The antennas need not be similar, but deep nulls require two antennas that both "hear" the noise. The sense antenna should be placed closer to and directly in-line with the noise source. The spacing can be nearly any distance, but $\lambda/4$ or more is always best. There must be a stable RF phase relationship between the noise received in the main receive antenna and the noise-sense antenna. Since local noise is received by surface or ground wave, the phase, polarization, and amplitude are constant. This allows a stable deep null to be obtained using equipment such as the MFJ-1025/1026 or the DX Engineering NCC-1 (see Section 1.37).

1.4.3.2. Distributed Radiation Source

Noise from a single source or multiple sources can be fully nulled if the distance to the radiation area is large compared to the length of the radiating area. This is true even if the noise follows power lines and radiates from multiple points or is from multiple sources. The sense antenna must clearly and strongly pick up the noise. The ideal case is where the sense antenna is very close to the source and the signal antenna is a much larger distance away. If however the noise source is right on your street and the radiating power lines are in front of your house, it is likely that all of this happens in the near field of both the receive and sense antennas, and in that case nulling will be impossible.

In all cases the sense antenna should ideally hear only the noise and not the wanted signals, which means it must be fairly close to noise source. And the sense antenna should be fairly small.

1.4.3.3. Nearby Man-Made Noise

If the noise source is very close (in the near or *induction* field), it becomes difficult or impossible to eliminate noise through antennas arrays. In this case the problem must be tackled in a different way, either by eliminating the noise source or experimenting (trial and error) with various antennas. Using a portable receiver or a fox-hunting (DFing) receiver for 160 or 80 meters, local sources can be found easily. If the noise cannot be killed, such a single source noise, even in the near field, can often be completely nulled out provided the sense antenna is installed near the noise source, and the main receiving antenna is located farther from the noise. It's obvious, however, that the best solution in this case is to "kill" the noise source directly.

1.4.3.4. Propagated Noise

This noise generally sounds like a smooth hiss, even though it is coming from hundreds or thousands of raspy or harsh noise sources. Propagated noise is rarely, if ever, audible in urban areas on 160 meters, since it is masked by harsh local noises. Propagated noise is sometimes audible in quieter directions of suburban areas on 160 meters, but not in "noisy" ground wave directions or if a local dominant noise is present. Propagated noise is often responsible for the entire noise floor in remote rural areas. It is often possible to find the direction of strong band openings by looking for highest propagated noise, because the enhanced propagation can sum countless noise sources for many thousands of km! Unfortunately, as Tom, W8JI, says: "Propagated noise reduces the advantage of super-quiet locations during the night." Hearing propagated noise is a good indicator of how quiet your location is and how good your receiving system is. For example, the winter season 160-meter daytime-to-nighttime noise level increase at W8JI has been measured at 15 dB. This is in the absence of thunderstorms within many thousands of miles.

While local man-made noise can often be nulled, propagated noise is another story. Canceling propagated noise only works with antennas of identical polarization and similar patterns. The antennas must be close to each other, so they receive signals in a constant phase and amplitude relationship, with no space diversity. However, propagated noise constantly changes phase, polarization and amplitude. Different types of antennas in a canceling system respond differently, making canceling impossible or very unstable. For any relief from such propagated noise, it must arrive from a significantly different direction than the desired signal.

1.4.3.5. QRM

Interference from CW clicks, splatter, noise sidebands, etc is usually called QRM, but it is just another form of noise. We do not deal with QRM any differently from the way we deal with other propagated noises. If you are lucky, the QRM does *not* come from the same direction as the desired signal. If it does, there is very little you can do about such noise with your antennas.

1.5. Noise Suppression and Noise Canceling

What is the principle of canceling and suppressing? It's really fairly simple. First, we must receive the unwanted signal with two different antennas. The main antenna would receive as much desired signal as possible. Ideally, the second antenna would hear only noise, with very little desired signal. The noise outputs of the antennas would then be adjusted so they are exactly equal, and the results combined exactly out-of-phase (180°). Total canceling would occur when these two conditions are met. If the noise antenna hears very little desired signal, all noise from the main source would be removed, without any change in desired signal level.

Noise cancellers are simple in theory. They allow adjustment of level, and rotation (or shift) of phase. When selecting a noise canceller, the following technical parameters are important:

- Low amplitude change with phase adjustment
- Wide amplitude range
- Wide phase range
- No loss
- Immunity to overload (good dynamic range).

Noise cancellers are most frequently used to cancel the noise from a single noise source. But, provided they are designed for it, they can also be used to combine signals of a receiving array without the aim of canceling a specific noise source (see Section 1.37).

Homebrewers should exercise caution in selecting a noise-canceling circuit design. Some designs are very poor because they do not actually rotate or shift phase. It is impossible to shift phase in a simple transformer system, since transformers only invert phase. We cannot mix only 180° out-of-phase signals to obtain phase variation. L/C circuits, R/C circuits, or delay lines must be used.

The best noise-canceling circuits are bridge-type phasing systems. These circuits look much like a standard Wheatstone bridge, except a relatively high value reactance is substituted for at least one resistance. If such a circuit drives a high-impedance load, considerable phase shift can occur with minimal amplitude change.

1.6. Directive Receiving Arrays: How to Obtain a Null

Let's develop an antenna array that has directivity in a given direction and that produces a null in another direction. In order to form a deep predictable null, we need two antennas with nearly identical patterns. Let's assume we will use two vertical antennas. In terms of physical size, the most efficient 2-element combination is an *end-fire array*. This is an array where maximum radiation occurs in-line with the elements. The ideal spacing ranges from $\lambda/3$ downward.

We will bring the RF signals received by the two antennas to a common point, where these signals will be *combined*. By judiciously determining the physical spacing between the two antennas, as well as possibly "slowing down" the wave traveling in one of the feed lines toward the signal combining point, we can develop an antenna system that receives better from certain directions. Signals are "slowed down" by making them travel a longer distance in its feed line — the difference in distance is called the phase delay line or the phasing line. The system receives better in the direction where the signals arriving at the *combining point* from the two antenna elements add up — combining in phase or "nearly" in phase. It does not receive as well in other directions, where these signals subtract or even cancel one another (combining out of phase).

In what follows we assume no "space diversity." In other words, the received signals arrive in a constant phase and amplitude relationship (which is usually the case if the two elements are spaced less than 1 to 1.5 λ apart). Since the two antenna elements are not at exactly the same physical point, an incoming wave takes a small extra time to travel to the furthest antenna. Put another way, the two antenna elements receive the same signal at a slightly different time, which also means with slightly different phase differences for different directions of arrival. The exact phase difference depends on the distance between elements and the angle at which the signal arrives (in both the horizontal and vertical plane). The largest phase difference occurs when signals arrive in-line with the elements.

Let us assume there are two vertical elements spaced λ/4 (90°). Refer to **Fig 7-1**. A signal coming from the right (in-line with the two antennas) at zero elevation angle will arrive at the second antenna (B) later than the first one (A). The phase difference will be 90°. To completely cancel this signal you must combine the 90° shifted outputs (due to the physical separation) exactly 180° out-of-phase. You can do this by connecting impedance-matched feed lines to both antennas, making the feed line to antenna B 90° electrically longer than the line to antenna A. If you connect these two lines together using a method that produces equal currents in each antenna, the signals coming from the back will cancel out.

Signals from the front direction add quite differently. As they arrive at A, they have a spatial delay of 90° at A. Since the feed line to B also delays phase 90°, the total phase shift is 0°. Signals from the direction of B are in phase and add together (fully in phase).

While many people use 90° shift with 90° phasing, it is not the optimum phase delay. What amateur operator wants maximum nulling at a 0° elevation angle? Few signals arrive at (or nearly at) a zero elevation angle, except ground-wave signals that are perfectly in-line with the array.

We often need to place the null at higher angles, or move it slightly off the back. This not only increases usefulness of the null, it increases gain and directivity of the array.

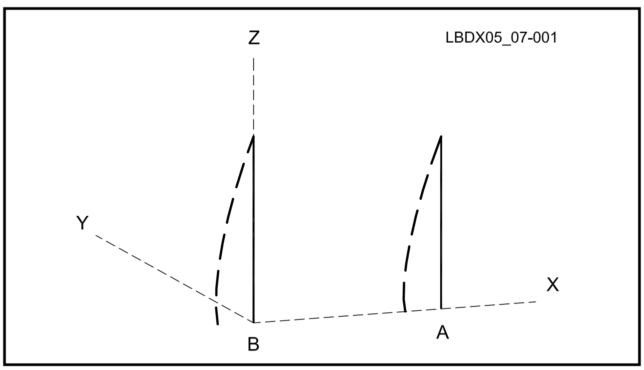


Fig 7-1 — A signal arriving off the back of the 2-element end-fire array hits element A earlier than element B. See text for discussion of how the directivity pattern is obtained.

Let's look at how to produce a null at a given wave angle, in this example 52.5°. Refer to **Fig 7-2**. Both antennas are still spaced spaced λ/4 apart. The rearward signal again arrives at antenna A before antenna B, but this time the difference will be shorter than 90°. A little trigonometry shows us that the spatial phase delay is now 90° × cosine (52.5°) = 55°. To create a null, we have to combine signals exactly out-of-phase, but the spatial delay is now 55°. The extra delay in the feed line to element B becomes 180° – 55° = 125°. Signals from the front are now 125° – 55° = 70° out-of-phase.

In this λ/4 spacing example, we can obtain a null at any wave angle by changing phase delay between 90° (λ/4 long for 0° null angle) and 180° (λ/2 long for 90° null angle). This is very useful, and it also works for other element spacings if we use the proper phase delay ranges. (We must always be sure the feed and phasing system compensates for impedance changes.)

Null elevation is not the only pattern change. The array phasing change we did above also affects the null in the azimuth plane. In **Fig 7-3** we can see the same geometry and trigonometry applies when examining the azimuth pattern at 0° elevation angle over a perfect ground. At a 0° wave angle, the 125° array phasing moves full cancellation to two points 52.5° either side of an imaginary line drawn through the two antennas.

The horizontal directivity patterns as developed by a 2-element end-fire array (as shown in Figs 7-2 and 7-3) are called *cardioid patterns*.

At different elevation angles, nulls are present in different directions. **Fig 7-4** shows a three-dimensional view of the pattern. The null actually forms a deep cone surrounding the back lobe. This much wider null greatly increases the area of zero response. Removing the response over this large area decreases noise pick up and improves gain and directivity of this array. It also gives two deep nulls along the ground and pulls in the sides of the pattern, decreasing antenna response to ground-wave noise.

The offset angle for maximum attenuation is identical in the horizontal and vertical plane. This is useful information. We can adjust an antenna for maximum attenuation at a 52.5° elevation by adjusting for maximum attenuation 52.5° offset

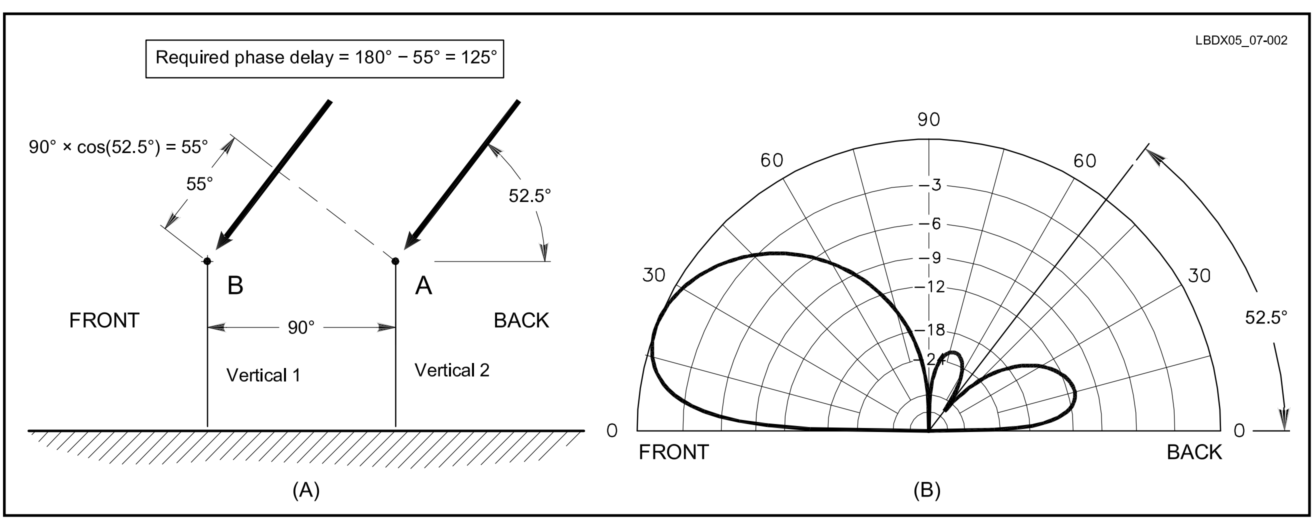


Fig 7-2 — Development of a null at a given elevation angle in a 2-element end-fire array. See text for details.

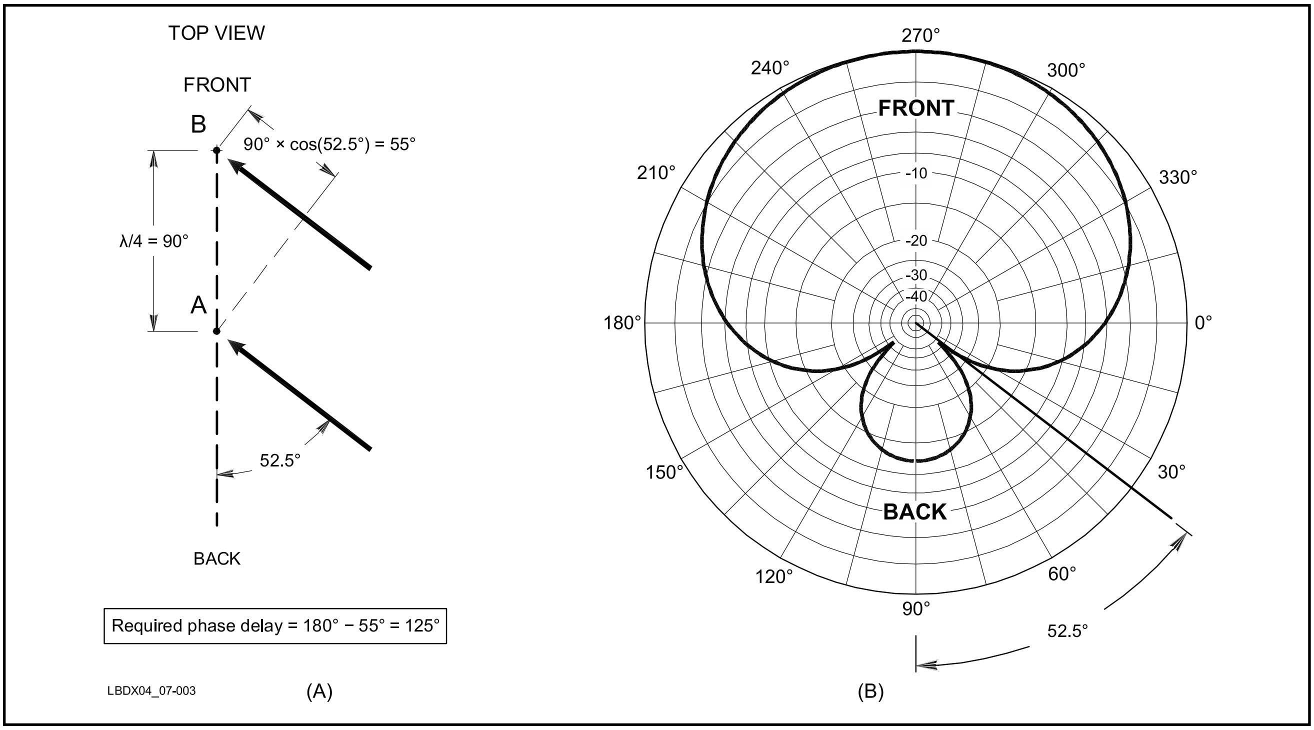


Fig 7-3 — The same trigonometry applies when looking at the null angles in the horizontal plane. See text for details.

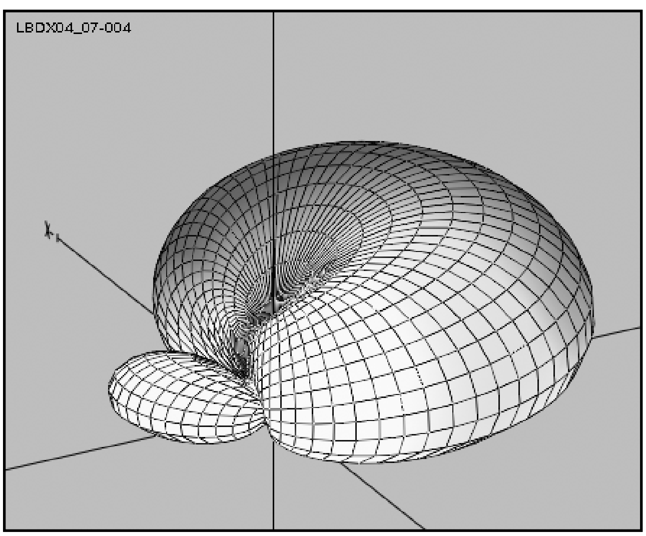


Fig 7-4 — A 3D radiation pattern of the 2-element end-fire array. (Plotted by *Antenna Model.*)

from the back. It is not necessary to hire a helicopter!

In the foregoing examples, elements were spaced λ/4 (90°) apart. The same analysis can be used with other spacings, such as λ/8.

Table 7-1 shows phase delay for different null angles and element spacings. Phase delay is given by

$$\psi = 180° - [\ S \times \cos\ (Na)\ ]$$

where
S = spacing in degrees
Na = null angle in degrees

1.7. Where Do We Put the Null?

We now understand how to move the null in an array, but haven't discussed the best position for the null. Since circumstances vary, there is no universal "best position." We always want to position the null to remove the maximum amount of *accumulated noise power* compared to the response at the desired signal elevation and azimuth angles. There are three distinct cases to consider:

- A location with desired signals arriving from a reasonably wide expanse of low noise, but with very high levels of noise covering a wide quadrant behind the array. An example of this would be a location in the suburbs of a high-noise city looking out over the very quiet ocean toward many DX contacts. Another case would be a noisy power distribution line going past a very quiet location. In this case the vast majority of noise (more than 15 dB higher than normal ambient in the forward direction) would always arrive from a well-defined relatively wide area, while desired signals arrive from a relatively low-noise wide forward area.
- The second case would be a quiet rural or somewhat quiet suburban location, with noise randomly distributed in all directions. This would be the case for most suburban and rural amateur stations, where widely distributed rearward local noise or point-source noise is not a major issue. This situation also applies where noise averages over time to be about the same from all directions and isn't greatly stronger

Table 7-1
Required phasing angle ϕ as a function of spacing and notch angle

Null Angle — *Element Spacing, Degrees*

Null Angle	*23*	*35*	*45*	*50*	*60*	*70*	*80*	*90*	*100*	*110*	*120*	*130*	*140*	*150*	*160*	*170*	*180*
0	158	145	135	130	120	110	100	90	80	70	60	50	40	30	20	10	0
5	158	145	135	130	120	110	100	90	80	70	60	50	41	31	21	11	1
10	158	146	136	131	121	111	101	91	82	72	62	52	42	32	22	13	3
15	158	146	137	132	122	112	103	93	83	74	64	54	45	35	25	16	6
20	159	147	138	133	124	114	105	95	86	77	67	58	48	39	30	20	11
25	160	148	139	135	126	117	107	98	89	80	71	62	53	44	35	26	17
30	161	150	141	137	128	119	111	102	93	85	76	67	59	50	41	33	24
35	162	151	143	139	131	123	114	106	98	90	82	74	65	57	49	41	33
40	163	153	146	142	134	126	119	111	103	96	88	80	73	65	57	50	42
45	164	155	148	145	138	131	123	116	109	102	95	88	81	74	67	60	53
50	166	158	151	148	141	135	129	122	116	109	103	96	90	84	77	71	64
55	167	160	154	151	146	140	134	128	123	117	111	105	100	94	88	82	77

(greatly would be 15 dB or more) from any single area.

- The final case is where a strong single-source noise profoundly dominates all other noise arriving at the site. This may be typical of amateurs living in an area where all noise comes from an electrical substation or from a somewhat distant group of arcs or noise sources concentrated in one narrow direction.

In the first case, we should compare response in the rearward direction (or any other exceptionally noisy area) to the desired signal direction. The signal must be from a point in the relatively quiet front area of the antenna and the noise from a "problem" area. Good performance means we need a very broad null in the direction of the strong noise rather than the typical requirement of high directivity. We can identify this situation by changing antenna direction. If the noise shows about the same F/S (front to side) and F/R (front-to-rear) as signals and is very clearly in one general direction (a F/R or F/S change on your "S" meter similar to that of regular signals), you should pay close attention to the *Directivity Merit Figure* (DMF), described below.

In the second case, noise averages to be within several dB from every direction. We need low average gain compared to point-gain at the specific elevation angle and azimuth of the desired signal. This translates to the need for high directivity.

In the final case, we are mostly concerned about maintaining a very deep null in one direction. Noise from one specific direction is hundreds of times stronger than normal band noise, ruining our DX. We need to totally remove it without removing the desired signal.

1.8. The Directivity Merit Figure (DMF)

In the 3rd edition of this book, I ranked receiving antennas by calculating the average front-to-back, which was calculated at 120°, 140°, 160°, 180°, 200°, 220° and 240° azimuth and between 10° to 80° in elevation. This gives 7 × 8 = 56 gain figures, which were then compared to the maximum forward gain of the antenna.

Keeping the same basic approach, I elaborated on the idea and now calculate the average gain in the entire back azimuth half of the antenna, from 90° to 270°, and over the entire elevation range from 2.5° to 87.5°. Doing all of this at 5° increments means we consider 37 × 18 = 666 gain values. The average rearward gain now is the average of 666 values (watch out, you cannot average dB figures directly and have to compensate for area). We can now define a figure of merit for the directivity (front response to back half-hemisphere) as being the difference between the forward gain at an optimum wave angle (for example, 20°) and the average rearward gain.

For the 2-element array we developed above (with 90°) spacing, this DMF is 13.11 dB. The DMF is the peak front lobe (at a specified elevation angle of 24°) gain versus the average back half-hemisphere gain, expressed in dB, in this case –9.74 dB.

This method of evaluating a receive antenna applies to a case where a dominant noise arrives from a relatively wide half-hemisphere. If the noise is evenly distributed in all directions (eg, in a very quiet location), the RDF ranking system discussed in Section 1.9 should be used.

Let's examine DMF further. What is the DMF (at 25°) of our vertical antenna? By definition the vertical is an omnidirectional antenna. Does that means the DMF is 0 dB? No, its peak lobe (in all horizontal directions) is at an elevation angle of approximately 25° (given soil with conductivity = 5 mS and $\varepsilon = 13$), but the antenna has good rejection at high elevation angles in all horizontal directions (**Fig 7-5**). The DMF of a single vertical ($\lambda/4$ long at 3.65 MHz) is 5.07 dB (calculated for a 25° wave angle).

A word of caution: The nonprofessional versions of *EZNEC* and other software programs calculate patterns at an "infinite" distance from the antenna. If your software does not allow you to set a pattern distance, it almost certainly will not accurately evaluate ground wave signals. This means that anything less than perfect ground causes vertically polarized, zero-degree responses to incorrectly appear as zero. If ground wave or directly propagated noise exceeds skywave noise levels, vertically polarized antennas (including Beverages) may thus appear significantly better than they actually are for locally noisy locations. Despite that, these models work quite well at any location where skywave exceeds ground wave noise level.

What does 13.11-dB DMF mean for our 2-element end-fire array and what does the 5.07 dB DMF mean for our single vertical? It tells us the 2-element array will, on average, deliver 8.04 dB better S/N ratio, provided the dominant noise is skywave in the rearward area and provided that the desired signal arrives in the center of the forward lobe peak (at a 25°

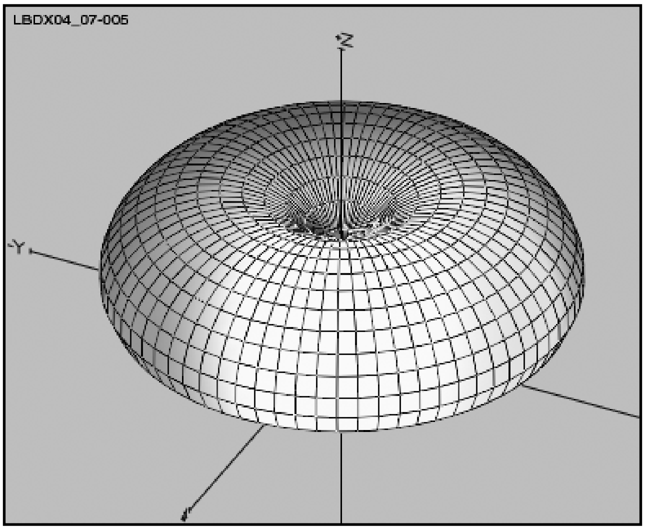


Fig 7-5 — A 3D radiation pattern of a single vertical. (Plotted by *Antenna Model*.)

elevation angle). The test for this condition is simple. If you see a change in noise level that is similar to the change in signal level when the array is reversed, you should be using DMF to evaluate antennas. You should set the rear measurement window to the direction of strong noise.

If you have a strong single-point noise and if that noise arrives in a deep notch in your receiving antenna pattern, the S/N improvement may be much greater than expected. If noise arrives predominantly from a higher antenna response area, the S/N improvement will be proportionally less. Another important thing to consider: Signals almost never arrive from a single angle or direction. A range of angles is involved, and a single-angle evaluation does not fully represent the real world. (If it did, signals would not have nearly as much QSB!)

Many noise sources vary in direction, arrival angle and polarization tilt. The same is true for desired signals. Because of this, we really only are considering "average" results over time. Averages are not foolproof under every condition. There is a story about a person who decided it was safe to wade across a river because the depth averaged only 4 feet. Well, he never made it to the other bank.

The *DMF* is the forward gain of the antenna at a chosen elevation angle (usually the elevation angle producing maximum gain) minus the average back quadrisphere gain. The back quadrisphere, also called the back half-hemisphere, is the area between 90° and 270° azimuth — provided the forward lobe is aiming at 0° azimuth — and 0° to 90° elevation.

Example: If the average back quadrisphere gain is –10 dB, and the maximum forward gain is 5 dB, the DMF is 15 dB. To calculate the average gain of an antenna in the back quadrisphere, we divide this whole area in a large number of small areas, areas measuring eg 1° by 1° degree. Now we calculate the gain figures for the center of each small area, and convert these gain figures into power ratios. Once this is done we need to normalize these as a function of the surface of the subject area. It is obvious that an area of 1° by 1° near the pole is substantially smaller than an area of 1° by 1° near the equator! Next you will need to calculate the average of all these normalized power ratio values (this is the tricky part) and finally convert them back to gain values. This gives you the average gain "in the back of the antenna." You really need a computer program to do that for you. For more detail see *Average Gain Computing.pdf* by Greg Ordy, W8WWV, which is available on this book's CD.

The easiest way to calculate the DMF of an antenna is to use *EZNEC* antenna modeling software in combination with W8WWV's *LBDXView* program (available on the CD that comes with this book). The *LBDXView* Help file gives you all the details. In short: You model the antenna using *EZNEC* and calculate the 3D pattern. From the Plot window, save the 3D plot (this is a *.PF3 file). Now run *LBDXView* and import the .PF3 file you just created using *EZNEC*. In the "Antenna Manager" window, right click on the subject pattern and select Properties. This will give you all relevant data for the subject antenna, including:

- Maximum antenna forward gain
- Radiation angle for maximum antenna gain
- Azimuth angle for maximum gain

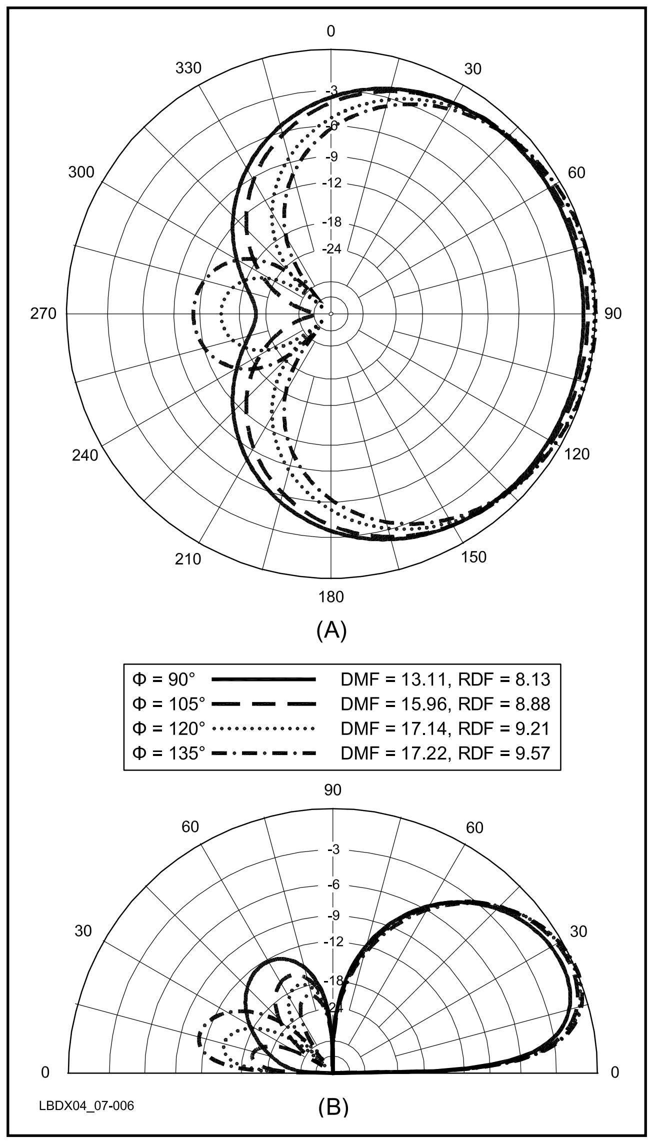


Fig 7-6 — At A, azimuth patterns and at B, elevation patterns for 2-element end-fire array with λ/4 spacing and various phase angles.

- Average back half-hemisphere gain
- Average hemisphere gain (required to calculated RDF, see section 1.9)
- RDF (= forward gain – average hemisphere gain)
- DMF (= forward gain – average half-hemisphere gain)

For example, using modeling file *CH11-2el-endfire-90-90phase.ez* from the CD: 3.37 dB gain at 24°; DMF = 13.11 dB; RDF = 8.13 dB.

The average hemisphere gain (4.89 dB) is exactly the same as what is calculated by *EZNEC*. However *EZNEC* does *not* calculate the average back half-hemisphere gain, while *LBDXView* does.

Fig 7-6 shows the radiation patterns for a 2-element end-fire array with λ/4 spacing and various phase angles. Note that the highest DMF is obtained for fairly high phasing angles. These phasing angles result in a good rejection at high elevation angles, although significant lobes at low angles are formed. On the average, though, higher phase angles achieve a better F/B, and thus a higher DMF. For higher phasing angles the forward lobe also gets narrower. Hence the Receiving Directivity Factor (RDF, discussed next) also becomes better for such angles.

1.9. The W8JI Receiving Directivity Factor (RDF)

W8JI has developed a similar measure to quantify receiving properties of antennas. While the Directivity Merit Factor (DMF) compares forward gain at the desired wave angle to the average gain in the rear half hemisphere, Tom's *Receiving Directivity Factor* (RDF) compares forward gain at a desired direction and elevation angle to average gain over the *entire hemisphere*. RDF includes all areas around and above the antenna, considering noise to be evenly distributed and aligned with the element polarization. Losses are factored out, and we find the directivity of the array. If noise, on average, is evenly distributed in *all directions* (including forward and side lobe areas) this method provides an accurate picture of receiving ability. (Keep in mind most antenna modeling programs used by amateurs calculate pattern at infinite distances and ignore ground wave response. RDF models, like DMF models, are not reliable when ground wave noise dominates skywave noise.)

For everything but an omnidirectional antenna, the RDF will be different from the DMF. You have to decide if your location has dominant skywave noise in the rearward area (DMF), or if skywave noise is evenly distributed on average (RDF). Do not compare RDF with DMF.

Calculating the RDF is very simple. First, carefully model the antenna with a *Windows* version of *EZNEC* by plotting the 3D pattern. The main *EZNEC* window shows average gain at the very bottom. You normally use this average gain figure (with all lossy antenna elements set to zero loss and in free space or over perfect ground) so that you can isolate actual ground and element losses from possible deficiencies in the model itself. You must fix any model deficiencies before proceeding. Once you've determined that the model itself is OK, you can resume using lossy elements and real ground to calculate the average gain figure.

Now, go to a two-dimensional elevation or azimuth pattern and select the desired elevation angle and/or azimuth of the desired signal with the gain cursor and note the gain. The difference between the overall average gain and gain at the desired direction and elevation angle is the RDF. The front lobe does not have to align with the desired signal. You can move the cursor around and look at the RDF for off-path signals.

You can also use the procedure described in Section 1.8 to calculate the RDF (and at the same time the DMF) using W8WWV's *LBDXView* software.

For our 2-element end-fire array (with 90° spacing and elements fed 130° out of phase), RDF = 9.21 dB (see Fig 7-6). The RDF of a single vertical is 5.09 dB. If noise or interference is somewhat evenly distributed, the end-fire array will show 9.21 – 5.09 = 4.12 dB *average* signal-to-noise ratio improvement over a single vertical (this is at 20° elevation in the main lobe peak).

Throughout this chapter we will assess the quality of the antennas by calculating both the DMF as well as the RDF, in addition to the –3 dB (half-power) beamwidth. Also remember a few dB of improvement in S/N, while meaningless on strong signals, can make a profound difference in readability of signals near the noise level.

1.10. RDF or DMF?

Both evaluation systems have their merits. If you're in a location that's always very quiet, with no specific noise or QRM sources from a particular direction, then RDF is most meaningful. The exception would be if you always had grossly dominant noise (or QRM) only from one direction. For a front-to-rear (F/R) selection to be valid, the dominant noise would have to be so strong as to consistently exceed distributed background noise by the null-depth ratio between an antenna selected by RDF compared to one selected by F/R.

For example, assume noise from a rearward quadrant was 20 dB higher than average noise from all other directions. Once the array had greater than 20 dB F/R ratio you could simply quit worrying about looking at F/R averages. Once the spot noise is down in the average noise, any additional depth is meaningless. At that point RDF takes over.

Another very important thing is when we work DX at local sunset or sunrise, when the rearward area is looking into a zone of poor propagation. W8JI wrote:

"At my very quiet QTH I see a 5-10 dB noise drop to the east at sunrise, and a 10-15 dB drop to the west near my sunset. This is because distant noise does not propagate in through the daylight areas. In this case, F/R is virtually meaningless and probably is 'over considered' even in RDF. Another thing is when we look into an area of good propagation, noise is enhanced from that direction also. The same mechanisms that enhance noise propagation enhance signals, so we had better consider beamwidth (which RDF does).

"RDF does not work well for local noise, but then nothing else will either. That's because *EZNEC* and other programs do patterns at "infinite" distance and do not show true response along the earth. They have no ground wave. If your modeling program does not have an input for distance, you can be sure it ignores ground wave. As such, there isn't an accurate model or method for those of us limited by local noise sources. RDF is exceptionally good for comparing similar antennas, such as a single Beverage to phased Beverages.

"I think there are very few skywave noise cases where anything but RDF applies. Even while it is far from perfect, it is the best overall method. If you have a case like those of us in the SE USA do, where a certain land mass has frequent thunderstorms (Florida and S Georgia), then it might pay to always be sure to have a deep null over that area. Even so, I would never pick the antenna exclusively based on the ratio of average gain in the null area to gain in the desired direction.

"At my QTH, antennas that have a poor 15 dB F/B hear just as well or better than antenna with huge 40 dB F/B ratios when the forward BW of the modest F/B arrays is narrower. The exception is summertime, when thunderstorms are off the rear. The worse thing you can do, over time, is go for extreme F/R at the expense of Half Power Beamwidth."

If you are in a less ideal situation, it seems to me that you first have to take care of the noise/QRM that is predominant from one direction. At my QTH that is the east/southeast. A good F/B (good DMF) is essential. Once that has been taken care of, further noise reduction can only be achieved by narrowing the forward lobe beamwidth, provided you have the room to do it, because broadside arrays that narrow the forward lobe require a great deal of space! In a nutshell: have a look at both the DMF and the RDF figures, and understand what they mean.

For most US stations, except those living deep down South, wintertime DX is not coming in from the same direction as the thunderstorm QRN (which is coming from the area between Texas and Caribbean). This means that an optimized front-to-rear performance for the receiving antenna is required. The DMF rating informs you best of this performance. In such a case close spaced end-fire fed Beverages are unbeatable (see Section 2.16.3).

Note that the RDF and DMF of a vertical (omnidirectional) antenna are identical. This is also true for a bidirectional antenna such as the broadside arrays described in the next section.

1.11. Broadside Arrays

In *end-fire* arrays, we considered the case of two elements where the signals produced by these two elements were combined after one of the signals was delayed by a proper amount to produce the wanted cardioid radiation pattern (see Fig 7-6). This is a unidirectional receiving array, where maximum reception happens in line with the line connecting the two antenna elements.

In a *broadside array,* maximum (bidirectional) reception occurs in a direction perpendicular to the line through the elements. What happens if we feed the elements in-phase and vary the spacing? Since both antennas are fed in phase, maximum radiation is perpendicular to a line bisecting the elements regardless of spacing. Areas in-line with the elements are the "side" of the array.

At zero spacing, there would be no spatial phase delay. Signals arriving from the sides would be in-phase at both elements. As spacing is increased, a point is reached where elements are separated λ/2. Signals arriving at 0° (ground wave) from the sides will be delayed λ/2 in space, exciting each element 180° out-of-phase. Since the feed system has no element-to-element phase shift, zero-degree elevation angle signals arrive at the common point out-of-phase and completely cancel.

As you increase the distance between elements, the bidirectional lobe becomes progressively narrower, and at the same time the vertical elevation angle at which the null occurs off the side is lifted off the ground, which is what we really want. However, beyond λ/2 spurious lobes begin to appear. These spurious lobes increase in strength as spacing is increased and may cause problems if they fall in noisy directions. It is obvious that with wider spacing more directivity is obtained through narrowing of lobes. If noise arrives in roughly similar amounts from all directions, narrower lobes (more directivity) will translate into higher S/N ratios and higher RDF numbers.

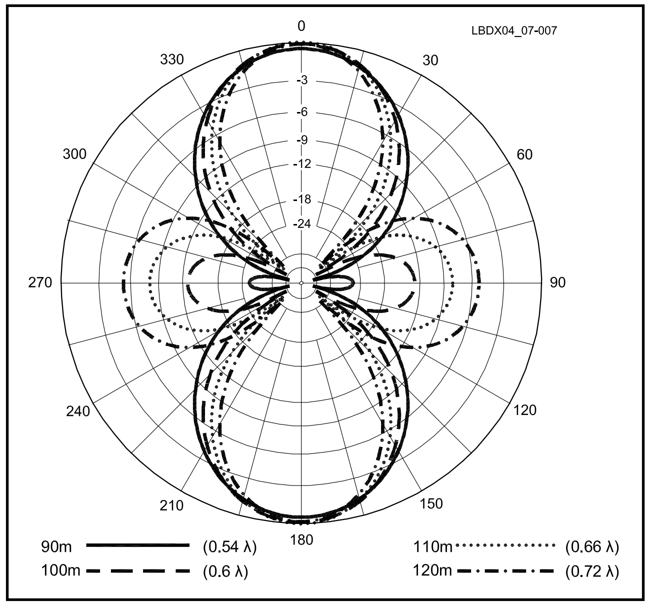


Fig 7-7 — Azimuth patterns for various spacings on 160 meters at 20° elevation for 2-element broadside vertical array (two verticals fed in phase).

Table 7-2
RDF For Various Spacings

Spacing	*90 m (0.54 λ)*	*100 m (0.60 λ)*	*110 m (0.66 λ)*	*120 m (0.72 λ)*
Gain	5.4 dB	5.8 dB	5.9 dB	5.8 dB
3-dB Angle	58°	52°	47°	43°
RDF	9.0 dB	9.5 dB	9.75 dB	9.7 dB

1.11.1. 2-Element Broadside Arrays

Fig 7-7 shows the bidirectional radiation pattern for various spacings on 160 meters. At first glance you might think that 90- or 110-meter spacings give better overall directivity than 110 or 120 meters, but this is not so. The RDF peaks for approximately 110-meter spacing (0.66 λ), as shown in **Table 7-2**. This is because at greater lateral spacings the sidelobes that appear become too important. Modeling files for these antennas are on the CD that comes with this book.

1.11.2. 4-Element Broadside Array

In a broadside array with more than two elements the current should taper away from the center in order to keep the effects of the so-called sidelobes under control. In a 4-element broadside array the outer elements should be fed with *half the current* of the center elements for best directivity. The pattern shown in **Fig 7-8** is for a 160-meter array with four elements spaced 110 meters apart. The 3-dB beamwidth is only 25° and the RDF is 12.8 dB. A modeling file is on the CD.

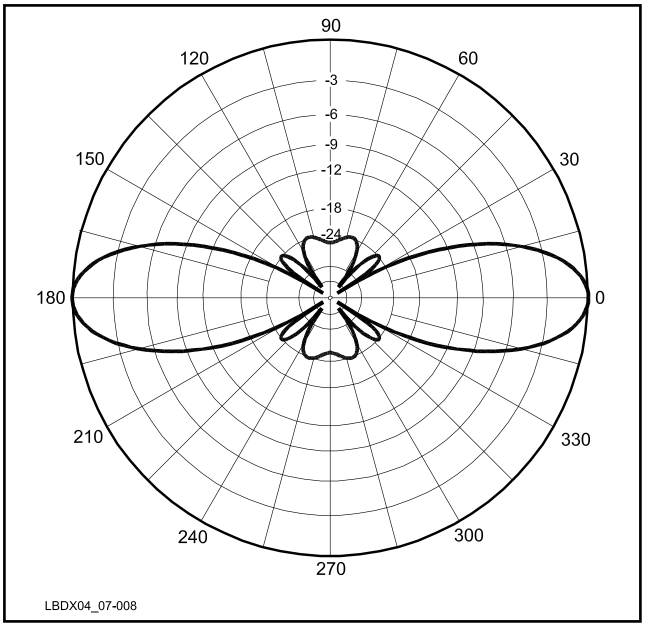


Fig 7-8 — Very narrow azimuth pattern for broadside array consisting of four verticals fed in phase.

1.12. End-Fire/Broadside Combination

End-fire arrays provide a unidirectional pattern, and you can move null angles by changing the phase delay and/or spacing between the elements. The forward lobe (cardioid) of such short end-fire arrays is rather wide (**Fig 7-9A**). A 2-element broadside array on the other hand is bidirectional, but can have a narrow lobe when the elements are spaced widely enough (Fig 7-9B). Combining these two systems can produce huge benefits, and is the most space efficient way to obtain very high directivity (Fig 7-9C). This is called *pattern multiplication* where you form a new antenna configuration by multiplying the patterns of two antennas. The feed systems for these arrays are described later in Sections 1.25.3 and 1.25.4.

Noise rejection is a three-dimensional problem. Wider spacing moves the nulls up off the ground and makes them more useful for distant QRM and noise. Wider spacing provides twice as many deep groundwave nulls and a noticeably narrower main lobe.

The elevation angle of the null center is given by

$$\alpha = \arc\cos \frac{\lambda / 2}{\text{Spacing}}$$

If we space the two "cells" a little wider than λ/2 we move the null up and form a cone reaching the ground, similar to the cone formed in end-fire arrays with larger phase lags. The same mechanism explained for end-fire arrays in Figs 7-2 and 7-3 applies to nulls in a broadside arrange-ment. Fig 7-9 illustrates that spacings slightly larger than λ/2 cause small sidelobes to form.

A spacing of λ/2 is optimum only when the dominant noise is coming in on ground wave, the source being at least a few "broadside spacings" away (which means that it must be outside the Fresnel zone) from the antenna, and directly off the side at exactly 90°. If the null is moved to higher elevation angles, patterns will change from the two right-angle nulls to

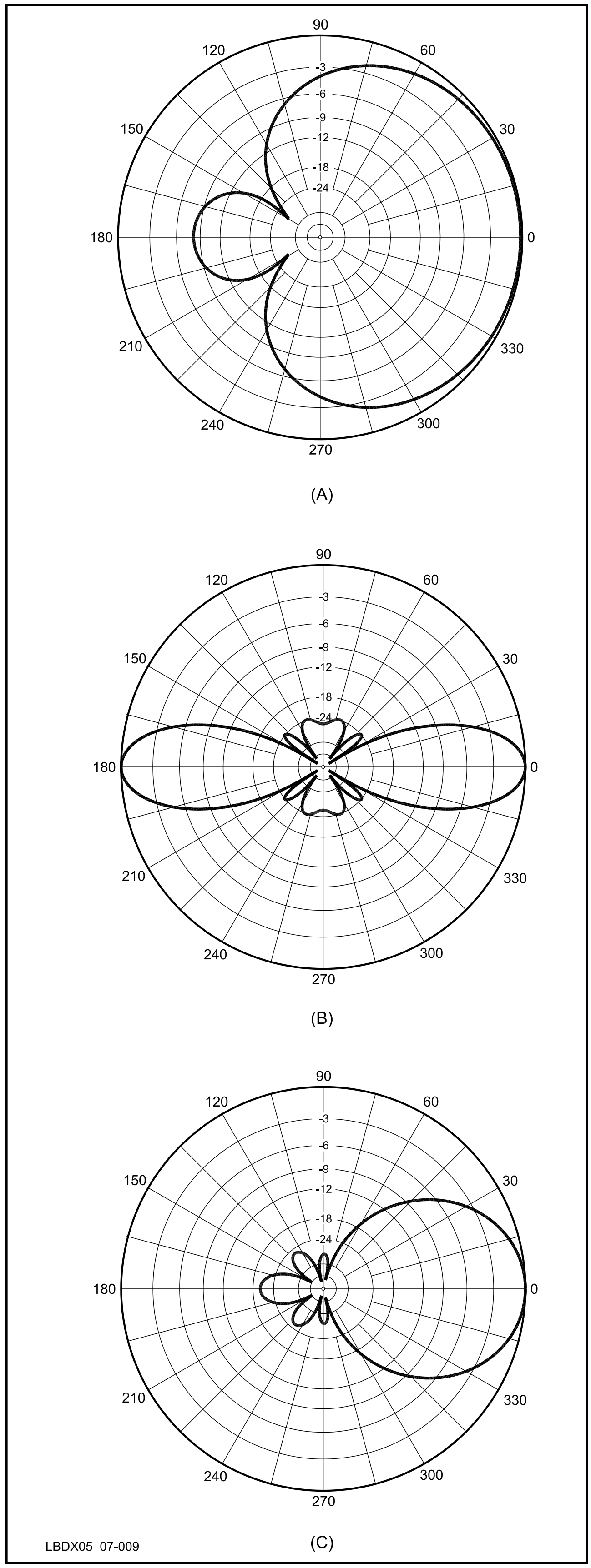


Fig 7-9 — Horizontal radiation patterns for a 2-element end-fire array (A) and a 2-element broadside array (B). The combination end-fire/broadside array (C) clearly combines the best characteristics of each individual component.

four nulls. A 25° to 30° side-null elevation is a good target angle if noise comes somewhat evenly from all directions. This is the angle that produces maximum directivity.

The plots in Fig 7-9 were generated using two end-fire groups (each 90° spacing, 105° phasing). These end-fire cells were then placed side-by-side with a separation of 198° (0.55 λ), placing additional side-nulls above the horizon and creating extra ground-level nulls. The side-null offset is arc cosine (180/198) = 25°.

DMF for this end-fire/broad-side array (with λ/4 spaced end-fire cells and 105° phasing, and 110-meter broadside spacing) is 23.45 dB, and the RDF is 12.4 dB. The 3-dB beamwidth is 57°. See the modeling file *ch7-4el-endfire-broadside-fig7-9.ez* on the CD that accompanies this book.

This is close to the optimum that can be achieved for a receiving pattern with four elements. End-fire/broadside combinations are used as building blocks for large multi-direction arrays (see Sections 1.29 and 1.30), or they can be expanded into super-directive receiving arrays. It was a large super-directive array of loop antennas that allowed W8JI to be the first station east of the western USA to work Japan in the presence of multiple extremely high-power pulse LORAN transmitters in the early 1970s. Pulse transmitters at 50 dB over S9 were taken to S2 with a custom blanker and a super-directive array.

1.13. Broadside Array Consisting of Four 2-Element End-Fire Cells

If you have lots of room (like 330 meters = 1000 feet) for broadside spacing on 160 meters, four end-fire cells will yield a razor-sharp pattern with a 3-dB beamwidth of approximately 25°. The design parameters for the array shown in **Fig 7-10** are:

- Broadside spacing between cells: 110 meters
- End-fire cells: ¼ λ spacing, 105° phase delay
- Current taper: 0.5, 1, 1, 0.5 (the outer element are fed with half the current of the inner elements)

Model: *ch7-8el-endf-broads-fig7-10.ez*

The performance is nothing short of phenomenal, with a 3-dB beamwidth of 24.5°, RDF = 16.06 dB and DMF = 29.35 dB. More practical design of such arrays is covered in Section 1.24. See also Section 1.35 (parasitic receiving arrays), where a similar array was designed around four cells, where each cell consists of a 2-element parasitic array (driven element and reflector).

1.13.1. The 3-dB Forward Beamwidth

A narrow forward beamwidth is great, provided you know exactly where signals are coming from (remember the crooked or skewed paths from Chapter 1). You will need more receiving arrays if each array has a very narrow beamwidth.

We probably should try to define *narrow*. A 2-element end-fire array has a 3-dB beamwidth of somewhere between 110° and 180°, depending on spacing and phasing angle. Three or four such arrays will work over the entire 360° azimuth without serious holes in coverage.

The end-fire/broadside combination just described has a 3-dB beamwidth of less than 60° (similar to a Yagi antenna). It would take eight arrays to cover the entire azimuth without significant pattern holes. The 3-dB beamwidth is actually a serious limit when you are looking at signals close to the noise floor because even one or two dB can make or break a contact.

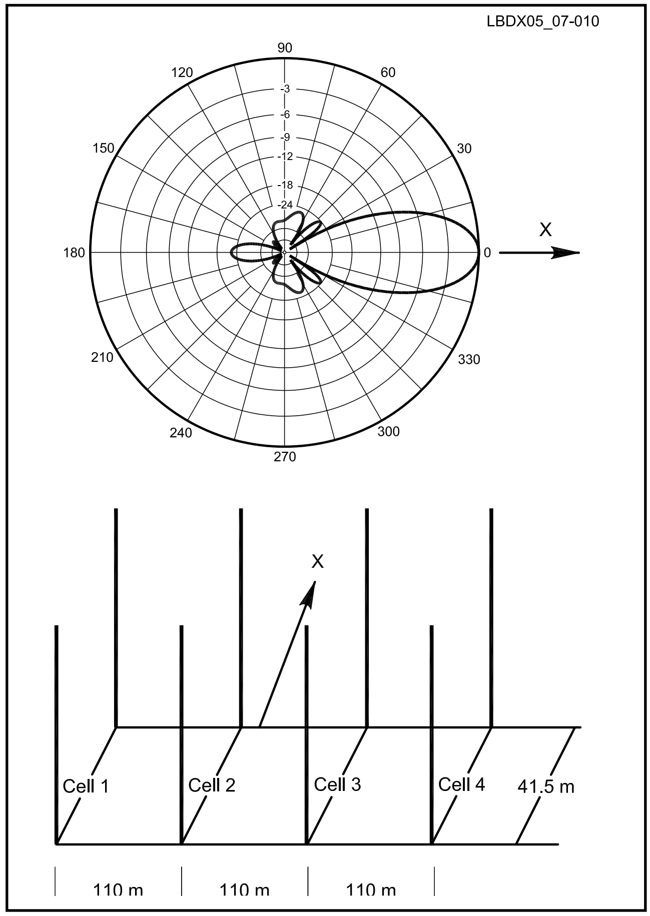


Fig 7-10 — Azimuth pattern (at 20° elevation angle) for a broadside array consisting of four 2-element end-fire array cells. See text for details.

Each person has to decide how much effort they want to put into making very weak signal contacts. When you hear a station consistently working weak DX you cannot hear, he is either in a much better location or has taken the time to build very directional antennas and a wide enough variety of them to cover every possible condition and direction.

There are two essential characteristics of a receiving antenna: the RDF (DMF if strong noise is in one defined area) and the 3-dB beamwidth. Gain is meaningless so long as external noise is several dB stronger than the receiving system's internal noise at the quietest time of operation using the narrowest selectivity.

1.14. Modeling Limitations and Tolerances

Models really are shortcuts, where everything is assumed to be simple and perfect. Sources are ideal in current, power and phase. The ground in the model is both flat and homogeneous, and there are no unwanted feed-line currents. The model often has no transmission lines, with no SWR and phase-shift errors that would plague real-world systems. Models work with numbers that are 32 digits long, or longer if we choose!

We frequently model receiving arrays as if they were transmit arrays. We talk about *feed currents*, although we do not feed the elements of such an array; in this case the voltages produced by the elements of the receive array are feeding one or more signal combiners, often after having gone through

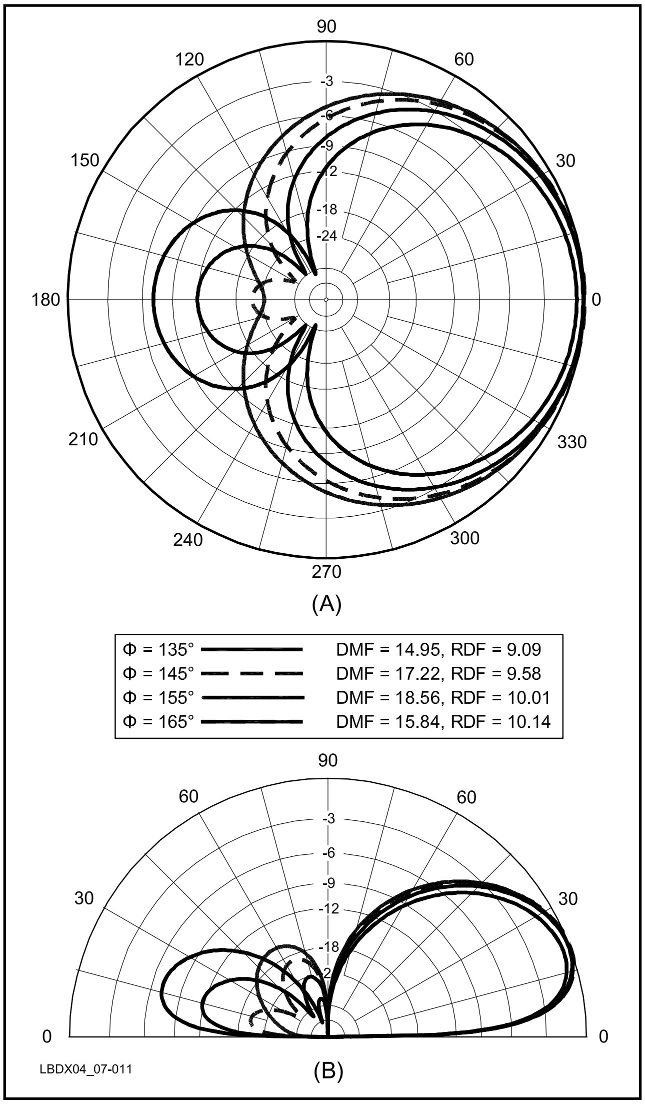


Fig 7-11 — Azimuth and elevation patterns for 2-element end-fire arrays with λ/8 spacing and various phase angles.

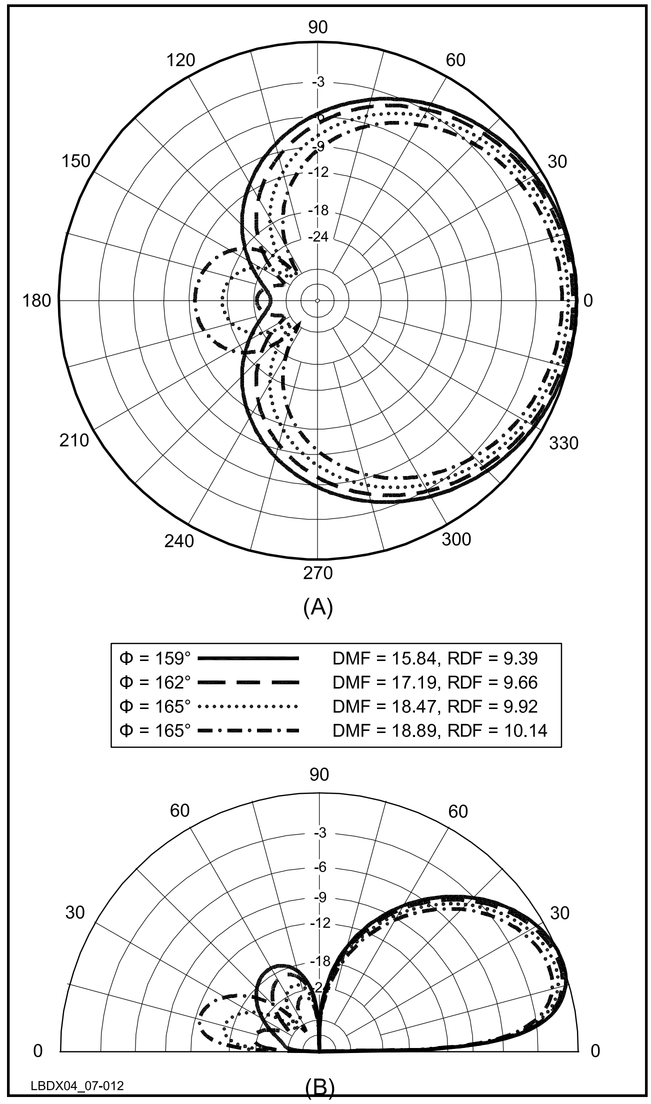


Fig 7-12 — Azimuth and elevation patterns for 2-element end-fire arrays with λ/16 spacing and various phase angles.

phasing lines (see Section 1.6).

Real-world antennas are often very different from models. In the real world everything is subject to tolerances. We should always keep this in mind as we examine antennas. If you model 2-element end-fire arrays with spacings closer than λ/4, you can obtain slightly better patterns. Spacings closer than λ/4 are often used for end-fire arrays, even in transmitting applications. (See **Fig 7-11**.) Modeling files are on the CD.

Note that the DMF (18.56 dB) peaks for a phasing angle of 155°, while RDF (10.14 dB) is still higher at a 165° phase angle. This is mainly due to the further narrowing of the forward lobe, which — in my opinion — overcompensates the worse F/R. **Fig 7-12** shows similar data for the λ/16-spaced end-fire array. The same remarks apply. Let's look at the impact of variations in feed-point phase and feed current amplitude.

If the phase angle is not exactly what we intend in a model, the only influence is a change in the position nulls. Figs 7-6, 7-11 and 7-12 show us that this is not a big problem, except for arrays with extremely close spacing or arrays requiring precise null locations.

Table 7-3
Modeling End-Fire Arrays at Various Spacings

λ/4 spacing, ψ =120°					
Ele 1, E(V)	1.0	1.0	1.0	1.0	1.0
Ele 2, E(V)	0.6	0.8	1.0	1.2	1.4
RDF (dB)	8.52	8.87	8.96	8.90	8.76
DMF (dB)	13.76	16.00	16.79	16.24	15.16
λ/8 spacing, ψ =155°					
Ele 1, E(V)	1.0	1.0	1.0	1.0	1.0
Ele 2, E(V)	0.6	0.8	1.0	1.2	1.4
RDF (dB)	8.40	9.47	9.82	9.57	9.08
DMF (dB)	11.54	15.90	18.76	16.59	13.89
λ/16 spacing, ψ =162°					
Ele 1, E(V)	1.0	1.0	1.0	1.0	1.0
Ele 2, E(V)	0.8	0.9	1.0	1.1	1.2
RDF (dB)	7.17	8.76	9.53	8.98	8.07
DMF (dB)	8.75	13.17	17.70	14.10	10.91

Ele 1 and Ele 2 are the relative voltage magnitudes delivered at the feed line combining points

But what if the *voltage magnitude* delivered by the feed lines coming from the two elements, and having possibly traveled through an extra length of feed line acting as a phase delay line, are not identical? Let us do a sensitivity analysis. Let's assume the feed line coming from element #1 delivers a signal with a magnitude of 1 μV at the signal combining point, while we vary the voltage magnitude delivered by the other antenna between 0.6 μV and 1.4 μV in the second element. This exercise was done on a λ/4 spaced end-fire array, a λ/8 spaced array and on a λ/16 spaced array. All calculations were done with 12-meter long elements over average ground (conductivity = 5 mS, ε = 13) on 1.83 MHz. The results are given in **Table 7-3**.

Note that the DMF is a much better indicator of what happens in the back quadrisphere. This is logical because RDF includes average gain (signal and noise pick-up) over the entire hemisphere, while DMF considers unwanted signals and noise arriving only from the rear quadrisphere. These are two very different cases. If you live where noise arrives from all directions with somewhat similar signal levels over time, use RDF. If noise comes largely from the rear of the antenna, use DMF for comparisons (see Section 1.10).

With normal somewhat-even noise (within several dB) from most or all directions, S/N will not change a great deal with slight differences in voltage magnitude of the signals delivered by the 2 elements. There is a clear rule for considering F/R (front-to-rear). If the ratio of arriving rearward noise to arriving forward noise approaches or exceeds the DMF, you will need to improve DMF to improve S/N ratio. If reversing the antenna produces a very clear noise increase of more than 10 or 20 dB, DMF should be the governing factor in array choice. If signals change a great deal but noise does not, use RDF and forget about extreme F/R ratios. F/R will not help a great deal.

Quarter-wave spaced arrays, even in situations where rearward noise is very high, can easily tolerate up to ±40% deviation in delivered voltage magnitude between both antennas, without showing much S/N deterioration. The acceptable feed current amplitude tolerance for a spacing of λ/8 (145° phase angle) is ±20% where rearward noise is a problem.

What if you space the elements even closer? Data for the λ/16 spaced array (162° phase angle) in Fig 7-12 shows that the range of voltage magnitudes (delivered at the combining point) that provides good directivity is much narrower. You now have only half the room for error. If you want to keep the DMF high (for concentrated rearward noise), you should keep the feed current tolerance to ±10%.

Close-spaced arrays are much less forgiving of errors than wider-spaced arrays. We can make an end-fire array quite small, and if properly designed and constructed, it will still perform well. There is no way, however, to reduce in size the width of the end-fire/broadside combination array described in Section 1.11. The large broadside spacing is required to reduce beamwidth and improve S/N. This is true for broadside arrays of all types, including Beverages.

1.15. Conclusions

In receiving arrays requiring deep nulls, control of the magnitude can be more important than control of phase angle. The exact phasing angle (in combination with the spacing between the elements) determines *where* the nulls are. If the magnitudes of the signals delivered by the two antennas are off a great deal, the nulls become less deep, and both RDF and DMF will suffer.

Close-spaced arrays require more accurate current and phase control (once more: no free lunch!).

1.16. Choosing Receiving Array Elements

It's very easy to look at pattern changes caused by current and phase variation in receiving arrays. We simply model the array in a modeling program's wire table, enter the correct element phase and current ratios in a source menu and insert proper loads. If we experiment with the phase and ratio of currents, we can observe changes in pattern as the elements depart from optimum phase and current ratios.

If our feed lines were terminated at both ends in resistors having an impedance equal to the line's characteristic impedance, we could simply adjust the line length to change element phase. In this ideal situation, where feed-line SWR is a perfect 1:1 ratio, phase shift is the same as electrical line length in degrees. Once we have a mismatch — that is, we have standing waves — the line no longer has a phase shift equal to line length, unless that line is an exact multiple of 90°. The higher the SWR becomes, the greater the phase error. Line lengths that are odd-multiples of λ/8 provide the worst SWR-related phase errors. Phase errors in each section of line will add, causing longer lines to have more accumulated error. Longer lines also become more frequency sensitive. A 3λ/4 line has more frequency/phase error than a λ/4 line.

Elements showing constant impedance over wide frequency ranges are very desirable, especially if mutual-coupling effects can be eliminated. Resistors have these qualities. They have a very wide SWR bandwidth and show no effects from mutual coupling at spacings of more than a few resistor lengths! Unfortunately, there isn't much useful electromagnetic radiation or reception associated with small resistors.

There is a solution to the lack of electromagnetic radiation and reception of a resistor: the resistor does not need to be the *entire* antenna. We can make the resistor a large part of the antenna, including just enough antenna area to receive useful amounts of signal. Broadband phasing systems are easily implemented in systems where feed-point impedance is stabilized through intentional loss mechanisms. If we make the losses large enough to swamp out or dilute mutual coupling and resonance effects, antenna feed-point impedance remains stable and predictable, even with close-spaced, very short elements.

1.16.1. Naturally Lossy Elements

You can use an antenna that has low radiation resistance and high loss resistance. A Beverage is just such a natural antenna. It has a radiation resistance of a few ohms and a loss resistance in the hundreds of ohms. The large antenna loss resistance caused by the nearby lossy earth below the antenna that dilutes or swamps out mutual coupling because the radiation resistance is a tiny fraction of the feed-point resistance. The bulk of feed-point resistance is due to losses. In addition, the termination resistor adds more loss and stabilizes impedance over wide frequency ranges (see Section 2.16). Nonresonant loops also meet these requirements (see Section 3).

1.16.2. Resistance-Swamped Elements

Short vertical elements have a very small radiation resistance, yet they are very sensitive. They are vertically polarized and the earth below the antenna does not try to cancel radiation. If you load a short vertical with significant resistance and then cancel the antenna's reactance, you can build an antenna with a wide SWR bandwidth. In addition, the high loss of the loading resistor swamps out mutual coupling effects.

Element Q is almost totally defined by the ratio of loading reactance to loading resistance. With 270 Ω of reactance and 75 Ω of resistance (see the W8JI element, Section 1.21.1), element Q is 270/75 =3.6, more than enough to cover the widest amateur band (and then some). Higher resistance reduces Q and increases array bandwidth, but it also decreases sensitivity. Lower values of lumped reactance reduce Q. Top-loading with a large capacitance hat increases radiation resistance and sensitivity, and also reduces reactance and Q. The combination of a top-loading hat and 75-Ω feed system results in an antenna with a relatively good bandwidth even if we require that the SWR be less than 1.2:1 (≥25 dB return loss) on the band edges (see Section 1.21).

Adding a resistance in series with R_{rad} (to obtain a total resistive part of 75 Ω) and canceling out the reactance of the short element by adding a coil takes care of the impedance aspect. With everything done properly, the feed line will see a very well matched load.

Having done this with the various elements of a receiving array does not, however, guarantee that each of the elements will deliver the same signal voltage to its feed line. It is clear that we need these voltages at the base of each of the short verticals to be identical (be careful, this situation is very different with transmit arrays!). These voltages greatly depend on the *height of the antenna element*. For short vertical antennas, this voltage varies with the square root of the R_{rad}. This means that in an array it is not sufficient that the elements exhibit the same driving impedance, but that the actual element lengths be physically as identical as possible to obtain proper directivity.

1.16.3. Active Antenna Elements

A third solution is to use active antennas as nonresonant elements. Each element consists of a fairly short vertical element (perhaps 3 meters high), with a semiconductor (FET) source-follower circuit.

The short antenna element of an active antenna looks like a very low resistive (R_{rad}) part and very high capacitive reactance part. The input impedance of the amplifier ideally should be a very high impedance. The output impedance of the amplifier however should be equal to the coax Z_0 such that the line sees a good return loss (a source follower circuit meets these requirements). As we can see in the equivalent circuit, the voltage at the antenna is dependent on the height of the vertical element (more specifically the square root of R_{rad}) and height needs to be controlled as well as possible.

The source follower presents constant impedance to the feed line, isolating reactive components from the element. Of course the circuit needs to fulfill a number of other criteria: It must withstand high RF level without damage from your own transmitting antennas, and it must not have intermodulation or harmonic distortion of signals while receiving. It also must withstand electrostatic fields and lighting discharges.

Tom, W8JI, used active elements in the 1980s, when he lived near Cleveland, Ohio. Those elements used very expensive 1.5-dB noise figure 28-V FETs operating at 400 mA quiescent current, not something the casual experimenter would have available. More recently W8JI has developed active vertical elements that are commercially available from DX Engineering (**www.dxengineering.com**).

1.17. Feeding the Elements of an Array

We need to combine signals coming from the elements of an array with carefully controlled amplitude and phase, but how do we achieve this? Transmission lines are ideal for

Table 7-4
Source Impedance for Various Vertical Arrays

	Ele 1	*Ele 2*	*Ele 3*	*Ele 4*
Single Vertical	2.0 – *j* 663 Ω			
2-Ele End-Fire array, λ/4 Spacing, 105°	0.67 – *j* 664 Ω	3.05 – *j* 662 Ω		
4-Ele End-Fire/Broadside, 90 m Spacing	–0.16 – *j* 664 Ω	3.17 – *j* 663 Ω	–0.16 – *j* 664 Ω	3.17 – *j* 663 Ω

Table 7-5
Resistor-Swamped Feed Impedance

	Ele 1	*Ele 2*	*Ele 3*	*Ele 4*
Single Vertical	74.0 + *j* 0.5 Ω			
2-Ele End-Fire Array, λ/4 Spacing, 105°	72.67 – *j* 0.5 Ω	75.05 – *j* 1.5 Ω		
4-Ele End-Fire/Broadside, 90-m Broadside Spacing	71.84 – *j* 0.5 Ω	75.17 + *j* 0.5 Ω	71.84 – *j* 0.5 Ω	75.17 + *j* 0.5 Ω

Table 7-6
SWR Values

	Ele 1	*Ele 2*	*Ele 3*	*Ele 4*
Single vertical	1.02			
2-ele end-fire array, λ/4 spacing, 105°	1.03	1.02		
4-ele end-fire/broadside, 90-m lateral spacing	1.04	1.01	1.04	1.01

moving radio frequency energy from antenna or array to the receiver, and at the same time they can act as delay lines. When improperly terminated, transmission lines become impedance transformers (see Chapter 5). A transmission line can be used as very flexible and easy to adjust phasing line if we are careful to follow good engineering practices.

So far we have not said much about impedances. I have modeled arrays on 160 meters using generic elements, each 12 meters long and 40 mm in diameter. Let's have a look at the 4-element array (end-fire/broadside) with end-fire cells spaced at $\lambda/4$ and end-fire cell phasing at 105°.

Looking at **Table 7-4** we see an almost constant imaginary part at $-j$ 663 Ω. The real part (the radiation resistance) is very low, and varies quite a bit from –0.16 to +3.17 Ω. While the R_{rad} of a single vertical is 2.0 Ω, mutual coupling between the various verticals causes the wide variation in feed-point resistance, even negative resistances, once elements are combined in an array. (See Chapter 11 for a fully detailed explanation.)

The solution popularized by W8JI is matching the short element impedance to the feed line (preferably 75 Ω) by inserting a resistor and inductor in series with the feed. Reactance is cancelled by the loading inductor's reactance. The resistor is selected so total loss resistance, including ground loss and loading inductor ESR (equivalent series resistance), is approximately 72 Ω. Adding enough loss resistance in series with an inductor of $+j$ 663.5 to equal 72 + j 663.5 Ω feed-point impedance is shown in **Table 7-5**. The resulting 75-Ω SWR is shown in **Table 7-6**.

The 75-Ω feed lines are terminated in very little radiation resistance, but have high loss resistances. The large loss resistance swamps out mutual-coupling effects, stabilizing feed impedances, regardless of element phasing and spacing (within reason).

With all feed lines operating at very low SWR it becomes very easy to design phasing systems. When the lines are matched, feed-line phase delay equals feed-line electrical length, for any length of feed line. Additionally, current and voltage along any length of line are equal, except for attenuation through normal feed-line loss. See Chapter 11.

1.18. Sensitivity Analysis

Section 1.14 examined current magnitude and phase errors and how they affect directivity. We estimated the tolerable magnitude of phase and current error in simple end-fire arrays. The next step is learning how to achieve our goals. Section 1.17 also described methods of making element impedance more constant and how to maintain very low feed-line SWR despite mutual coupling effects in end-fire arrays.

Section 1.16 clarified the important fact that phase delay equals feed-line length only if the line is flat (SWR = 1:1 or $Z_{load} = Z_0$) or a critical length (multiples of $\lambda/4$). We know that exact voltage magnitude is the most critical parameter for null depth, while phase controls the exact null placement.

Let's have a look at what happens in a feed line. Using the Voltage, Current, and Impedance Along Line module of the *Low Band Software*, we can calculate important parameters along the line in step sizes we desire.

Table 7-7 shows the voltage (magnitude vs a reference of 100 V at 0° phase angle) at the end of a coaxial feed line (with a loss of 0.2 dB per 30 meters on 160 meters) terminated in two different loads: 75 + j 7 Ω (SWR = 1.1:1, return loss approximately 26 dB) and 75 + j 14 Ω (SWR = 1.2:1 or 21 dB return loss).

- Column 1: Line length, in degrees
- Column 2: Voltage magnitude along the line for SWR = 1.1:1
- Column 3: Voltage angle along the line (in degrees) for SWR = 1.1:1
- Column 4: Voltage magnitude along the line for SWR = 1.2:1
- Column 5: Voltage angle along the line (in degrees) for SWR = 1.2:1

The table shows the relationship between voltage (magnitude and phase angle) at the end of the line and the length of the line.

In our antenna feed systems we cut our feed lines as if line length is equal to phase shift, and that assumes SWR is 1:1. When the SWR = 1:1 the phase angle tracks the cable length (both expressed in degrees).

In Table 7-7 you now see that both the phase and the magnitude of the voltage on the feed line are slightly off from what it should be in the ideal case. How much can you tolerate? There are a large number of variables involved: We already know that in a way, voltage magnitude is more important than the phase angle because a slight shift in phase angle merely moves the nulls around in the back of the antenna. Incorrect magnitude will make it impossible to obtain full cancellation — at whatever wave angle.

Doing a complete sensitivity analysis involving all parameters is quite complex and beyond the scope of this book. However, to give you an idea, an

Table 7-7
Voltage Along RG-6 Feed Line Terminated in Two Different Loads

	SWR = 1.1		SWR = 1.2	
Line Length (degrees)	*Voltage Magnitude (V)*	*Voltage Angle (degrees)*	*Voltage Magnitude (V)*	*Voltage Angle (degrees)*
0	100.0	0.0	100.0	0.0
10	101.6	9.8	103.0	9.4
20	103.0	19.2	105.6	18.3
30	104.0	28.5	107.3	26.8
40	104.5	37.7	108.1	35.2
50	104.5	46.8	107.9	43.5
60	104.0	56.0	106.7	51.9
70	103.0	65.3	104.6	60.6
80	101.7	74.9	101.9	69.8
90	100.1	84.7	98.9	79.5
100	98.6	94.9	95.8	89.8
110	97.3	105.3	93.2	100.8
120	96.4	116.1	91.3	112.4
130	95.9	127.0	90.5	124.3
140	96.1	137.9	90.9	136.3
150	96.8	148.8	92.4	148.0
160	98.0	159.5	94.9	159.3
170	99.5	169.9	97.9	169.9
180	101.1	179.9	101.1	179.9

SWR of 1.1:1 (return loss ~26 dB) on the phasing lines can introduce *voltage phase-angle* errors ranging from 0° to 6°, depending on the line length. Note that lines that are 90° long will always give a 90° phase shift between input voltage and output current, whatever the SWR.

The same 1.1:1 SWR can cause *voltage magnitude* errors of up to 5%, again depending on line length. With an SWR of 1.2:1, these deviations (in both voltage magnitude and phase) are almost doubled.

The amount of deviation we can tolerate for voltage phase angle and magnitude will greatly depend on the size of the array and the end use of the array. Wide-spaced arrays are much more tolerant than smaller spaced arrays. In large arrays, all these small deviations will do is slightly move the lobes (maxima and nulls) around in the back and make them a little deeper or shallower. The RDF and DMF will hardly change.

Without going into further detail, it is safe to state that you should try to design the array so that the SWR at the band edges is as low as possible, preferably less than 1.2:1.

Understand that we are analyzing the behavior of receiving arrays, where the antenna elements are the signal generators and where the final load of the feed system is the receiver. When dealing with transmitting antennas the antenna elements are the loads and we will analyze current rather than voltage (see Chapter 11).

In practice there are several things we can do to minimize phase and amplitude errors:

- Make as large an array as possible without compromising directivity. If you want to build a receiving Four Square and you have room for an λ/8 spaced or larger array, do not build a λ/16-sided array!
- Make sure you have a stable ground system that does not change with weather and season. Long and short-term impedance and loss stability with climatic changes is very important.
- Carefully measure and adjust SWR of the elements. Make sure the SWR at the band edges is low enough. Shoot for 1.2:1 SWR maximum at band edges, and use proper line-length planning if SWR is higher.
- Checking element feed impedance *regularly* is a must. If it is not stable over time, you will have to add radials (and/or increase the lumped constant resistor value).

1.19. What About Gain? (Signal Output)

I intentionally have not given a single gain figure so far, since I have insisted that gain (array output) is not an important issue for receiving antennas. That is also why I left the dBi figures out of all plots. Let us now analyze gain figures for arrays using 12-meter long loaded elements. See **Table 7-8** which includes both the dBi gain figures and the gain vs a single element with a gain of –12.25 dBi. These low gain figures are of course due to the resistive swamping of the input impedance, necessary to bring the value up to 75 Ω.

As we will see later, the output of the λ/4 wave spaced array is similar to the output of a reasonably long Beverage antenna. Under normal circumstances, with feed-line losses of less than a few dB you should not need a preamplifier — unless you are in a very quiet location and use narrow selectivity. With λ/8 spacing, a little amplification (see Section 6, covering preamplifiers) may be necessary, while λ/16 spacing requires at least 10 dB additional gain.

1.20. Feeding the End-Fire Array (Crossfire Feeding)

1.20.1. Calculating the Phasing Angle

Refer to Table 7-1 to determine the required phasing angle ψ. Let's assume we want to put our "null" at a wave angle of 30°. The spacing of our end-fire array is 20 meters, which is 43.5° on 160 meters (approximately λ/8) and 84° on 80 meters (a bit less than λ/4). From the table we can see that the required phasing angle on 160 meters is approximately 142° and for 80 meters approximately 107° (in both cases for a null angle of 30°). The angles can also be calculated as follows:

$$\psi = 180° - [S \times \cos(Na)]$$

where

S = spacing in degrees

Na = null angle in degrees

For 160 meters:

$$\psi = 180° - [43.5 \times \cos(30°)] = 142°$$

For 80 meters:

$$\psi = 180° - [84 \times \cos(30°)] = 107°$$

1.20.2. The Classic or Straightforward Phasing System

The circuit in **Fig 7-13A** shows what I would call old fashioned straightforward phasing: as you want the signal arriving from the back into element A to cancel the signal arriving into element B, you will have to delay the signal picked up by this antenna the same amount of time as it takes the signal to travel the extra distance from antenna A to antenna B (this is done by using a delay line). This is the feed system most currently used with transmitting arrays (see Chapter 11). The biggest disadvantage is that using this system there is no tracking between phase delay and frequency. In other words: if you design a receiving array with this feed system, it will only work well over a fairly narrow bandwidth and will certainly not cover two bands.

Table 7-8
Gain Figures For Arrays Using 12-Meter Long Loaded Elements

λ/4 Spacing, 12-Meter Long Elements, F= 1.89 MHz

Spacing	90°	105°	120°	135°
Gain, dBi	–9.2	–9.4	–9.7	–10.2
Gain vs1 el	3.1	2.8	2.5	2.0

λ/8 Spacing, 12-Meter Long Elements, F= 1.89 MHz

Spacing	135°	145°	155°
Gain	–12.6	–13.4	–14.5
Gain vs 1el	–0.3	–1.1	–2.2

λ/16 Spacing, 12-Meter Long Elements, F= 1.89 MHz

Spacing	155°	165°	175°
Gain	–17.1	–18.0	–19.1
Gain vs 1 el	–4.9	–5.7	–6.9

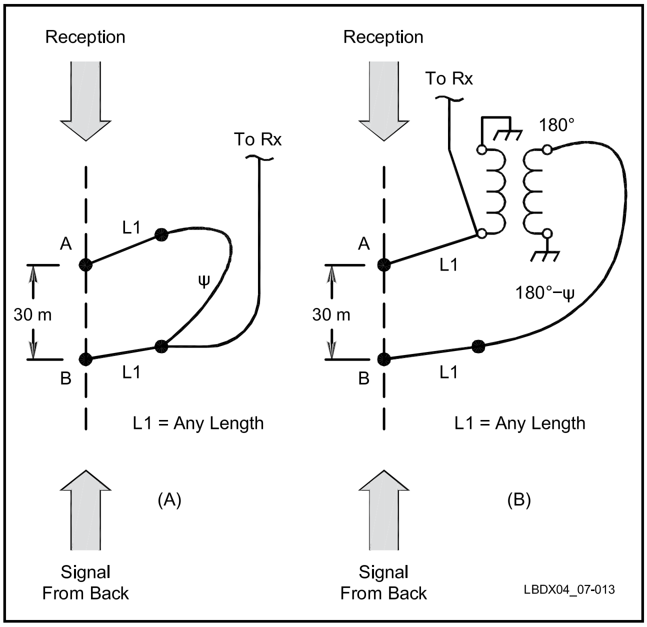


Fig 7-13 — Two ways of feeding the 2-element end-fire array. The system on the left is good for one frequency, while the system on the right can be used with the same length of phasing cable over a very wide range of frequencies (easily two bands).

1.20.3. Crossfire Phasing

Crossfire phasing is shown in Fig 7-13B. In this case the phasing line is inserted in the feed line to the back element, together with a phase inverting (180°) broadband transformer. Assume a 160 meter end-fire array having a spacing between the elements of 43.5°. If we want maximum rejection at 0° wave angle, the delay line in the crossfire feed system will have be 180° – 43.5° = 136.5°. Let us now analyze what that means on frequencies far away from our design frequency.

At 1.81 MHz

We calculated the required phasing angle ψ to be 142° (see Section 1.20.1).

- We install the (180° – 142° = 38°) long phasing line in the feed line to element B (leading current element).
- Delay angle to element A: L1° (L1 is the two equal-length pieces of coax going from the phasing/switch box to the two elements).
- Delay angle to element B: L1° + [180° – (180° – ψ)] = L1° + ψ°. The first 180° comes from the phase inverting transformer; the (180° – ψ) comes from the coaxial phasing line.

Note that element B has the leading feed current, as required. (In practice the 180° inversion can be on either element, a useful tool for building multiple element arrays.)

At 3.5 MHz

For 80 meters we calculated the required phasing angle ψ to be 107° (see Section 1.20.1).

- The same phasing line we calculated as being 38° long for 160 meters now has a length of: ψ = (38° × 3.5/1.81) = 73° long.
- The phasing angle on 80 meters is (180° – 73°) = 107° (180° from the inversion and 73° from the phasing line), exactly as required (see Section 1.20.1).

We can do the same analysis looking at what happens with signals coming from the front of the antenna. We will see that for all frequencies below λ/2 element spacing, signals from the front will never be out of phase! (W8JI has a detailed explanation at **www.w8ji.com/crossfire_phasing.html**.)

Using the same phasing-line length, the feed system maintains correct phase delay on both 160 and 80 meters. In actuality, phasing is correct from just above dc to the frequency where element spacing greatly exceeds 90°. This is a unique phasing system.

The use of the phase-inversion transformer, and the fact that we put the delay line in the back element instead of the front element, results in subtraction of phase, causing the phase delay system to fully track with changes in frequency.

The *phase-inversion transformer* is identical to a regular 1:1 transformer, where input and output are "cross-connected." See Figs 7-20 and 7-25 later in this chapter for details. This combining approach of connecting the coaxial cables together is not a good design, as will be explained in detail in Section 1.22.

At the feed point in Fig 7-13 (To Rx) the impedance is now 37.5 Ω. For a perfect match to our feed line we should provide a small wideband matching transformer. See Fig 7-26 later in this chapter for details.

The crossfire phasing system maintains the correct phase over a wide frequency range. Note, however, that the distance between the two elements should not be greater than λ/2 on the highest frequency to be used.

Tom, W8JI also used this system in transmitting arrays (Chapter 11, Section 3.4.4) as early as the mid-1970s. Despite trying to popularize this system over the years, it wasn't until the advent of the Internet that the word got out.

1.21. The Vertical Elements in Our Receiving Arrays

The issue is to make elements that have a very low Q over the entire band. Low Q means low SWR at band edges. Why do we want this? Because we use the feed lines to the elements as phasing lines, and to ensure proper phasing, the line SWR must be very low. Only a 1:1 SWR means line length in degrees is equal to phasing in degrees. Two parameters influence the variation of the impedance in an array as a function of frequency:

- The Q of the element itself.
- The amount of mutual coupling in the array. Large arrays with wide element spacing have much less mutual coupling than small, narrow-spaced arrays.

This means we can live with higher-Q elements in a wide-spaced array as compared to a narrow-spaced array. This constitutes the limiting factor in small arrays: Low-Q elements have very low output. As we make our arrays smaller the output will drop, as will bandwidth. Maybe most important of all is that small arrays are very critical to build and to adjust.

Let's examine a few types of short elements that can be used to build 80- and 160-meter vertical receiving arrays.

1.21.1. W8JI-Style Element (Umbrella Loading)

One of the nice things about this element is that you can easily build it to be resonant on 80 meters. This makes it a very attractive element for a two-band array, since you will not need to load to resonance on 80. See **Fig 7-14**. If you want to

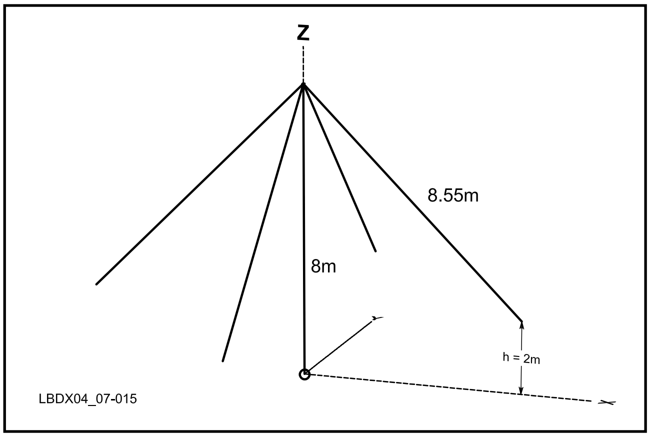


Fig 7-14 — W8JI-style element with slanted top-loading wires. This element is resonant on 80 meters.

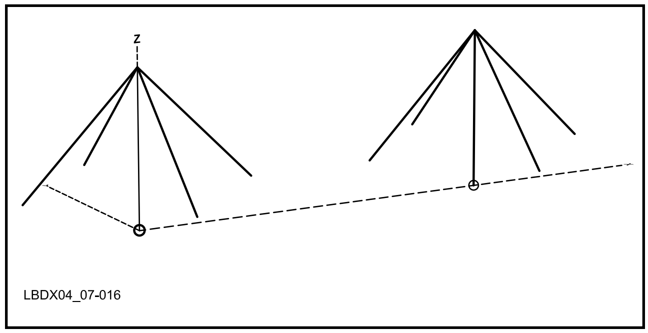


Fig 7-15 — Two W8JI-style elements are also evaluated in an end-fire array (see text for details).

model or build the element, first tune the element for 80 meters (3.65 MHz). Modeling was done with a 30-mm OD vertical tube and 2-mm loading wires. Exact length will depend on the diameter of the vertical tube and the size of the sloping wires.

The properties of the element on 160 meters are given in **Table 7-9**. The third column shows the impedance for the element loaded with 73.8 Ω in series with a coil of X_L = 277 Ω on 1.83 MHz (this represents a coil value of approximately 25 μH). Note that when modeling an element with a loading coil on various frequencies, if you specify the loading element(s) as Laplace Transforms in *NEC-2*, the impedance of the coils is tracked on various frequencies. Gain on 1.83 MHz over average ground is –16.7 dBi.

We should not forget that all of this is modeling. In real life we have tolerances and extra unknowns and unstable parameters involved.

After having checked the behavior of the W8JI-style top loaded element by itself, we will see how it behaves in an array. I modeled the element in a 2-element end-fire array with 20-meter spacing in **Fig 7-15**, using 105° phase shift on 80 and 140° phase shift on 160 meters. The results are listed in **Table 7-10**. The SWR levels at the band edges are very acceptable. On 80 meters we can tolerate a little more SWR, with a little deviation from ideal phase shift, as I explained and calculated in Section 1.14.

1.21.2. The K8BHZ Element

Brian Mattson, K8BHZ, developed another form of top-loaded element for his HEX-array in Section 1.29. He uses just two flat-top top-capacitance wires that run from one element to the next one in the circle containing the six elements. See **Fig 7-16**. The length of the top-hat wires is obviously half the spacing between the elements. As the wires are not 100% in-line (120° instead of 180°), horizontal radiation from these wires is not fully cancelled, but it is down just over 30 dB, which is acceptable.

The R_{rad} on 160 meters is a little higher than for the W8JI-element. As a consequence the output is a little higher (–14.9 dBi) on 160 meters. Logically, the bandwidth in the test configuration array (2-element end-fire) on 160 is a little bit less than with the W8JI element, as shown in **Table 7-11**.

Note that with this type of element the required loading coil will be approximately 39 μH, which is approximately 50% higher than in case of the W8JI element. This will have

Table 7-9
W8JI-Loading (Umbrella)

Freq (MHz)	*Zant*	*Zant-Loaded*	*SWR*
1.81	1.2 – *j* 282 Ω	75 – *j* 7.8 Ω	1.11
1.83	1.2 – *j* 277 Ω	75 Ω	1.00
1.85	1.2 – *j* 272 Ω	75 + *j* 8 Ω	1.11
1.87	1.2 – *j* 267 Ω	75 + *j* 15.7 Ω	1.23

Table 7-10
W8JI-Loading (Umbrella) in a 2-Ele End-Fire Array

Freq (MHz)	*Element*	*Ω*	*SWR*
1.81	Ele 1	73.0 – *j* 7.9	1.12
	Ele 2	76.86 – *j* 6.6	1.09
1.83	Ele 1	73.0 – *j* 0.5	1.03
	Ele 2	76.8 + *j* 0.7	1.03
1.85	Ele 1	73.1 + *j* 6.7	1.10
	Ele 2	76.8 + *j* 8.0	1.11
1.87	Ele 1	73.1 + *j* 13.9	1.21
	Ele 2	76.86 + *j* 15.2	1.22

Table 7-11
K8BHZ Elements in End-Fire Array

Freq (MHz)	*Zant*	*Zant + 73.5 Ω*	*SWR*
1.81	1.6 – *j* 448	75.0 – *j* 11.3	1.16
1.83	1.6 – *j* 442	75.0	1.00
1.85	1.7 – *j* 436	75.1 + *j* 11.1	1.16
1.87	1.7 – *j* 430	75.2 + *j* 22.1	1.34

Table 7-12
Base-Loaded Elements in End-Fire Array

Freq (MHz)	*Zant Ω*	*Zant + 75 Ω*	*SWR*
1.81	1.7 – *j* 700	75.0 – *j* 9	1.19
1.83	1.8 – *j* 691	75.0 + *j* 0	1.00
1.85	1.8 – *j* 682	75 + *j* 9	1.19
1.87	1.8 – *j* 674	75 + *j* 17	1.25

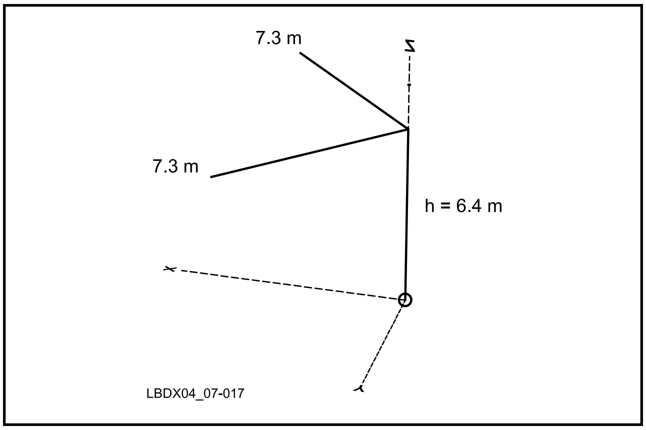


Fig 7-16 — K8BHZ-style element for the Stone-HEX array.

Fig 7-17 — Concrete base for the self-supporting elements used by the author. The concrete base goes down about 0.75 meter.

its consequences when we consider the array's operational bandwidth (see also Section 1.22).

1.21.3. Base-Loaded Elements

In some environments (in my front yard, for example) it is impossible to use top capacitance and guy wires. I'm lucky enough to be able to put up four self-supporting verticals, using slender tapering 11-meter long elements, which are mounted on bases set in concrete. See **Fig 7-17**.

The results in **Table 7-12** show that an array with these unloaded elements will have a slightly narrower bandwidth on 160 meters than an array made with W8JI-style elements, as the required loading coil is much larger (approximately 60 µH).Gain is –14.7 dBi.

1.21.4. Mechanical Construction of ON4UN's 11-Meter Long Elements

The bottom 6 meters of the 11-meter elements are made of a galvanized steel pipe measuring 60.3 mm OD with a 3 mm wall thickness (see **Fig 7-18**). Above that is a tapering element similar to a half element of a 20-meter Yagi (tapering from 35 mm OD to 20 mm OD). **Fig 7-19** shows the transition from a large-diameter tube to a much smaller one. Doughnut-like adapters are used, made of short lengths of aluminum tubing. Two stainless-steel bolts are driven through the element to secure everything.

Fig 7-18 — Roger, ON6WU, working on one of the self-supporting elements of the array at ON4UN's QTH.

1.22. Improved Feed System

(This section is based on contributions by Robye Lahlum, W1MK.)

In the previous editions of this book we described various types of small vertical element receiving arrays. Well known is the 2-element end-fire array as shown in Fig 7-13. The required phase shifts are obtained by using coaxial cable of a given length. Each element of the array consists of a "short" vertical. Each vertical, at the base feed point has an external resistor R and an inductor L, connected in such a way that the impedance looking toward the antenna is 75 Ω at resonance (see Sections 1.16 through 1.21). This means that, at the antenna, the coaxial feed line is terminated in exactly 75 Ω (over a relatively narrow bandwidth). The other end of the coax is connected to a combining network.

These receiving arrays used a *conventional combiner*

Fig 7-19 — Transition of the 60.3-mm OD steel pipe to the 35-mm OD aluminum element.

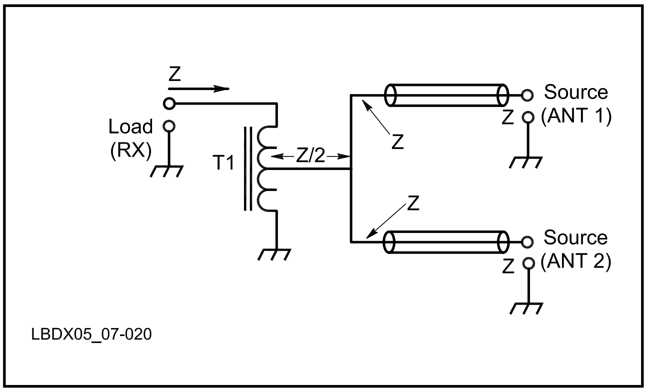


Fig 7-20 — Simple parallel combiner.

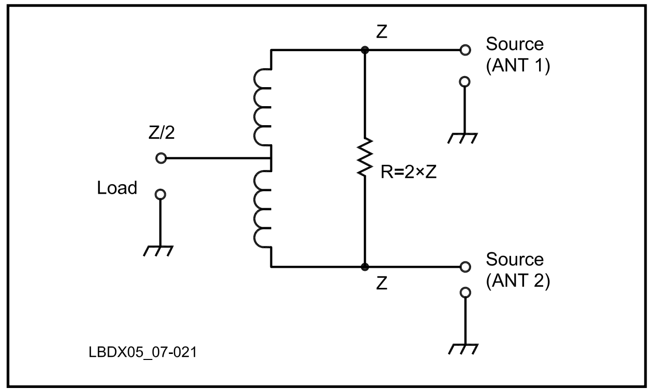


Fig 7-21 — Simplified schematic diagram of the 0° hybrid combiner.

consisting of a simple parallel connection as shown in **Fig 7-20**. In the case of the 2-element end-fire array, looking from each of the antenna elements (called "source") toward the parallel connection, we see a load impedance of 37.5 Ω (assuming T1 is a 2:1 impedance transformer and the receiver input impedance provides a correct match) in parallel with 75 Ω. This means that each of the feed lines is terminated in a load that creates an SWR of 3:1 (which equals 6 dB return loss). In addition to the 6 dB return loss there is 3 dB of insertion loss for the power at Ant 1 or Ant 2 going into the load port over that of a perfect combiner. For a perfect combiner all power from Ant 1 or Ant 2 will appear at the load port.

The family of arrays we consider here are using lengths of coaxial cable to obtain a desired phase shift. We know that, for the phase shift to be exactly the same as the cable length (both expressed in electrical degrees) the line must have a 1:1 SWR (see Section 1.18). For a receiving array, the generators are the antenna elements, and the final load is the receiver (when transmitting, the antenna is the load!). It all boils down to the point that, in order for the amplitude and phase shift through the coaxial phasing line(s) to be correct, the *sum of the return losses* looking in each direction should be in the 25 dB or greater range; in other words the SWR in both directions should be not greater than approximately 1.1:1 if we want to achieve the directivity we calculate with our antenna model.

The data collected in Tables 7-10, 7-11 and 7-12 (different kinds of short loaded 160 meter verticals) tell us that the 25 dB return loss bandwidth (1:1 SWR bandwidth) is quite low, on the order of 20-40 kHz. In practice it is even lower since the resonant frequency moves with temperature and other physical changes. Even if all the antennas track each other with these changes in resonant frequency, looking toward the load (the receiver) the coax is not at all "properly" terminated. It is terminated in an impedance that (in our example of a very simple 2-element array) causes a 3:1 SWR (merely 6 dB return loss). This means that the average return loss (looking at both directions in the cable) is certainly much less than the 25 dB we proposed.

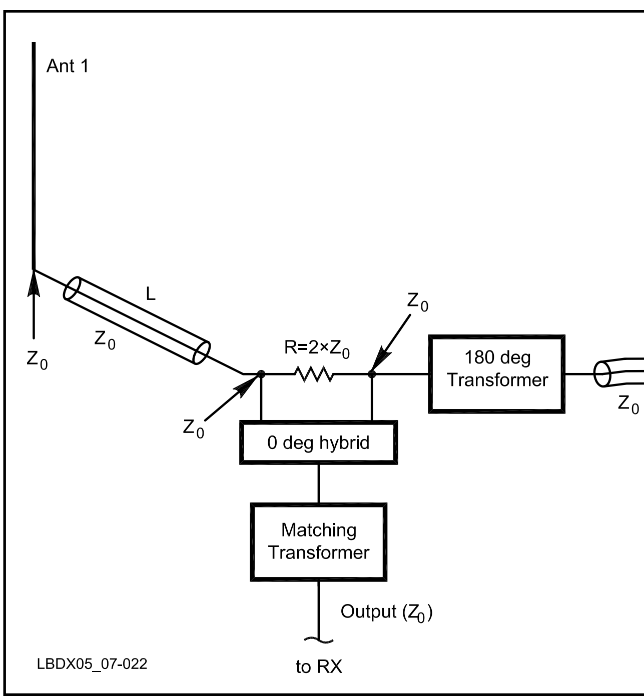


Fig 7-22 — Block diagram showing the different parts that make the feed system for the 2-element end-fire array. For simplicity, the direction switching is omitted. A complete diagram can be found in Fig 7-27.

In order to increase the bandwidth of the small receiving array, W1MK suggests using an improved combiner. This improved combiner has a broadband return loss of greater than 20 dB, and that over a very large bandwidth, from 1 MHz to 10 MHz. The coax that produces the desired phase shift will now be terminated in a good return loss over a much greater bandwidth.

This improved combiner is shown in **Fig 7-21** and consists of a *0° hybrid* of which an early description is given in Ref 1268. The circuit can be designed to present a constant loading impedance for both input ports, which ensures a return loss over a large bandwidth of greater than 20 dB in the coaxial lines feeding the combiner. This ensures a much improved array directivity over a much larger bandwidth. An important asset of this device is its isolation between the two input ports. The isolation is so good that the impedance looking in at one of the antenna ports is for all practical purposes independent of the impedance at the other antenna port. The conventional parallel combiner has very little isolation between the antenna ports.

This type of combiner/splitter is now also commercially available from DX Engineering (RSC-2), after a design by Tom, W8JI. Tom also explains the use of this type of splitter/combiner on his Web site (**www.w8ji.com/combiner_and_splitters.htm**).

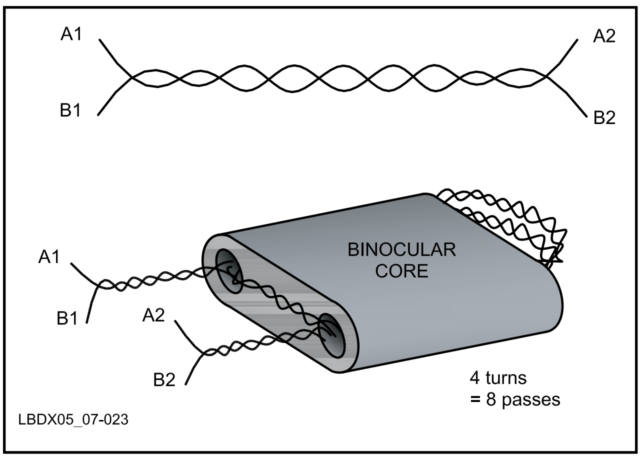


Fig 7-23 — This schematic shows how both the hybrid combiner network and the phase inverter are made. Only the connections of the wire ends (A1, A2, B1 and B2) are different. See Fig 7-25 and Fig 7-26. The 73 material binocular cores are Fair-Rite model # 2873000202, equivalent to Amidon part number BN-73-202 and CWS Bytemark type B-202-73.

One of the issues with the conventional combiner was that temperature changes produce a change in resonant frequency. The 0° hybrid combiner does not suffer from this problem as long as all the antennas track each other with temperature. Each antenna looks into a 75 Ω impedance, so we can calculate the effect of the antennas "not tracking" each other. For example a 1.8 MHz vertical using a 30 µH loading coil will suffer a 3.5° error with a 25 kHz movement in resonant frequency relative to the other antennas. As the value of the loading coil increases the phase error increases, for example an L of 40 µH produces a 4.5° phase error for the same 25 kHz shift.

While such a circuit, used as a signal (power) *splitter*, always causes a loss of just over 3 dB (half of the power goes to load 1 and half to load 2), when using it as a *combiner* there is *no loss*, provided we combine signals that are in phase.

Fig 7-22 shows how the 0° hybrid is used to combine the signals coming from the two antennas that are part of the end-fire receive array. Note that the output impedance of the combiner hybrid is half of the impedance of the combiner's inputs. Therefore the combiner is usually followed by a 2:1 impedance transformer.

Construction details of three components (0° hybrid, phase inverter and impedance matching transformer) are shown in **Figs 7-23** through **7-26**. All three components can be wound on a binocular core made of type 73 material, as manufactured by Fair-Rite (model # 2873000202). These cores are marketed by Amidon as part number BN-73-202 and CWS Bytemark as part number B-202-73. These cores saturate at about 25 W of power. The 0° hybrid and the 1:1 phase inverter are wound with 4 turns (8 passes) of the twisted wire, as shown in Fig 7-23.

Figs 7-24 and 7-25 show how a hybrid combiner and a 1:1 (impedance ratio) phase inverter are connected into the circuit. The matching transformer serves to transform the output impedance of the hybrid combiner (in the case of a 2-element end-fire this impedance is $Z_0/2$) to the impedance of the feed line going to the receiver. In this case it is likely to be a 2:1 impedance ratio transformer.

This transformer requires a $\sqrt{2}$ = 1.4 turns ratio. Although one can use an "autotransformer" (one winding with a

Fig 7-24 — Schematic showing how the wire ends A1, A2, B1 and B2 are connected to make a 0° hybrid combiner.

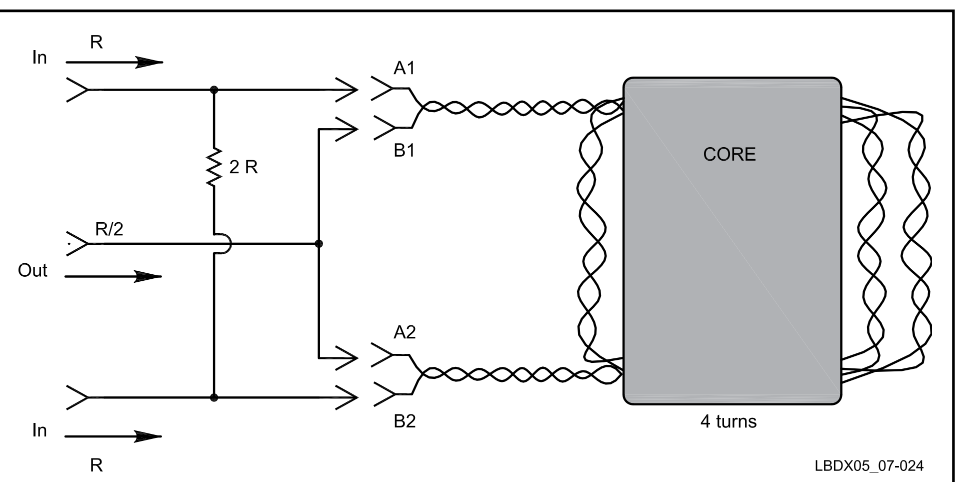
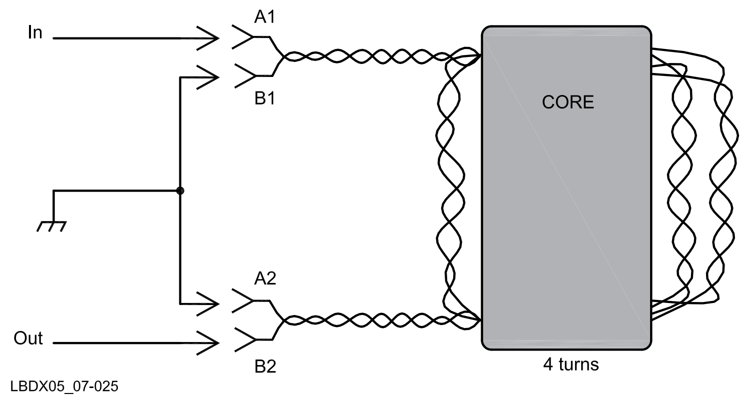


Fig 7-25 — Connected as shown in this schematic, the unit serves as a phase inverter (180°, 1:1 impedance ratio transformer). For load impedances between 100 and 33 Ω, the measured insertion loss is less than 0.1 dB.



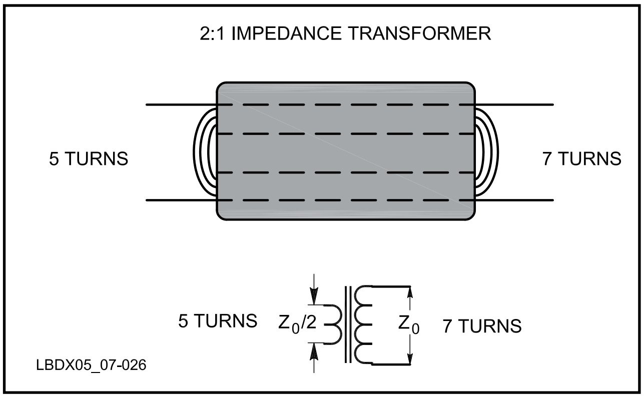


Fig 7-26 — The transformer with a 2:1 impedance ratio is wound with 7 turns (14 passes) for the 75 Ω winding and 5 turns for the 37.5 Ω winding.

tap), it is always better to use a transformer with galvanically separate windings, so that the transformer also acts as a *braid breaker* (see Sections 2.7.2.8 and 2.7.2.10). This can help reduce ingress of common-mode signals coming off the feed line. When using such a transformer, do *not* connect the shield of the feed line to the antenna ground (the ground to which the "cold" side of the low impedance winding is connected). Such a transformer can be wound on the same binocular core with 7 turns for the high-impedance winding (75 Ω) and 5 turns for the low impedance winding (37.5 Ω). See Fig 7-26.

In other arrays described in this chapter, this transformer may require different impedance ratios (hence different turns ratios), usually either a 2:1 or a 4:1 impedance transformation ratio. This will be dealt with when each of these arrays is described. The 4:1 transformer is shown later in Fig 7-30.

In order to illustrate the bandwidth improvement of this improved combiner, W1MK modeled both combiners when used with a small λ/8 wave Four Square vertical array on 160 meters (phasing step = 86°). Each antenna needed 28 μH of inductance to bring it to resonance. The antenna elements used were 13.2 meters high using four 6-meter long sloping tip wires. *EZNEC* v 5.0 was used, allowing the matching network to be included as part of the model. Using the conventional combiner the 20 dB F/B bandwidth was 40 kHz. With the improved combiner, the 20 dB F/B bandwidth increased by a factor of 5, to 200 kHz.

Considering the benefits of the use of the 0° hybrid combiners, these types of combiners have been used throughout this chapter when combining signals from different array elements.

1.23. Choosing Switching Relays for Receiving Antennas

In the past we used multiple pole general purpose relays, such as those commonly used in switching units for transmitting antennas. This has proven not to be a good choice for switching feed lines of receiving antennas and arrays.

General purpose relays with open contacts use the principle of "self cleaning" contacts. The contact material (unless pure gold) inevitably oxidizes from exposure to air, which in general means that the contact resistance increases. If such a relay switches "heavy" loads (as is the case in transmit antenna systems) the high current will literally burn away the insulating oxide layer. Hence the name "self cleaning" contacts.

When we switch feed lines going to receive-only systems, there is no current high enough to do the cleaning job, and after a couple of years (depending on how polluted the environment is), contacts will go bad. They will show high resistance or even open.

There are two remedies:

1) In addition to the RF signal, route dc voltage through the contacts (decoupled with a capacitor and RF choke). This dc voltage and the load should be designed to deliver enough power to do the cleaning job. This is the brute force approach, which I do not like and do not apply because it is hard to determine the amount of current needed to clean the contacts. In addition, decoupling components may degrade the isolation between antennas if not engineered correctly.

2) Use relays with hermetically sealed contacts. You can use vacuum relays (expensive) or reed relays (cheap). Reed relays with a single-make contact can be bought inexpensively, and suitable relays are available from distributors such as Digi-Key and Mouser.

Get a bag of reed relays, and get rid of the bad contact problems! SPST reed relays are the most common, and a good choice (lower capacitive coupling), even if you need twice as many relays. If you live in an area with lots of thunderstorm activity, make sure you have sealed gas-discharge tubes installed at the base of the antenna. These tubes are made by Bourns. Use the lowest voltage type, which is 90 V. This is the Bourns # 2035-09-B, which is available from Mouser as their #652-2035-09-B (cost less than $2). Now you are all set! Brian, K8BHZ uses such reed relays and gas discharge elements

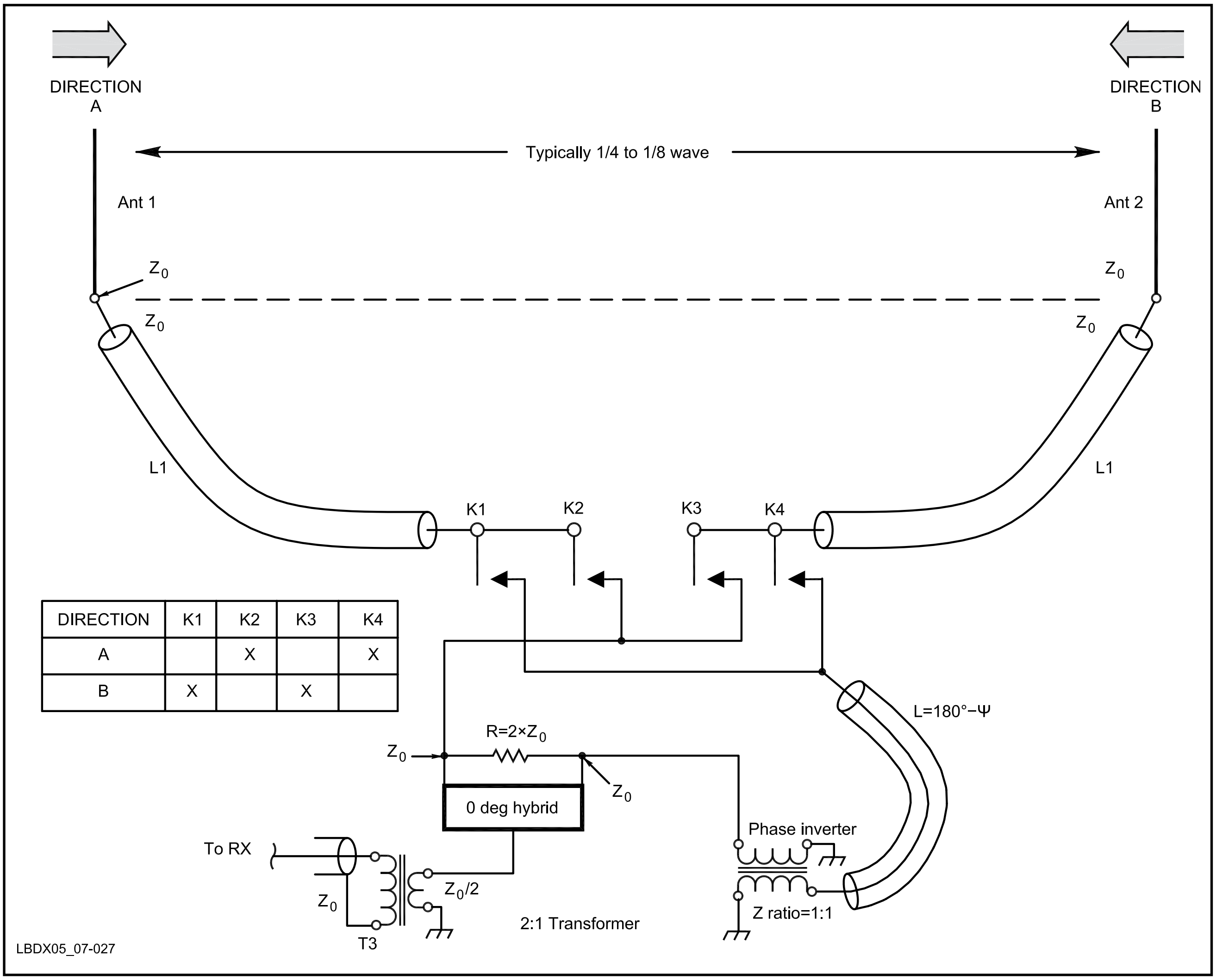


Fig 7-27 — End-fire 2-element array using the 0° hybrid combiner instead of a simple parallel connection in the feed circuit. Compare the schematic with Fig 7-13. The circuit uses three cores: one for the 0° hybrid combiner, one for the 180° phase inverter and one for the 2:1 autotransformer.

and has never had troubles. (See Fig 7-52 later in this chapter where you can see the array of reed relays for switching the feed lines to his Six Circle array.) Since I have replaced all my old fashioned power relays with small relays using hermetically sealed contacts, all my bad contact problems are gone.

1.24. Array Configurations

In the foregoing sections I worked with the example of the 2-element end-fire array to explain the mechanism by which directivity is achieved and to show how we feed the array using the crossfire principle.

Using the 0° hybrid combiner (see Section 1.22) for feeding the 2-element end-fire array is shown in **Fig 7-27.** You can use four small SPST reed relays to do the job (see Section 1.23). In Fig 7-27 all four relays are shown in de-energized state. The winding data for the 0° hybrid combiner, the phase inverter and the matching transformer (T3) are given in Section 1.22.

There are more configurations than the 2-element end-fire and the 4-element end-fire/broadside arrays we have been studying. I examined these because they are the simplest and most forgiving, and I used them as a step-by-step exercise to show all the aspects involved.

Other arrays developed according to the guidelines and procedures explained above are:

- *Broadside* arrays and *broadside/end-fire* combination arrays
- The *Four Square* array: Four elements set up in a square, yielding an array that is switchable in four different directions (receiving directions along the diagonals going through the corners of the square).
- The *Hex (Six Circle) array*: Six elements spaced 60° in a circle, yielding an array that is switchable in six different directions.
- The *Eight Circle* array: Eight elements spaced 45° in a circle, yielding an array that is switchable in eight different directions.

We will now analyze these configurations one-by-one and evaluate various versions. As with the 2-element end-fire, it is possible to develop these arrays on different scales. As is the case with the end-fire array, smaller arrays can sometimes have slightly better theoretical performance, but they are more difficult (less forgiving) to build, have a narrower bandwidth and have substantially less output.

Many other configurations are possible. Once you understand the design procedure, you can design your own array.

1.25. Broadside Arrays and Broadside/End-Fire Arrays

In Section 1-11 we reviewed the 2- and the 4-element broadside array, which develop narrow patterns. How do we feed these arrays?

1.25.1. Feeding the Two-Element Broadside Array

This bidirectional array has its elements spaced approximately ⅝ wave (110 meters on Top Band). It has a narrow 3-dB forward angle of 47° and an RDF of 9.7 dB (see also Fig 7-7). It is obvious that the same setup can be used with less spacing (see Section 1.11.1), in which case the forward lobe becomes wider.

Feeding is extremely simple: Run two 75-Ω feed lines of identical length (if the feed impedance of the elements is 75 Ω) to a 0° hybrid combiner (see Section 1.22), as shown in Fig 7-23.

The output of the combiner (Z = 37.5 Ω) goes to a 2:1 transformer to get the impedance back up to 75 Ω as shown in **Fig 7-28**. The construction of the 2:1 impedance matching transformer is given in Fig 7-26.

1.25.2. Feeding the Four-Element Broadside Array

The feed principle is the same (**Fig 7-29**). However, in this array, with a 3-dB beamwidth of only 25° and an RDF of 12.4 dB, we get the cleanest directivity pattern if we reduce the received voltages coming from the two outer elements to half of the signal voltages coming from the two inner elements when the signals are combined at the Receiver port. All antenna elements have the same signal voltage magnitudes, hence the 6 dB pad. The value of the resistors R1 and R2 in the 6 dB attenuator can be calculated using one of the many Pi (π) attenuator calculators available on Internet. For a system impedance of 37.5 Ω, R1 = 28 Ω and R2 = 110 Ω.

Sometimes we see that the outer elements are fed via an extra feed line length of 1 λ to save on cable and obtain some extra attenuation (part of the 6 dB). In such a case the four elements are fed in-phase only on the frequency where the extra cable is exactly 1 λ long, but as we move from this frequency the phase change will be much faster in the longer cables to the outer elements than in the cables going to the center elements.

We use 0° hybrid combiners to accept the inputs from the four vertical elements (see Section 1.22). Combining the outputs of these two combiners is done in another such combiner. This one has an input impedance of 37.5 Ω (assuming the array is designed around a system and feed line impedance of 75 Ω), and an output impedance of 37.5/2 = 18.75 Ω. Hence the 4:1 step-up matching transformer used in the system (2:1 turns ratio, 4:1 impedance ratio).

Construction data for this 4:1 impedance ratio transformer are given in **Fig 7-30.** The core used is the well known Fair-Rite model # 2873000202 binocular core, equivalent to the Amidon BN-73-202 and the CWS Bytemark B-202-73.

In a 160-meter array you will need at least 640 meters of coaxial cable. RG-6 seems to provide the most cost effective solution, though still expensive. Don't forget that you also need a stretch of land of more than 300 meters to put up the four verticals in line…

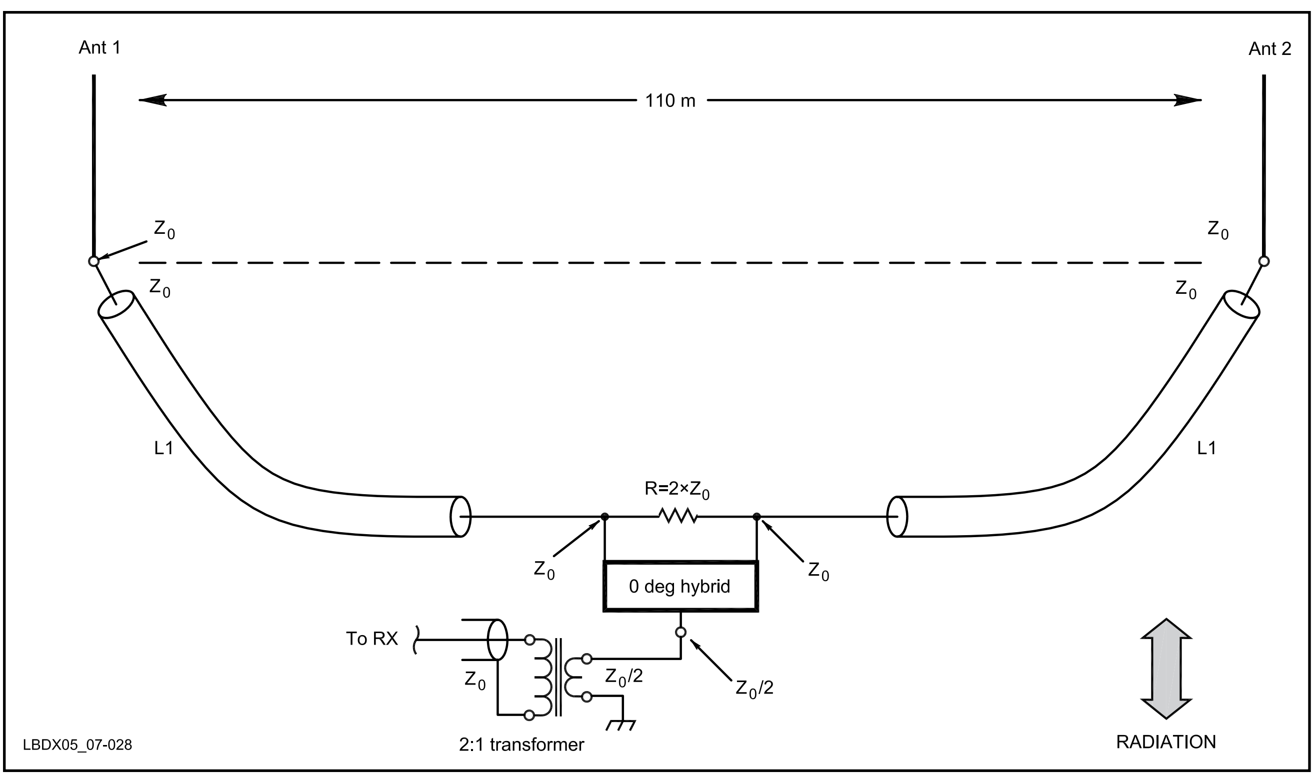


Fig 7-28 — Feed system for a 2-element broadside array using a 0° hybrid combiner and a 2:1 impedance transformer.

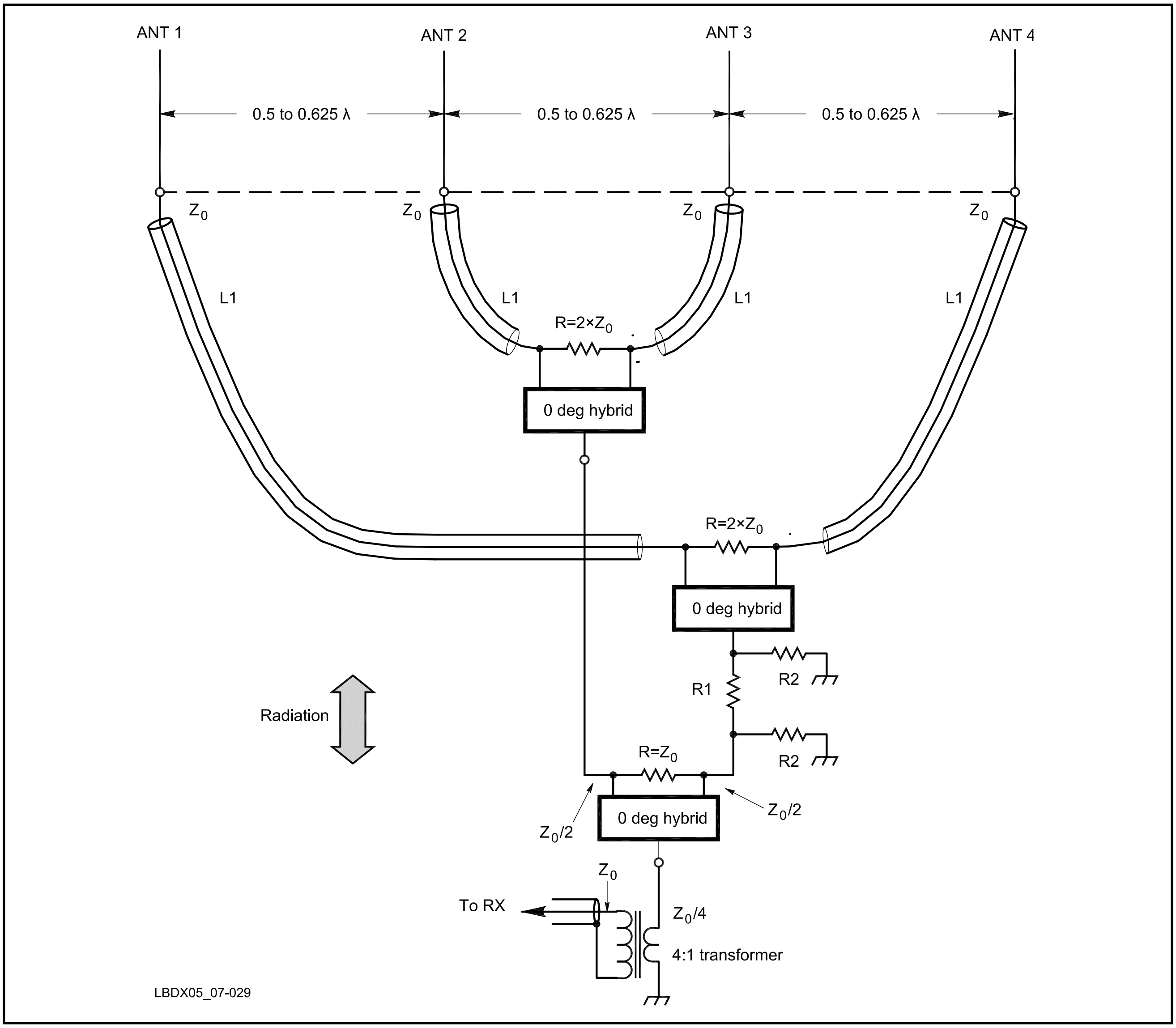


Fig 7-29 — The feed system for the 4-element bidirectional broadside array uses three 0° hybrid combiners and one impedance matching transformer. See text for details.

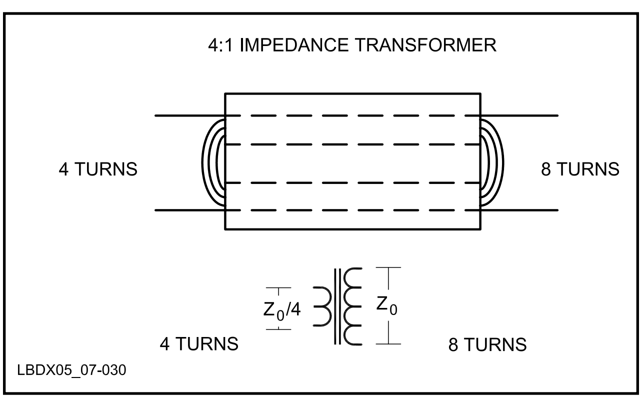


Fig 7-30 — The 4:1 transformer with galvanically separated windings is wound on the same binocular core used for the 0° hybrid.

1.25.3. Feeding the Broadside/End-Fire Array with Two End-Fire Cells

Section 1.12 explained how the array shown in **Fig 7-31** works. Each of the end-fire cells is fed as described in detail in Section 1.20. Box A and Box B contain everything described in Fig 7-27. Each of the two end-fire cells can now be treated as individual antennas in the case of a 2-element broadside array (see Section 1.25.1) and the further feed system is identical to what's shown in Fig 7-28.

1.25.4. Feeding the Broadside/End-Fire Array with Four End-Fire Cells

Feeding this array (described in Section 1.13) is a combination of what is explained previously. As shown in **Fig 7-32**, ideally, we would run four feed lines of identical length to the

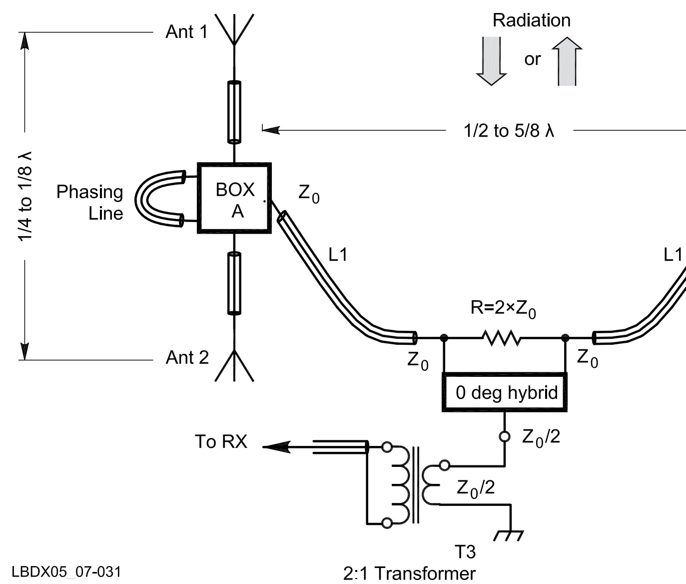


Fig 7-31 — To combine the directivity characteristics of an end-fire array and a broadside array, two end-fire cell are combined. The feed system of each cell is identical (as shown in Fig 7-27) and both cells are fed in phase using a 0° hybrid combiner and a step-up transformer.

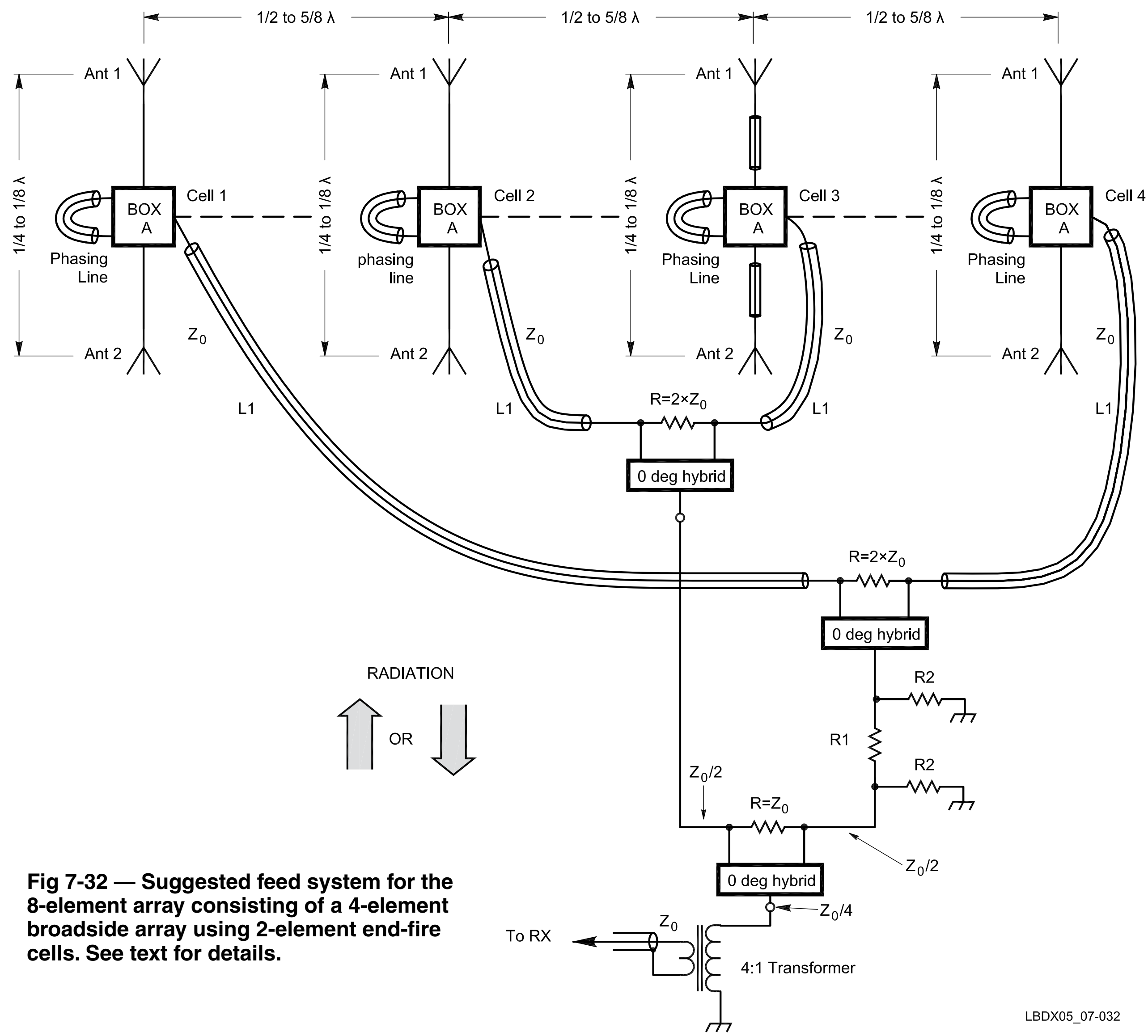


Fig 7-32 — Suggested feed system for the 8-element array consisting of a 4-element broadside array using 2-element end-fire cells. See text for details.

end-fire cells, and use a 6-dB attenuator to reduce the voltage supplied by the outer cells by 50% (see Section 1.24.2).

1.26. The Four Square Receiving Array

Most of us undoubtedly know about the classic Four Square, with λ/4 sides and elements fed (in theory) in increments of 90° (quadrature). See **Fig 7-33**. This configuration became popular because it was the first described in literature, and because it can, as a transmit antenna, be fed with a 90° *hybrid network* (see Chapter 11).

But there is no reason why this 90°/90° (90° side dimension, 90° phase increment) would be magical or better than other configurations. I analyzed three types of Four Squares:

- Large footprint Four Square, with sides = λ/4.
- Small footprint Four Square, with sides = λ/8.
- Very small footprint Four Square, with sides = λ/16.

Fig 7-34 shows the horizontal radiation patterns at a 20° elevation angle for Four Squares with λ/4 side spacing, with various phasing-step increments. Changing the side-element phase angle from 90° to 130° (with rear-element phasing at twice that of side element) does the following:

- It narrows the forward lobe.
- It increases the size of the side lobes.

Note that the geometric F/B remains very high for all elevation angles. The RDF gets better as you increase the phase angle, simply because you substantially narrow the forward lobe. Looking only at the back (RDF), 90° phasing seems to be the best choice. Your choice of best phase angle should be dictated by whether or not you need to look at RDF or DMF. (See Sections 1.8, 1.9 and 1.10.) However, the two-dimensional patterns shown in **Table 7-13** can easily fool you! I would opt for 120° phasing.

Table 7-13 also shows the performance data for a "medium-size" Four Square and for a "mini-size" Four Square. Note that the smaller the array, the lower its gain, but the directivity (RDF and DMF) remains fairly constant. Anyhow, gain is rarely an issue with receiving antennas as long as the antenna gain remains higher than approximately –15 dBi. Compare those figures with those listed in Table 7-8 for the 2-element end-fire array. The horizontal radiation patterns for these Four Squares are shown in **Figs 7-35** and **7-36**.

As for the large-size Four Square, the higher-angle steps result in better directivity and a narrower forward lobe. As we make the array smaller, however, its output (gain) drops. Compared to the large Four Square (λ/4 sides), the medium-sized

Table 7-13

Performance Data for Four Square With λ/4 Side Dimension

Phasing Step	*3-dB Angle*	*DMF (dB)*	*RDF (dB)*	*Gain* (dB)*	*Gain** (dBi)*
90°	100°	19.64	10.13	5.5	–11.2
100°	92°	22.77	10.62	5.1	–11.6
110°	86°	25.70	11.07	4.6	–12.1
120°	81°	25.51	11.49	3.8	–12.9
130°	74°	22.62	11.85	2.9	–13.8

Performance Data for Four Square With λ/8 Side Dimension

Phasing Step	*3-dB Angle*	*DMF (dB)*	*RDF (dB)*	*Gain* (dB)*	*Gain** (dBi)*
130°	96°	19.72	10.43	–1.4	–18.1
140°	86°	25.12	11.24	–3.3	–20.0
150°	76°	27.35	11.83	–5.6	–22.2
160°	66°	22.85	12.39	–8.4	–25.0

Performance Data for Four Square With λ/16 Side Dimension

Phasing Step	*3-dB Angle*	*DMF (dB)*	*RDF (dB)*	*Gain* (dB)*	*Gain** (dBi)*
157.5°	89°	23.12	10.95	–13.3	–29.9
160°	85°	25.88	11.27	–14.4	–31.0
162.5°	80°	28.40	11.60	–15.6	–32.3
165°	76°	27.65	11.93	–17.0	–33.7

*Gain over a single (resistor swamped) element
**Assuming the W8JI-type elements are used (Section 1.21.1)

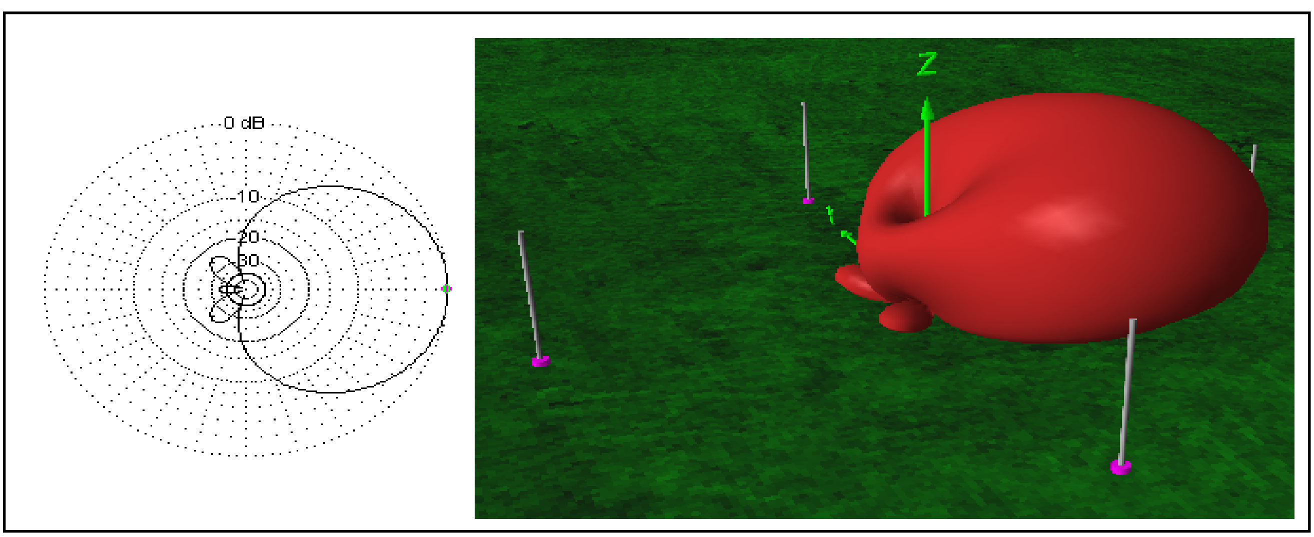


Fig 7-33 — The Four Square has its main direction along the diagonals of the square.

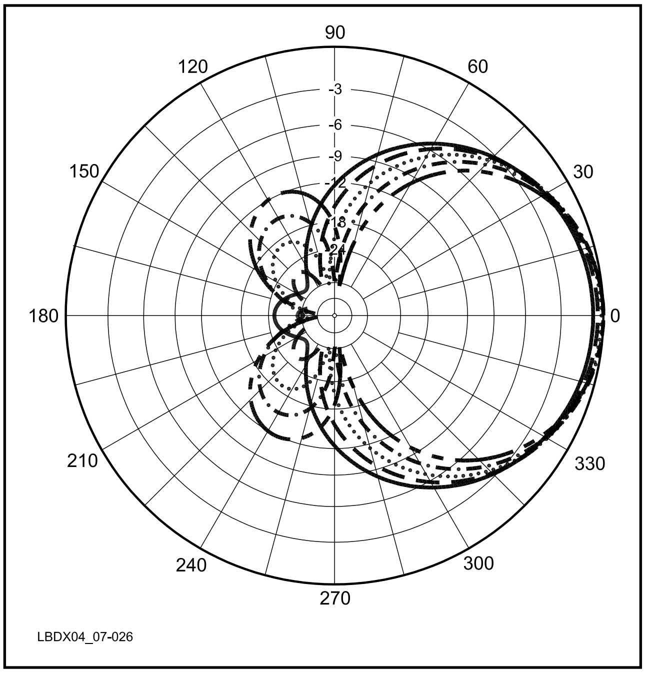


Fig 7-34: Azimuth patterns at a 20° elevation angle for various values of phasing step for a Four Square measuring λ/4 (side dimension). Solid line = 90°; dashed line = 100°; dotted line = 110°; dashed-dotted line = 120°; dotted-dotted-dashed line = 130°.

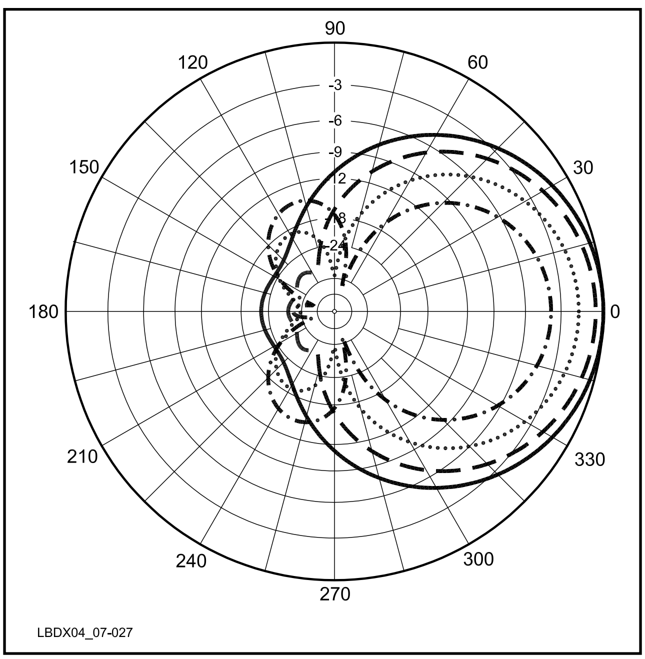


Fig 7-35 — Azimuth patterns at a 20° elevation angle for various values of phasing steps for a Four Square measuring λ/8 (side dimension). Solid line = 130°; dashed line = 140°; dotted line = 150°; dashed-dotted line = 160°.

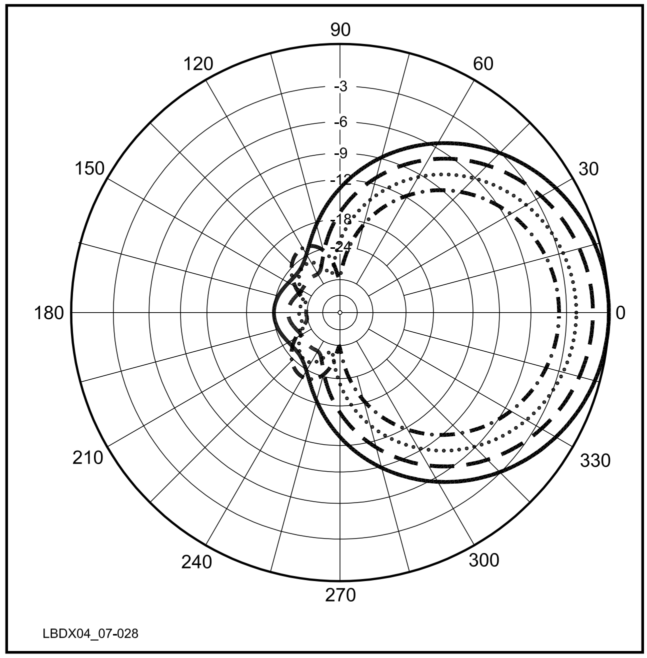


Fig 7-36 — Azimuth patterns at a 20° elevation angle for various values of phasing steps for a Four Square measuring λ/16 (side dimension). Solid line = 157.5°; dashed line = 160°; dotted line = 162.5°; dashed-dotted line = 165°.

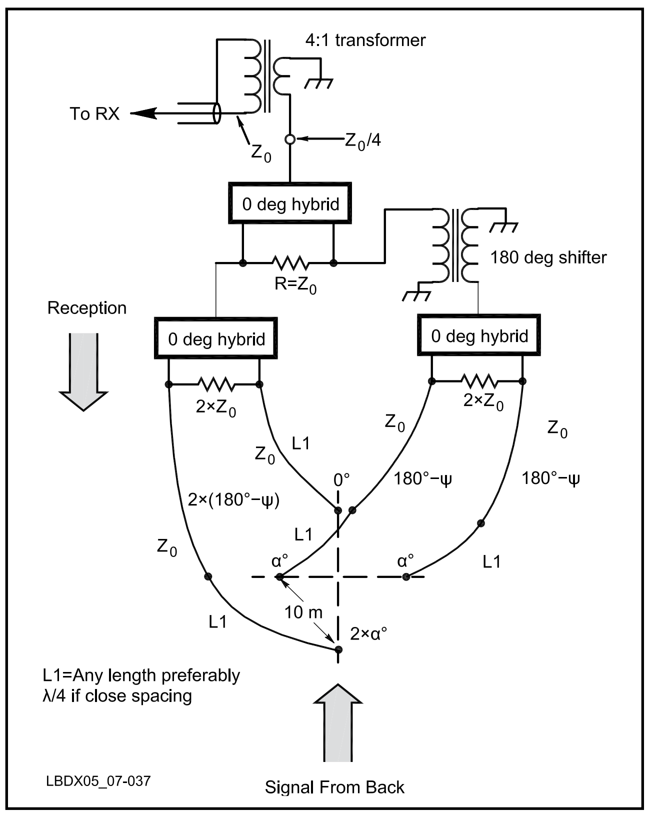


Fig 7-37 — Feed system for the Four Square array, based on the crossfire system. With one set of phasing cables the correct phasing is easily maintained for two adjacent bands (80 and 160 meters). Here too we use the 0° hybrids instead simple parallel combining.

one has 7 to 11 dB less output, and the mini-size Four Square ($\lambda/16$ side) has 19 to 20 dB less — rather dramatic drops, but not relevant for a receiving array.

What we see in these square receiving arrays with various phasing steps is very similar to what we saw happening to the 2-element end-fire arrays (see Fig 7-6).

Transmitting Four Square arrays have sometimes acquired a reputation for not being very good on receiving. The reason is that most transmit Four Square arrays have never been optimized for best directivity, and they have never been fed correctly. Merely hooking such an array up to a 90° hybrid coupler feed system will *not* deliver the anticipated feed current magnitude and phase angle. You can build a Four Square transmit array with element currents and magnitudes way off from what they should be and yet they will still show almost maximum gain, even though directivity might suffer seriously (see Chapter 11, Section 3.4.6).

1.26.1. Crossfire Feed System for the Four Square Array

Tom, W8JI, introduced the crossfire phasing feed system described here. It can be used with any array that requires feed currents with three or more different phase angles. This means it can also be used for the HEX-array described in Section 1.29. As explained in Section 1.20 it tracks frequency over a very wide range.

This system cannot be used for transmitting arrays, as it is based on the antenna elements of the array showing essentially a constant and purely resistive impedance at all frequencies. Refer to **Fig 7-37**. The front element is fed without delay lines. The two center elements are fed like the reflector in the 2-element end-fire array, with $(180 - \psi)°$ delay lines from the inverted output of the phase inverter T1. The rear element of the array is fed by a phasing line that measures $2 \times (180 - \psi)°$.

Let us check if the phase angles remain OK over a wide frequency spectrum. Assume the Four Square has elements that show 75-Ω feed impedance on both 3.5 and 1.83 MHz. We'll neglect the extra phase shift caused by L1, which is the same for all elements, and look at the phase delays. The required feed currents are:

On 160 meters

- $\psi = 150°$
- Element 1: 0°
- Element 2 and 3: 180° – (180° – 150°) = 180° – 30° = +150°
- Element 4: 0° – 2 × (180° – 150°) = 0° –2 × (30°) = –60° = +300°
- Phasing line length to Elements 2 and 3: 30° long
- Phasing line length to Element 4: 60° long

On 80 meters

- $\psi = 120°$
- The phasing lines remain physically the same, but are now twice as long in degrees (60° and 120°)
- Element 1: 0°
- Elements 2 and 3: 180° – 60° = +120°
- Element 4: 2 × (180° – 120°) = –120° = +240°

The required feed currents are listed in **Table 7-14**. The results obtained above are correct. This proves that the system does track and keep the correct phase shift for (theoretically) any frequency. This means we can make a Four Square for 80 and 160 meters using the same feed system. All we have to do is make sure the elements are resonant for the bands we want to use it on (for example, by switching loading coils).

Table 7-14
Required Feed Current Phase for Crossfire Fed Four Square

	160 meters	*80 meters*
Ele 1 (front element)	0°	0°
Ele 2 and Ele 3	150°	120°
Ele 4 (back element)	300°	240°

Fig 7-38 shows the complete wiring of the Four Square receiving array, including a direction-switching system. The circuit uses three 0° hybrid combiners (see Section 1.22, where you can find construction details). The phase inverter (T1) is also described in Section 1.22 (Fig 7-25). The 4:1 transformer (T2) is shown in Fig 7-30. SPST reed relays are used because they have sealed contacts and are cheap (see Section 1.23). Fig 7-38 includes the diode matrix providing the control voltages to the 12 relays from a control cable using five conductors (or four conductors plus shield).

DX Engineering sells the Four Square receiving array hardware (model RFS-2P). On his Web page W8JI shows a simplified schematic of the Four Square system. While in this schematic simple "parallel connection" combiners are used, in reality the RFS-2P hardware uses a 0° hybrid combiner exactly as shown in Fig 7-38. The unit uses three DPDT relays to do the direction switching. While these make the schematic look much simpler, the relay contacts are not sealed and are prone to oxidation after time.

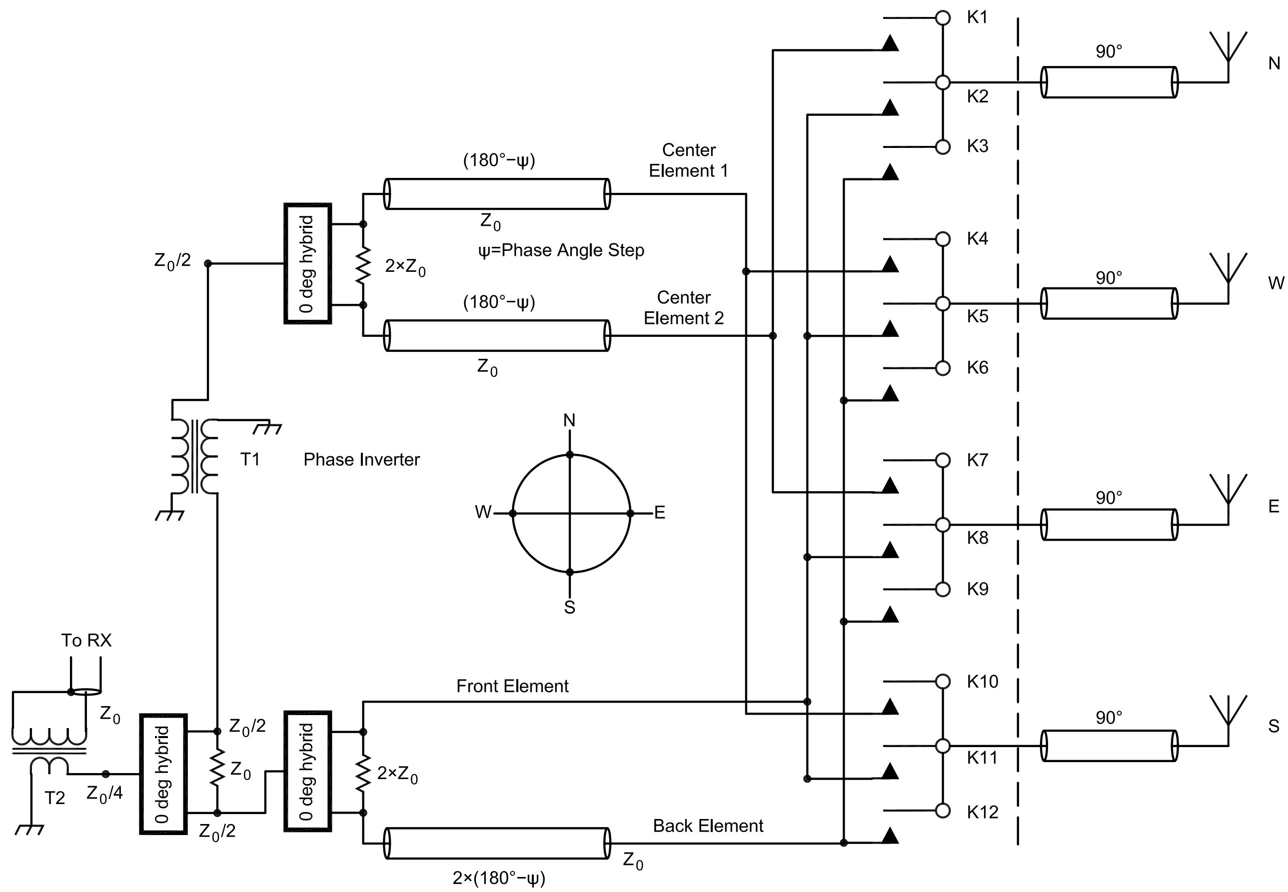
1.26.2. A Word of Caution About Small Receiving (Mini) Four Squares

We have seen that small arrays, although mathematically the equivalent of their bigger brothers, are much more critical to build. Down to $\lambda/8$ spacing there should be no problems, if the array is built with sufficient care. Very small Four Square arrays (eg, $\lambda/16$ spacing) are even more sensitive to variations in feed angle and amplitude. The tolerances are small and extreme care must be taken to keep the operating parameters within strict limits.

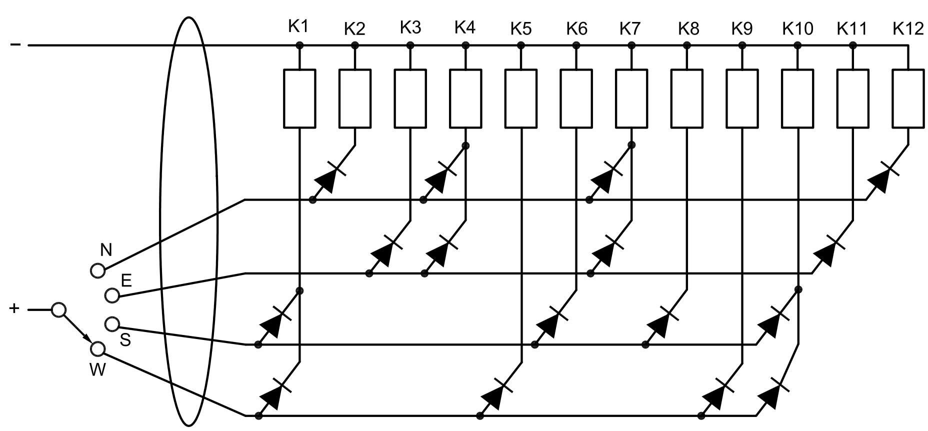
In Figs 7-11 and 7-12, I showed the 2-element end-fire array with $\lambda/8$ and $\lambda/16$ spacing, and explained issues with smaller arrays. The same concerns exist for the Four Square receiving array, but to an even higher degree.

1.27. Checking and Adjusting the Four Square Receiving Array

The first thing you must do is measure the impedance at the base of the elements. Make sure you have a good ground system. How do you make sure? Put a long wire on the ground as a single radial approximately 25 to 30 meters long for 160 meters. Use a single radial of approximately half that length on 80 meters (don't forget that the velocity factor of a wire *on* the ground is around 60%). Measure the impedance of the loaded and resonated vertical with your antenna analyzer using a very short coax, no longer than 1 meter. Connect the radial wire to the ground at the antenna. If you see any appreciable



	N	E	S	W
K1			X	X
K2	X			
K3		X		
K4	X	X		
K5				X
K6			X	
K7	X	X		
K8			X	
K9				X
K10			X	X
K11		X		
K12	X			



LBDX05_07-038

Fig, 7-38 — This feed system for a receive Four Square uses no parallel connections as signal combiners. Only 0° hybrid combiners are used (three in total). T1 is a 1:1 phase inverter, and T2 matches the system impedance (Z/4) to the line impedance to the receiver (usually 75 Ω). Switching is done by using 12 small SPST reed relays. The relays use hermetically sealed contacts, which guarantees long error-free operation (corrosion of contacts being impossible).

change in readings on your antenna analyzer, you will need to improve your ground.

Make sure your coax sees 75 Ω at the design frequency. Measure the impedance at the band edges. The absolute value of the imaginary part of the impedance should not be greater than 8 Ω.

It is also important that the feed lines have exactly the same length. It is not necessary that the feed lines be λ/4 long as is the case with transmit Four Squares. If the array elements are properly loaded and tuned, the SWR on each of the feed lines going to the antenna elements will be ≤1.15:1 (return loss ≥23 dB), which means that phase delay will be almost identical to line length, both expressed in degrees (see also Section 1.18).

Don't use poor quality coax to feed the array. Bury the feed lines in a plastic flexible cable duct about 50 cm deep (if possible). The deeper they are buried the less the chances for common-mode signal pick-up.

To test and adjust the array for best performance, we can temporarily turn it into a very low power transmit array and inject a low power signal. Use just a few watts — make sure the antenna element loading resistor can take it! At the end of the equal length feed lines we have installed small sampling resistors measuring 5 kΩ. This is where we can connect our test equipment. A vector voltmeter (such as a surplus HP-8405) or a more expensive vector network analyzer is ideal for the purpose, but a good dual-trace oscilloscope can be pressed into service if the user is careful.

If you are serious about antennas, you might consider acquiring a vector network analyzer such as the VNA 2180 (see Chapter 11, Section 3.5.2.4) and the matching vector scope hardware as described in Chapter 11, Section 3.6.2. With such a setup you'll be equipped to measure and adjust your receiving arrays as well as your transmit arrays.

1.27.1 Level Adjustment

Most important is to adjust the signal levels in the four branches of arms. In view of the fact that different components (phase inverter) and different coaxial cable lengths are used in each arm, the attenuation will be slightly different in each branch.

This is why we should fine-tune the current magnitudes using small attenuators installed in two of the network branches. In **Fig 7-39** we can see that the most attenuation will likely occur in the leg going to the back element, as it uses the longest cable (150°). So that leg should *not* have an attenuator. The next highest attenuation will occur in the center elements. So an attenuator (Att2) needs to be in this leg to increase the loss so that it equals the loss incurred in the back element leg. A single attenuator after the 180° transformer will work (Z_0 however is 37.5 Ω here, not 75 Ω). The front element leg uses no coaxial delay cable and has the least amount of attenuation, so it requires an attenuator (Att1).

You can use your transceiver, followed by a good filter such as a W3NQN bandpass filter, as an RF power source. Make sure you inject no more than a few watts, as the small cores used for the 0° hybrids and other components will not tolerate more than approximately 10 W if you want to play it safe. Anyhow, the resistors in use in the hybrids may be the weakest link as far as applying too much power!

Similar precautions must be taken with the components used to load the antenna elements. That means that miniature coils used on printed circuit boards are out of the question. Wind your own antenna loading coils on small powdered iron cores. Use loading resistors of sufficient wattage; connect several in parallel if necessary.

The amount of adjustment required is on the order of 0.3 to 1.0 dB. A couple of 4.7 Ω series resistors (R_s) and a 500-Ω carbon variable for the parallel resistor (R_p) can be used to make the attenuator for the leg with a 37.5 Ω impedance. For 75 Ω, R_s is 1000 Ω and R_p 2.2 Ω. These values give close impedance matches and any mismatch created by imperfect attenuator values can be compensated by adjusting the attenuation.

1.27.2. Phase Checking

If you have access to an RF vector voltmeter or similar equipment capable of accurately measuring RF voltage magnitude *and phase*, you can measure the voltages at the ends of the feed lines going to the elements. The T attenuators can be fine-tuned for identical voltage magnitude on the three lines. There is not really a way to adjust the phase of the four signals. If the SWR on the feed lines is ≤1.15:1 (≥23 dB return loss), the measured phased should come out to be more or less exactly what was calculated. In this case the phase delay in the cable will track the line length within a few degrees, which is more than adequate. Such slight variation will only move around the angles (both horizontal and vertical) at which the greatest rejection is obtained (see Table 7-1), but will not greatly affect the overall directivity.

Robye, W1MK found that since using the 0° hybrid combiners, problems with incorrect phase shifts are almost nonexistent because of the excellent isolation between the antennas. Since using this form of combiner he merely uses the 5 kΩ pick-off points to connect to an RF power meter which reads out in dBm. A suitable power meter (which also includes a frequency counter) is the FPM1 available from M Cubed (**www.m3electronix.com/fpm1.html**). It can measure the relative power between the different ports to within 0.1 dB. Using the attenuators Att1 and Att2 he can adjust the levels very quickly.

1.27.3. Using the W1MK 90° Hybrid

If you do not have access to a vector voltmeter or VNA (vector network analyzer) and still want to check the phase, you can build a simple piece of test equipment designed by Robye Lahlum, W1MK. Using two 90° hybrids shown in **Fig 7-40**, you can adjust the attenuator values for the deepest null at the output port (3) of the hybrid. You will, however, need a few watts of RF to do this (see Section 1.27.1).

The hybrid coupler is a one-band device. The values of the components are calculated from:

$$L = \frac{Z_0}{2 \pi F}$$

where

Z_0 = characteristic impedance (here 75 Ω)

F = frequency in MHz.

For 1.83 MHz the coil inductance is

L = 75/11.5 = 6.5 μH

and

$C = 10^6 / (4 \pi F Z_0) = 580$ pF.

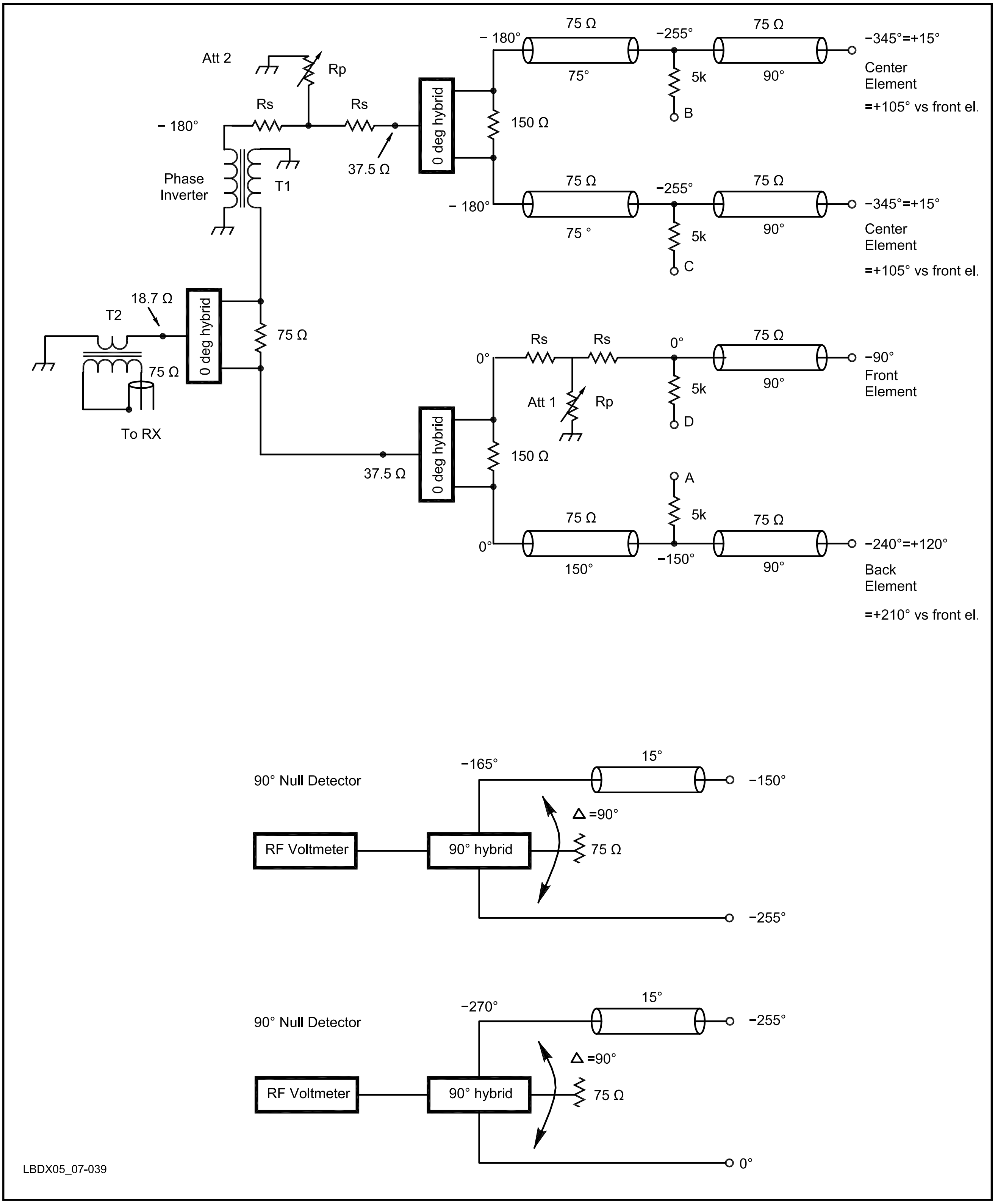


Fig 7-39 — Two small attenuators are inserted in the front and center legs, where the least total attenuation is needed because of feed cable losses to the side elements. These attenuators can be adjusted to obtain the same feed-current magnitude in all elements, as witnessed by a full null on the RF voltmeters of the two hybrid circuit/ null detectors.

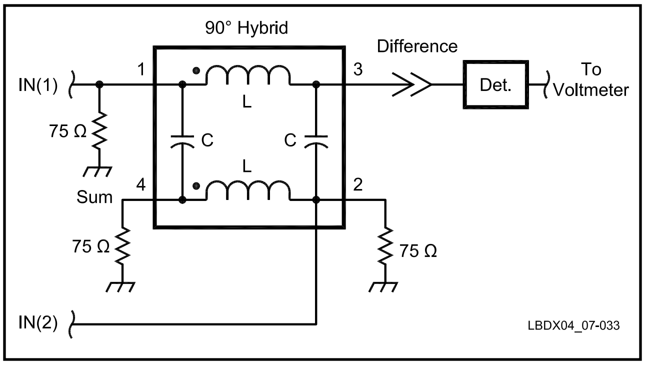


Fig 7-40 — A 90° hybrid: if signals at IN(1) and IN(2) are exactly 90° out-of-phase and have the same amplitude, the output of the coupler to port 3 will be zero.

The coils must be wound in a bifilar fashion (the wires can be twisted). Note the phasing dots in Fig 7-40. Powdered-iron cores must be used. A T-50 core (red mix) has an A_L of 49 (meaning that 49 turns are required for 100 µH inductance). The required number of turns is given by

$$N = 100\sqrt{\frac{L}{A_L}}$$

where

L = inductance in µH

A_L = inductance index (µH/100 turns)

If the array requires 90°-phase shift between the elements ($\psi = 90°$), you need not add any additional phasing line to one of the inputs of the hybrid. In Fig 7-39, I inserted 15°-long phasing lines, since the ψ of this array is not 90° but 105° (15 = 105 – 90). Make sure you have the 75-Ω line termination resistors at ports 1, 4 and 2 of the hybrid. You can connect the hybrid to proper points (A, B, C and D) at the input of the λ/4 lines to the elements, with short but equal lengths of coax.

There is no real need for these circuits to be made for a 75-Ω system impedance; they can just as well be made for 50 Ω. In that case the extra phasing lines should be made with 50-Ω cable.

In Chapter 11, Section 3.6.1, there is more detailed information on this adjustment procedure.

When using low RF power, you can use a receiver instead of a detector and voltmeter for a null indicator. Make sure you have an extra attenuator inline with the input of the receiver and make sure you apply the minimum amount of power necessary to obtain a clear null. Do the adjustment during the day when there are no signals on the band.

The two hybrid test circuits can be left inline permanently. To test the array, inject a few watts of power, connect a receiver to the output port of the hybrid and adjust the T attenuator(s) for minimum signal. You can, of course, use a small dedicated detector/voltmeter instead of a receiver for this purpose (see Chapter 11, Section 3.6.1.1).

One of the problems associated with this method is that for some types of arrays the extra length of coaxial cable going to one port of the 90° hybrid will introduce some small amount of attenuation and thus falsify the results. In the example of the Four Square shown in Fig 7-39 the lengths of the extra coaxial cables are only 15° which, if low loss cable is used, will not influence the test method a great deal.

1.28. The Mini Receiving Four Square at ON4UN

During the winter I can use the terrain you can see in the background of the picture in **Fig 7-41** to put up 12 Beverage antennas (the terrain is approximately 160 by 300 meters). Fortunately most of the DX on the low bands is worked in

Fig 7-41 — You can see three of the elements of the mini Four Square in the front yard at ON4UN in this photo.

Fig 7-42 — Radial layout used at ON4UN's mini Four Square. Each vertical has 18 short ground radials, which are soldered to a bus wire. A 1.5-meter long ground rod is used at each element.

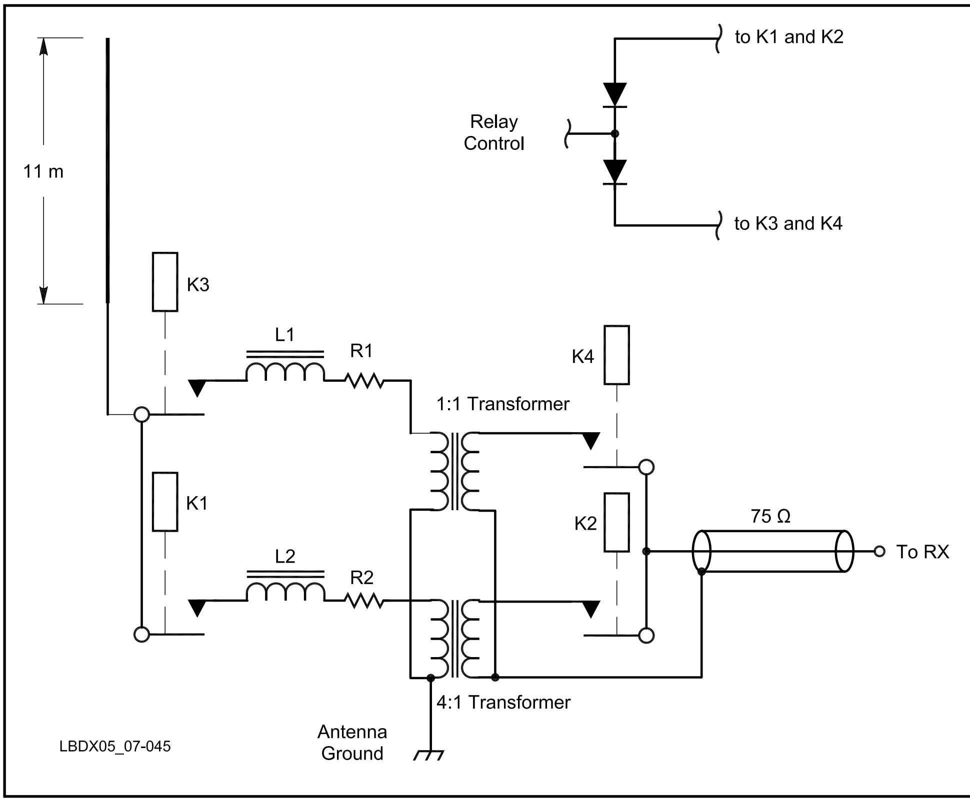


Fig 7-43 — Duo-band switching system used by ON4UN. See text for details.

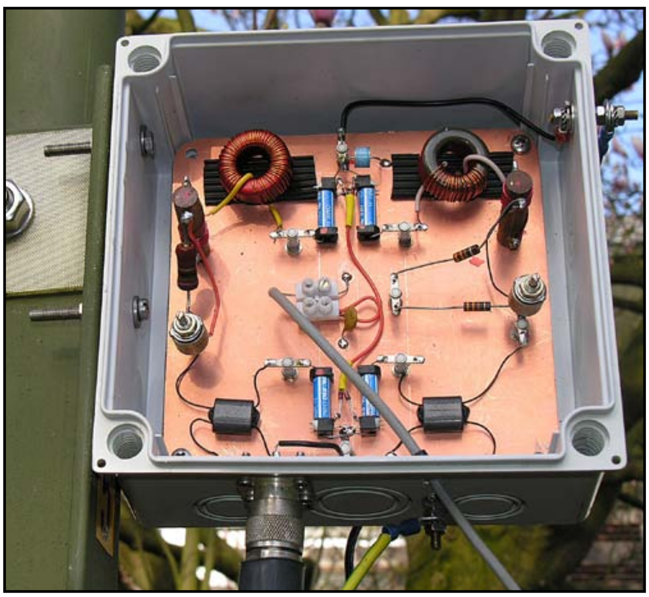


Fig 7-44 — A plastic box houses the matching and loading (swamping) components, as well as the two "braid-breaker" transformers and the reed relays for switching bands.

Fig 7-45 — Base of a vertical. All copper radials are soldered to a copper ring, and connected to the strap going to the 1.5-meter long ground rod.

winter, although I have often put up a single Beverage across the cornfield in the summer when it was necessary to work a new one! You really should try to lay a Beverage wire on top of 2.5-meter tall corn: a unique experience!

To have some decent receiving capabilities in the summertime, I decided to try a receiving Four Square in the small yard in front of the house. The array is about 40 meters from the 160-meter transmit antenna. This is really too close and requires the transmit antenna to be detuned while I'm listening. (See Section 3.11.) **Fig 7-42** shows a bird's eye view of my front yard in a picture taken from the top of one of the other towers. The schematic for the switching employed at each of the duo-band 80/160-meter elements is shown in **Fig 7-43**. **Fig 7-44** shows the weatherproof plastic box used to hold the switching/matching/loading components for each vertical. The Four Square measures 11 meters on each side.

Long radials were not possible, so I put down about 18 radials per element, after installing two crossing bus wires to which the ends of the inward-looking radials are soldered. See **Fig 7-45**. With a 1.5-meter long ground rod at each element, I estimated the equivalent ground loss resistance to be approximately 20 Ω.

It is quite easy to determine the ground loss resistance. Assume your element radiation resistance is 2 Ω, and the loading-coil equivalent series loss resistance is 2 Ω. We need to add a 50-Ω series loading resistor to obtain a 1:1 SWR in a 75-Ω system, so we can conclude that the ground loss resistance is 75 – 2 – 2 – 50 = 21 Ω.

I wanted to use the small Four Square on both 80 and 160 meters. Using a loading resistor with a typical value of 50 to 60 Ω, to end up with a 75-Ω resistance (R1 plus ground losses plus losses in loading coil L1) this 11-meter long vertical gives adequate signal output (–15 dBi) on 160 meters. It has reasonable SWR bandwidth, measuring approximately 1.2:1 on 1810 kHz and on 1850 kHz (see Table 7-12). The required loading coil has an inductance of ~60 µH.

On 80 meters I needed much more bandwidth, which means much more resistive loading. This is no problem as the 11-meter tall vertical is fairly long for this application and extra swamping is required to reduce the effects of mutual coupling. Instead of loading the antenna to a total resistance of 75 Ω, I loaded the antenna to 300 Ω, which resulted in an output of approximately –15 dBi, the same level as on 160 meters.

On both bands I use a transformer with galvanically separated primary and secondary windings to connect the loaded antenna to the feed line. On 80 meters the transformer has a 1:1 impedance ratio, and on 160 meters a 4:1 ratio (transforming 300 Ω down to 75 Ω). These transformers also serve as "braid breakers" (see Sections 2.7.2.8 and 2.7.2.10) which attenuates the ingress of common-mode signals on the feed lines by some 42 dB on 160 meters. This assumes the antenna ground system resistance is approximately 50 Ω, the surge impedance of the feed line braid working as an antenna is about 300 Ω, and the inter-winding capacitance of the transformer is 16 pF (Z = 5.5 kΩ). See also Fig 7-95 later in this chapter.

On 160 meters the 1:1 transformer can be constructed as explained for the common-mode choke in Section 2.7.2.9. On 80 meters we need to use a small 4:1 broadband transformer, to bring the impedance down to 75 Ω. A suitable 4:1 impedance ratio transformer is shown in Fig 7-30.

The bandwidth of this 80-meter element is extremely wide, with an SWR of 1.1:1 on 3.5 MHz, 1:1 on 3.65 MHz and 1.1:1

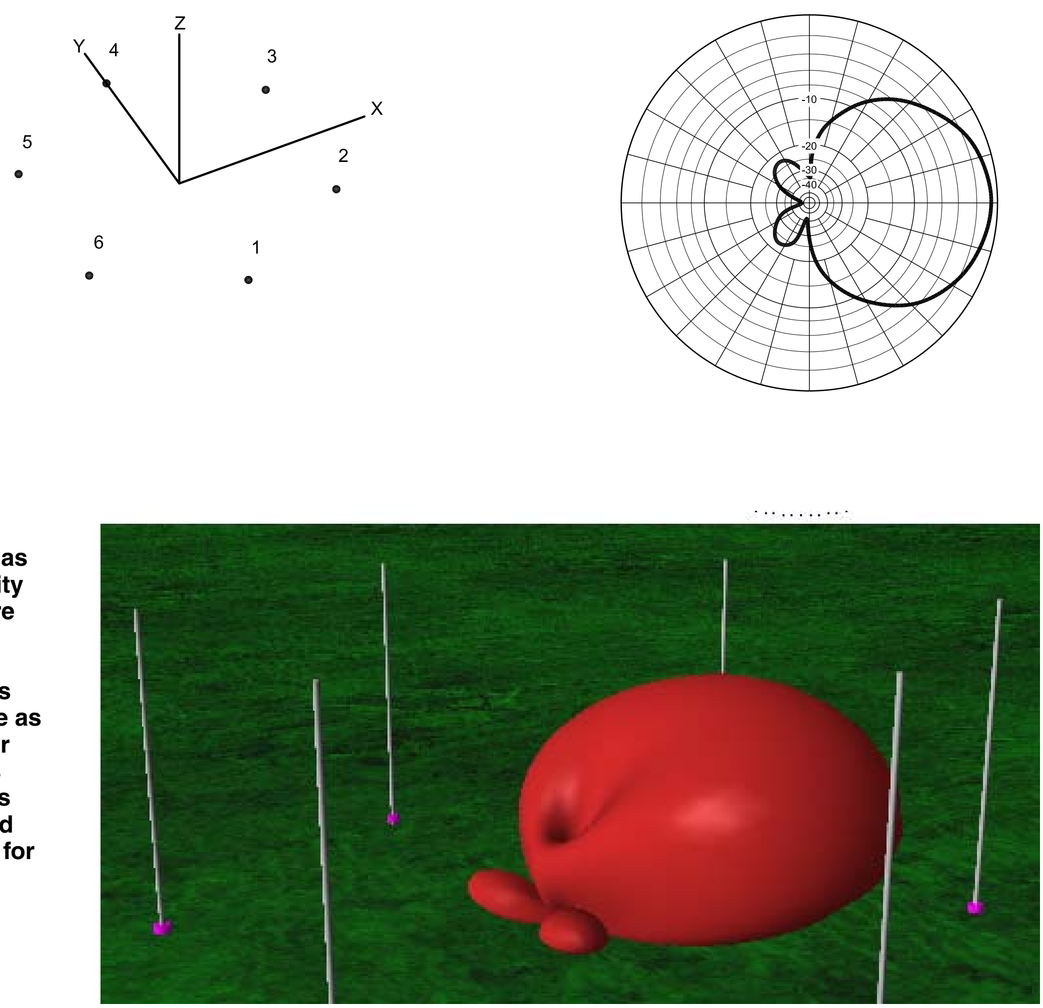


Fig 7-46 — The Stone-Hex array has improved directivity over a Four Square and is switchable in six directions. The feed system is basically the same as used with the Four Square, since this array also requires three different feed currents. See text for details.

on 3.8 MHz. Fig 7-44 shows the band-switching arrangement at the base of each element. Four small reed relays (SIL housing) are used and selection can be done via a positive/negative voltage. If necessary the dc voltage can be brought in via the coaxial cable and the usual C plus RFC circuit.

Using 16 short radials and a 1.5-meter ground rod at each vertical, the required value of the loading resistor for 160 meters (R1) was approximately 50 Ω to achieve a 75-Ω feed impedance. For 80 meters a value of 270 Ω was used (R2).

Use metal-composition resistors (not film resistors), such as Ohmite OY/OX ceramic composition resistors or equivalent in this application. I wound the coils on 1.3-inch OD powdered iron cores (T-130, Red mix, $A_L = 100$). The number of turns is given by the equation shown in the previous section. A coil with an inductance of 50 μH requires:

$$N = 100\sqrt{\frac{50}{100}} = 67 \text{ turns}$$

After installing the elements, check the feed impedance on the design frequency. It should be as close as possible to 75 Ω. Also check the band edges. The SWR at the band edges must in no case be higher than 1.2:1 on 80 meters (3.5 MHz and 3.8 MHz) and on 160 meters (1810 kHz and 1850 kHz).

1.29. Stone-HEX Array (K8BHZ)

Brian, K8BHZ, sent me details of a 6-element round array, which he calls a Stone-HEX (referring to Stonehenge) — see **Fig 7-46**. K8BHZ's hexagonal array is small: the six elements are located on a circle measuring only 30 meters in diameter. **Fig 7-47** is a panoramic view. The array can be switched in six directions, which is exactly what's needed with its 3-dB beamwidth of 78°.

Look at this Six Circle array as if it were a Four Square but with two front and two back elements in addition to two central elements. Refer to Fig 7-46, where the antenna is shooting along the X axis. In that case elements 2 and 3 are front elements, element 1 and 4 are central elements, and elements 5 and 6 are the back elements.

Table 7-15 gives the design and performance data for the Stone-HEX array for three different sizes (modeling files: *chap7_6circle-30mdiam.12mvert.ez*, *chap7_6circle-45mdiam.12mvert.ez* and *chap7_6circle-60mdiam.12mvert.ez*. At first glance it may look like there is little reason to make the array larger than necessary. But read Section 1.18, which will tell you not to make the array any smaller than necessary, and explain why.

1.29.1. The W1MK Improved Feed System

First read Sections 1.24, 1.25, 1.26 and 1.27 before proceeding here. This array also uses the crossfire feed system like we did with the Four Square, as this guarantees perfect frequency tracking over a wide range.

Just like the Four Square array, the Six Circle array requires three different feed currents. In addition, these do not all have the same feed-current amplitude. The antenna currents required at the antenna element feed points are:

- At the 2 front elements : I = a ∠0° (is the reference element with phase = 0°)
- At the 2 middle elements: I = 2a ∠+ψ°
- At the 2 back elements: I = a ∠+2ψ°

To obtain these phasing angles in a crossfire feed system, the lengths of the phasing lines are:

- To the front element: no phasing line (is the reference element)
- To the center elements: (180° – ψ), where ψ = the phasing angle step
- To the back elements: 2 × (180° – ψ)

For the 30-meter diameter Stone-HEX, with ψ = 149°, the length of the phasing line to the central elements is (180° – 149°) = 31°. The length of the phasing line to the back elements is 0° – 298° = 360° – 298° = 62°. These are electrical lengths. When converting to coax length take into account the velocity factor of the coax.

As in the Four Square array, the feed lines running from the switch box to the array elements are equal in length, preferably λ/4 long. This allows us to measure voltage at the end of these lines and know the currents at the element feed points.

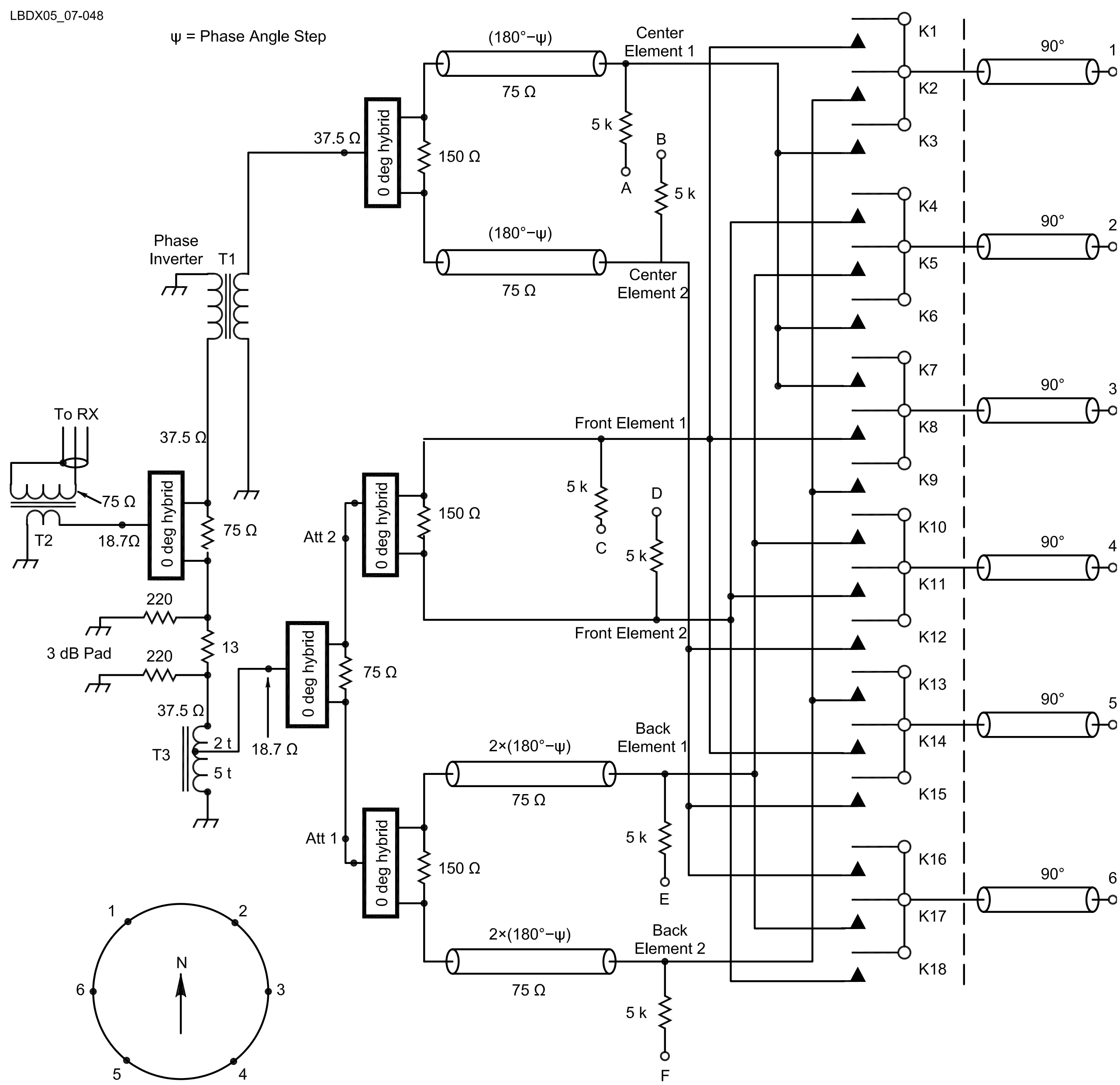
As a receiving array, each element sends equal currents toward the load. Robye, W1MK, designed the feed circuit which is shown in **Fig 7-48**. It uses four 0° hybrid combiners as described in Section 1.22. Details of the phase inverter T1 are shown in Fig 7-25. Details of the 2:1 impedance transformer (T3) are shown in Fig 7-26 and the 4:1 impedance transformer is shown in Fig 7-30.

The direction-switching shown in Fig 7-48 uses the reed relay matrix switching system

Table 7-15
Three Sizes of Stone-HEX Arrays

Circle Diameter (meters)	*Phasing Angle ψ (degrees)*	*Gain Over One Element (dB)*	*3-dB Beamwidth (degrees)*	*DMF (dB)*	*RDF (dB)*
30	149°	–2.42	79	28.02	11.70
45	130°	+3.28	80	26.68	11.47
60	120°	+5.34	75	26.54	11.67


Fig 7-47 — The Stone-HEX array as set up at K8BHZ.



	N	NE	SE	S	SW	NW
K1	X					X
K2			X			X
K3		X			X	
K4	X	X				
K5					X	
K6			X			X
K7	X			X		
K8		X	X			
K9					X	

	N	NE	SE	S	SW	NW
K10	X			X		
K11			X	X		X
K12		X			X	
K13	X	X		X		
K14				X	X	
K15			X			X
K16	X			X		
K17		X	X			X
K18					X	

Fig 7-48 — Optimized feed system for the K8BHZ Stone-HEX array. The system uses 18 small reed relays, which have hermetically sealed contacts, and thus provide long life. The circuit also used a total of four 0° hybrid combiners which help in broadbanding the directivity characteristics of the array (see Section 1.22)

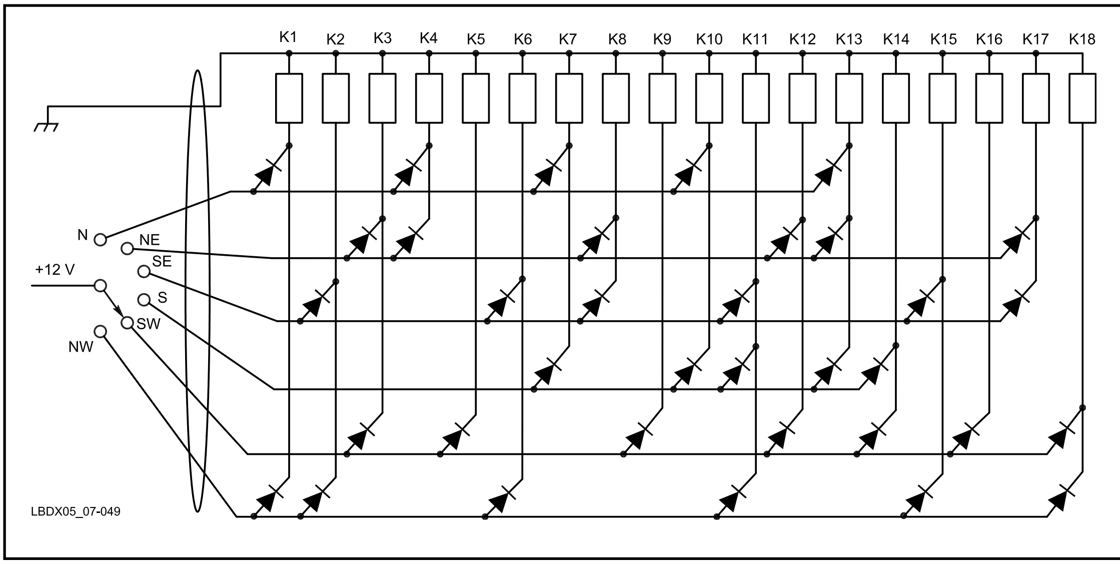


Fig 7-49 — The 18 reed relays can be switched via a 6+1 conductor cable, using the small diode matrix as shown.

originally developed by Brian, K8BHZ. The circuit uses 18 small, inexpensive SPST reed relays. **Fig 7-49** shows a diode matrix for switching the relays.

The improved feed system for the Stone-HEX (or Six Circle) array uses 0° signal combiners in a system originally developed by W1MK (Section 1.22). As we are combining the signals beyond the phasing lines using 0° hybrid combiners, we now have six "rails" — two for front, two for center, and two for rear elements. Three simple SPST reed relays per element (a total of 18 relays) connect each of the feed lines to these elements with one of the six rails.

Note also that the original K8BHZ Six Circle Stone-HEX uses a 50-Ω system impedance because the builder had easy access to 50-Ω cable. If you have to go out and buy cable, let me suggest building everything with 75-Ω cable. For a similar construction and diameter, 75-Ω cable has less loss than the 50 Ω variety. Flooded CATV cable such as RG-6 is used widely and is available at very reasonable cost. I use 75-Ω cables for all my receiving antennas.

1.29.2. Six Circle Level Adjustment

To perfectly balance the amplitude of the feed currents, we can add two T-attenuators, one in the feed line going to the front element (Att 2) and one going to the back element (Att 1) as shown in Fig 7-48. It is likely that the total loss in the leg going to the middle elements (including the loss in the phase inverter) will be greater than the loss in the two other legs. If that is *not* the case, you may have to change one of the attenuators to another "leg" of the circuit. See Section 1.27 for details on these attenuators.

Test points A through F are provided on the feed lines at 90° distance from the antenna elements. This means that the voltages measured at this point are a perfect picture of the current at the antenna feed points (I = E/Z).

For further details see Section 1.27.1.

1.29.3 Six Circle Phase Adjustment

We can use the "W1MK 90° hybrid method" to check correct phasing, as described in Section 1.27.3. In the case of a Six Circle with a 149° phase step (diameter 30 meters, see Table 7-15), we will need to put an extra length of coax going to the leading phase input of each 90° hybrid of (149° – 90°) = 59°.

1.29.4. The Six Circle Ground System

K8BHZ uses an unusual ground system for his array that relies on many judiciously located ground wires plus rods rather than an elaborate radial system. Brian had measured the ground loss for a single 2.4-meter long rod to be about 90 to 100 Ω. He came up with the idea of connecting multiple ground rods to lower this and to average out variations in that value.

Brian used the "Moxon monopole/counterpoise symmetry," which means that the connecting wires between the three ground rods are an exact replica of the top-hat wires; just as long and directly underneath. This way the ground resistance at each vertical was brought down to about 30 Ω.

The resistance of the ground system for resistor loaded verticals is not very low, as we add resistance anyhow, but it ought

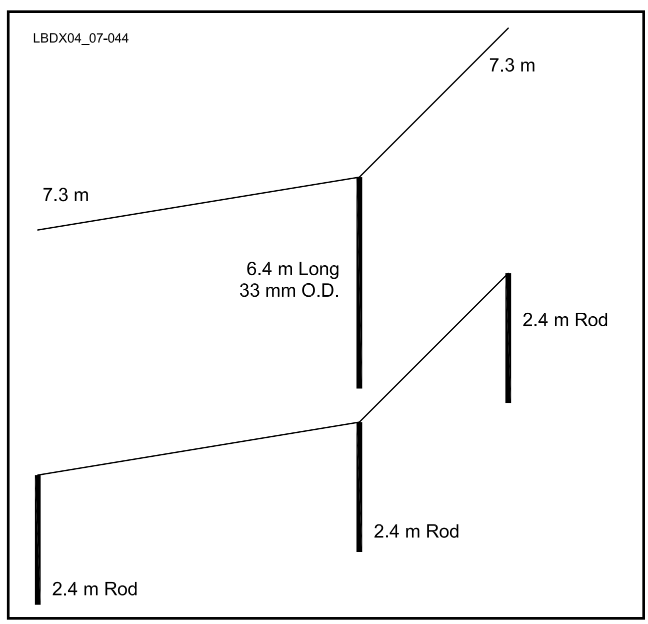


Fig 7-50 — The K8BHZ ground system uses three 2.4-meter long ground rods connected with a wire that is an exact mirror image of the loading wires.

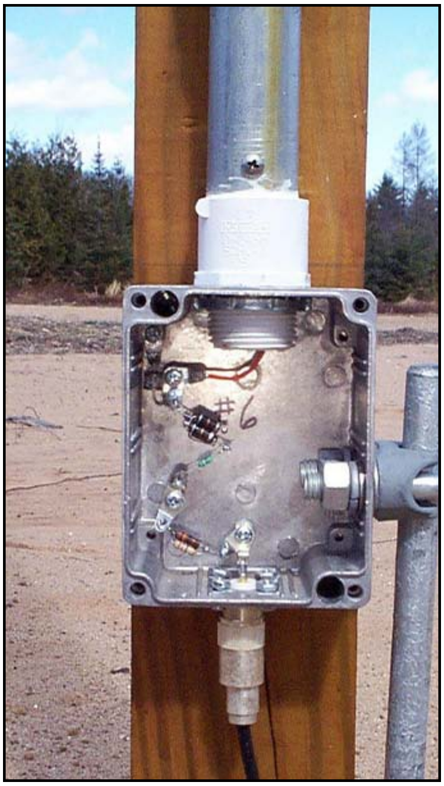


Fig 7-51 — K8BHZ uses 10 cm × 10 cm treated wooden fence posts to support his 6.4-meter long elements. The 2.4-meter long ground rod emerges above ground and is clamped directly to an aluminum die-cast box.

Fig 7-53 — A 10-gallon garbage can is used to house K8BHZ's phasing/switching circuitry, as well as the extra phasing-line coax (coiled up at the bottom of the can). The can is grounded using two 2.4 meter long ground rods.

to be stable, between winter and summer, dry and wet weather.

Brian uses three 2.4-meter long ground rods, one at the base of the vertical, and two more at the end of the 7.3-meter long radial wires running exactly beneath the top loading wires (see **Fig 7-50**). There are no other ground wires, which means that no extra room is required outside the circle on which the array is built.

Brian says that the Moxon monopole/counterpoise symmetry is one of the keys to the very clean pattern seen on *EZNEC*. *EZNEC* is one thing, real life another, but if the *EZNEC* model is good, you have a good starting point anyway. I am of the opinion that, if you use just a few radials, the Moxon geometry is a good thing. With so few radials, they also may act as poor radiators and upset the pattern. If you use a lot of fairly short radials, they act like a ground screen and tend to not upset the radiation pattern. On my small Four Square I use 18 short radials, and they form a fairly dense "screen" which, I am sure, will not influence the radiation pattern a lot.

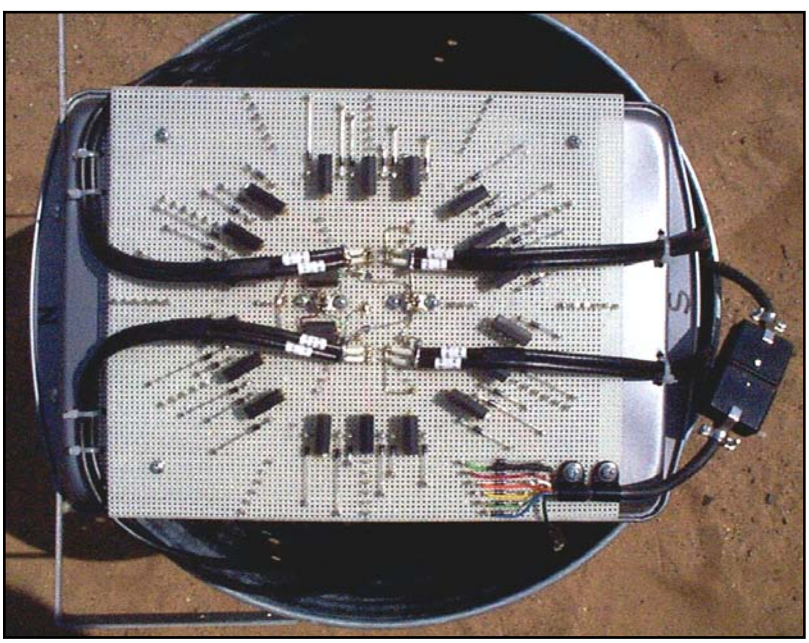


Fig 7-52 — The phasing/matching/switching board made by Brian, K8BHZ. Note the small reed relays.

Fig 7-51 shows the base of a vertical. Brian uses small printed-circuit type inductors to tune the elements. Note how the box is clamped to a ground rod with a large U-clamp.

Personally, I have had extremely poor experience with galvanized and copper-clad steel ground rods. They have proven to rust up and literally disappear. Now I use copper tubing, approximately 30 mm diameter. This a little more expensive, but so much more reliable.

If you solder a conductor to a solid copper-bar ground rod, cover everything with a heat-shrink tube with a hot-melt adhesive inside. Never expose solder joints done with regular solder to bare ground. Always cover the solder joints liberally with a couple of layers of liquid rubber before burying.

Fig 7-52 shows the construction of the matching/phasing/switching board by K8BHZ, using a large perfboard. Note the excellent RF layout, which results in minimum stray coupling. **Fig 7-53** shows the board in a 10-gallon garbage can used to protect it from the elements.

The direction-switching lines are the long outer conductors connected to the rotator cable. The symmetrical relay triads are self-explanatory; the 6-element connectors are directly below them. The short inner rings are for Front/Center/Rear, and the two transformers are visible, as are the phasing lines coming and going. The output connector is directly at the center. Brian also mentioned he made a second layout for a two-sided PCB with ground plane.

1.30. The Eight Circle Array (W8JI)

The Eight Circle array developed by Tom, W8JI, is undoubtedly one of the best performing receiving arrays using vertical elements, with directivity results equaling those of phased Beverages. In Section 1.12 I described the end-fire/broadside array. W8JI developed a direction-switchable array based on this four-element cell. **Fig 7-54** shows the design and the relationships between the end-fire spacing (EFS) and the broadside spacing (BSS).

A broadside spacing of 0.55 λ (90 meters on 1.83 MHz) results in the highest attenuation off the side for a 24° elevation angle [elevation angle = arc cosine (82/90)], where 90 meters is the separation and 82 meters is λ/2 at 1.83 MHz. W8JI used a separation of 107 meters (0.65 λ), which results in a few more sidelobes but a substantially smaller forward beamwidth. The "gain" remains the same within 0.1 dB and is approximately 7.6 dB over a single element.

All of this means it is not an antenna for a small backyard. You need almost 20,000 m^2 (180,000 sq ft, or a little over 4 acres) of land to build this array.

The highest RDF and DMF are obtained with a phasing angle of approximately 100° to 110° (**Table 7-16**). From the point of view of total noise these are the choice phasing angles. Feed currents:

Elements 1 and 3: 1 ∠0° A

Elements 3 and 4: 1 ∠+ψ° A

With these feed currents the array will be pointing North (in Fig 7-54). Note than only four of the eight elements are in use in any direction. (See modeling file *ch7-8circle-90mbroadside.ez* on the CD.) It is obvious that the elements that are not used should be left floating (as they are short elements they will be resonant at a much higher frequency than the design frequency of the array), and certainly not connected to ground.

Contrary to what we saw with the 2-element end-fire, the Four Square and the Stone-HEX array, this array cannot be scaled to a smaller version and still maintain its excellent directivity. This is because scaling would change the broadside phasing distance, which must be a little over λ/2 for proper operation. This is also the reason why the array only works on one band.

It's interesting to compare this array with the Stone-HEX array (Section 1.29). Gain-wise there is just over 1 dB difference, but who cares about gain for a receiving array? From a DMF point of view (looking at directivity in the back), there is very little difference between the two arrays. As expected the RDF is about 1 dB better with the Eight Circle because of its narrower forward lobe. The λ/2 broadside spacing of the two end-fire cells does the trick.

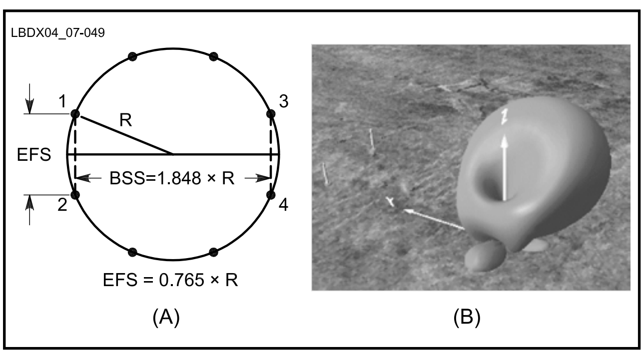


Fig 7-54 — The Eight Circle is nothing more than a set of end-fire/broadside arrays, with the elements arranged at 45° intervals on a circle. Properly dimensioned (slightly more than λ/2 side-spacing), this configuration makes a super receiving array that can be switched in 45° intervals.

Table 7-16
End-Fire Cells with 37.25 Meter Spacing

Phasing step	*90°*	*100°*	*110°*	*120°*	*130°*	*140°*
DMF (dB)	19.8	21.39	23.37	21.78	20.15	18.11
RDF (dB)	11.7	13.48	12.10	12.2	12.22	12.15
–3 dB Angle	60°	59°	58°	57°	56°	55°

1.31. Feeding the Eight Circle

Remember how easy it was to feed the 2-element end-fire array? This is just as easy because we have here two end-fire array groups (see Fig 7-27) fed in-phase. The switching too becomes quite easy (**Fig 7-55**). The system uses 32 small SPST reed relays set up in a matrix (see **Fig 7-56**), at least if we want to use 0° hybrid combiners to combine the signals rather than simple parallel connection (see Section 1.22).

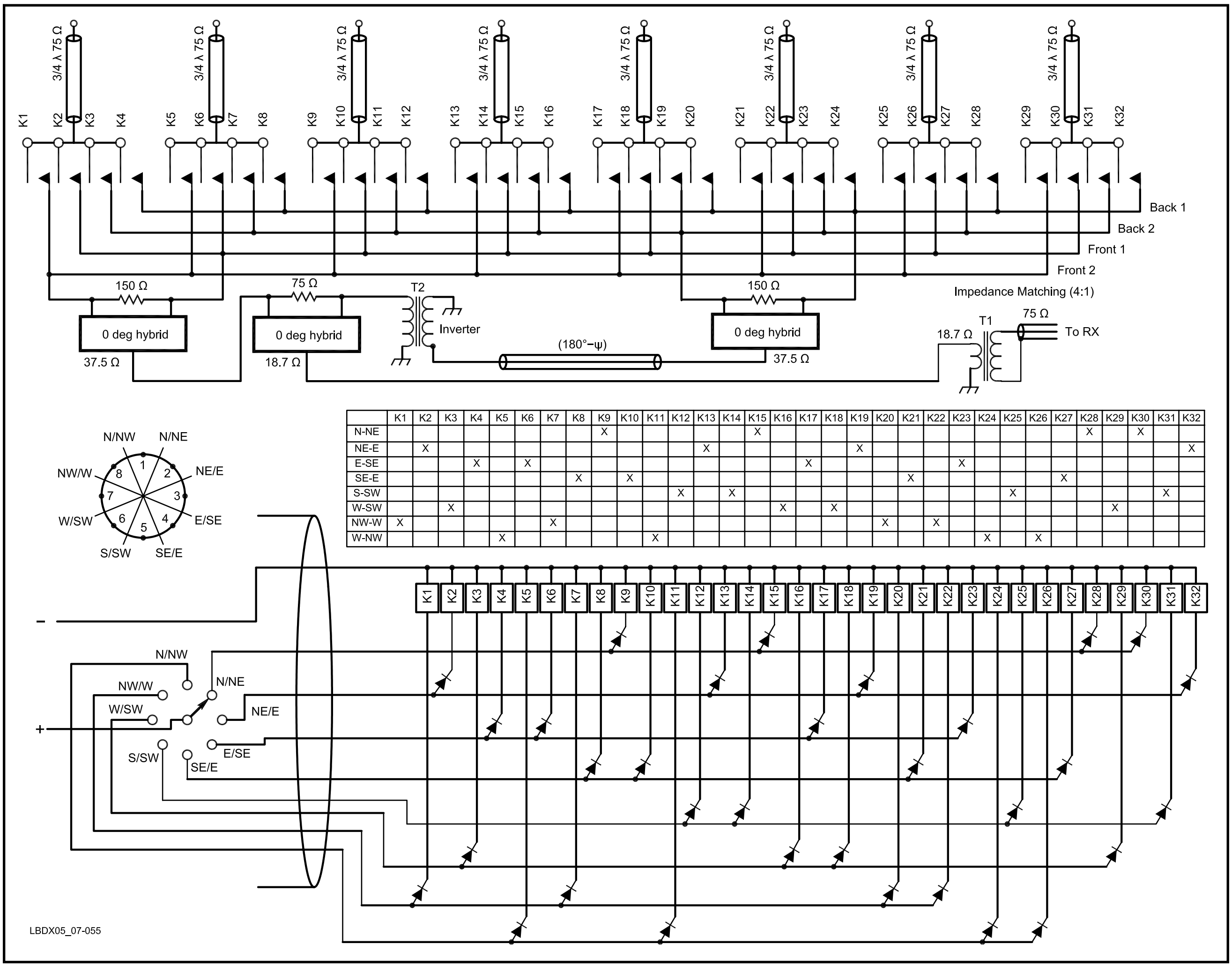
If the return loss on the feed lines going to the antenna elements is ≥23 dB (SWR ≤1.15:1), these feed lines can be any length, as long as they are all equally long. The rule that applies to elements of a transmit array (use λ/4 long feed lines) does not apply here, as the SWR is low enough to ensure that phase delay equals cable length (both expressed in degrees).

The two front verticals are fed at 0° phase angle. The feed lines to these arrays are combined via the 0° hybrid combiner (see Fig 7-24). The feed lines to the back elements are also combined via a second hybrid combiner. This output goes via the phasing line to the 1:1 phase inverter T2 (see Fig 7-25). A third hybrid combiner now combines the signals coming from the front elements and the back element. Its output impedance is 18.75 Ω. T1 is a 4:1 impedance ratio transformer (see Fig 7-30) matching this impedance to the 75-Ω feed line.

If you want to experiment with various null angles, you could easily add a few relays and change the length of the delay line from 50° (gives 180° – 50° = 130° phase shift), over 60° (120° phase shift) to 70° (110° phase shift). By switching between these lines it is possible to position the nulls in the back of the array at various angles in both the horizontal and vertical planes.

1.32. The Eight Circle in Practice

As far as I know, there are still only a few Eight Circle arrays around. That's not surprising because you need more



	K1	K2	K3	K4	K5	K6	K7	K8	K9	K10	K11	K12	K13	K14	K15	K16	K17	K18	K19	K20	K21	K22	K23	K24	K25	K26	K27	K28	K29	K30	K31	K32
N-NE									X						X													X		X		
NE-E		X											X						X													X
E-SE				X		X											X						X									
SE-E								X		X											X						X					
S-SW												X		X											X						X	
W-SW			X													X		X											X			
NW-W	X						X													X		X										
W-NW					X						X													X		X						

Fig 7-55 — If we want to use 0° hybrid combiners, we need 32 small SPST reed relays to switch the Eight Circle in eight different directions. The remote control box uses a single-pole 8-position rotary switch, and an 8+1 conductor cable is required. You can reduce this number of lines if you use a few more diodes and positive as well as negative voltage to switch the relays. See text for details.

Fig 7-56 — Receiving antenna switching matrix (8 × 4 lines = 32 SPST relays) using low-cost hermetic reed relays. Note also the 90-V gas discharge tubes and the static drain resistors (56 kΩ). This relay matrix is made by Charlie, NØTT and used at his Nine Circle array (see Section 1.33)

than 4 acres of open land to put one up. Tom, W8JI was the "inventor" of the array, and the second one (to my knowledge) was put up by Wally W8LRL. Like W8JI, W8LRL has lots for acres for receiving antennas, and he finds the Eight Circle a good complement to the extended range of Beverages he runs on his very impressive 160-meter antenna farm. Joel, W5ZN, and Bob, N4HY detailed the design and construction of W5ZN's Eight Circle in March/April 2010 *QEX*.

It's proven over and over again, that the most successful Top Band DXers are those with the best receiving antennas in the quietest locations!

1.33. The Nine Circle Receiving Array

To my knowledge, the Nine Circle transmitting array, described in Chapter 11, conceived by John, WØUN, was first built by K9DX. A few hams have built a receive only version of the array, using short resistor-loaded elements as described in Section 1.21.1. One of them was Charlie, NØTT (see **Fig 7-57**). The feed system described hereafter is not the feed system used by NØTT.

In Chapter 11, Section 4.13 we can find a full description of the transmit version of the array. The receive-only version can be made with a much simpler feed system. For a transmit antenna we normally specify the antenna currents (magnitude and phase). The Nine Circle not only has different feed current phase angles, but also different feed current magnitudes. We will use these values (see table insert in

Fig 7-57 — To prevent having to remove the grass and weeds that cover the base of verticals, Charlie, NØTT, slipped a disk made of roofing shingles over the wooden support poles at his Nine Circle array. Nice and tidy.

Fig 7-58 — This schematic shows both the impedance and the current values in different points of the circuit. We have started from the equivalent transmit array, taking into account the feed currents (magnitudes and phase). In this case we can make use of attenuators to adjust signal levels in the different branches to the levels required to obtain the proper signal combination in the combiner. Phase inversion (0° to 180° and 90° to –90°) is done with phase inverter transformers T1 and T2. The 90° shift in the branch going to elements 6, 7, 2 and 3 is obtained by using a 90° long coax. All signal combining is done in 0° hybrid combiners (H1 through H8).

Transmit Equivalent		
Feed Current		
EL	I (A)	θ (°)
1 and 8	1	180
2 and 3	1.66	−90
4 and 5	1	0
6 and 7	1.66	90
9	3	0

Z_0=75 Ω		
att (dB)	Rpar	Rser
4.4	302 Ω	40 Ω
3.9	340 Ω	34 Ω

Attenuators
4.4 = 20 log (3.32/2)
3.9 = 20 log (4.7/3)

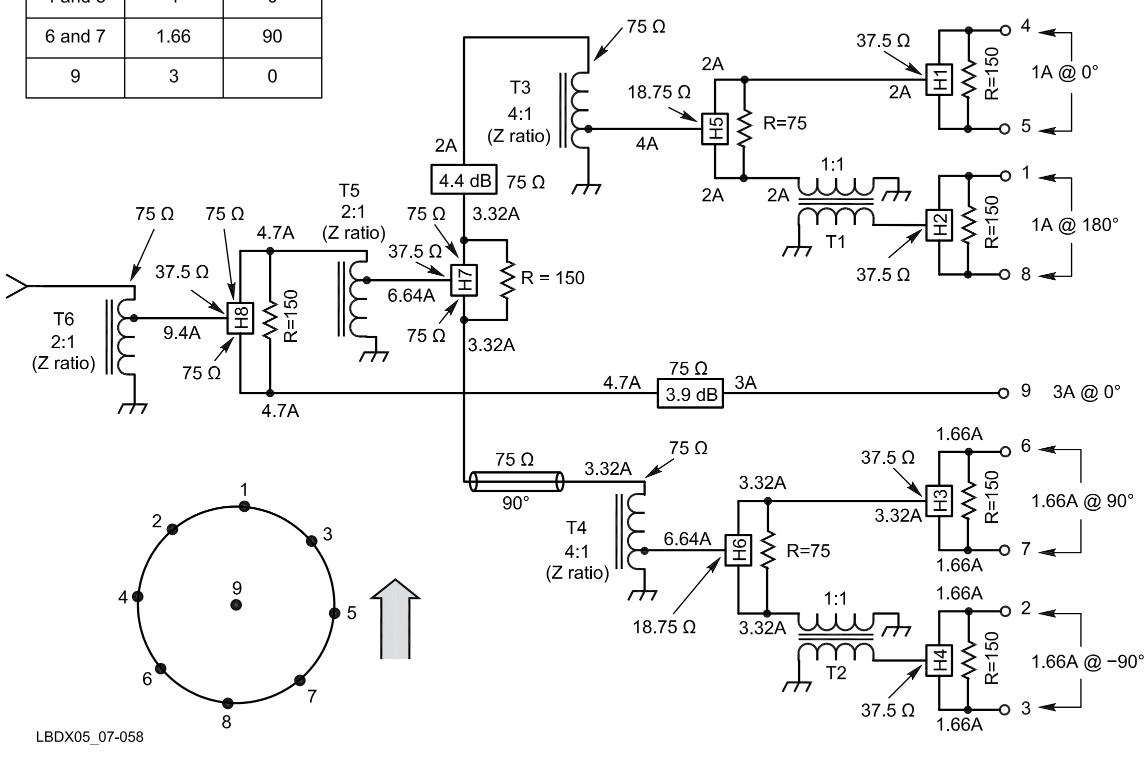


Fig 7-58) to design the combiner network.

The Nine Circle's DMF is a spectacular 33.5 dB and the MDF is 12.63 dB. See modeling file *ch7-9CIRCLE-RX-ant.ez* on the CD.

Fig 7-58 shows an elegant feed system where the difference in concept between the transmit array and the receiving array is highlighted (the use of attenuators). The feed lines going from the "combiner" box to the antennas are of equal length (there is no need for them to be λ/4 or λ/4 long). The combiner box can best be located at the central element. In this case all elements can be fed with equal length (approximately 65 meter) cables, and no "surplus" lengths of cable need to be coiled up, except for the center element where the total length will need to be coiled up.

The antenna elements are loaded to be resonant and have a feed point impedance of 75 Ω (that includes approximately 70 Ω resistance loading). This means that the SWR on these lines is very low (very little mutual coupling between the elements). In the combiner box we will combine the signals from these elements (taking into account the required magnitude and phase) which will result in the wanted directivity pattern. There are different ways of doing it, but the system shown in Fig 7-58 has some advantages. It uses the 0° hybrids (see Section 1.22 explaining why these should be used in receiving antenna combiners) and has the distinct possibility to fine tune the voltage magnitude (to compensate for possible unequal losses in different branches of the network) by using simple resistive attenuators to adjust the signal levels as required.

In the branch going to elements 2, 3, 6 and 7 we need to introduce a 90° phase shift. This is done by inserting a 90° feed line in that branch. Transformers T3 and T4 are 4:1 (Z ratio) transformers providing 75 Ω in the line where we insert the 90° phasing line. The line from T3 to H7 incorporates an attenuator (6.5 dB), required to obtain the same signal magnitude to the combiner H7 as coming from the branch with 1.66 A feed current. Between H8 and H7 we need to place a 2:1 impedance ratio transformer (1.41:1 voltage ratio) to provide a correct impedance match between the two combiners. This transformer (T5) reduces the current from 6.64 A to 4.7 A (ratio is 1.41:1). The values of the resistors used in the π-type attenuators are shown in Fig 7-58. It is obvious that this approach, using attenuators, cannot be used in transmitting arrays.

The hybrid combiners are described in Section 1.22 and Fig 7-24, the phase inverters (T1 and T2) in Fig 7-25 and the matching transformers T5 and T6 in Fig 7-26.

Checking and adjusting the combiner box is quite simple. Terminate all except one of the nine input ports (which for the test become "output" ports) with quality 75-Ω resistors. Use a vector network analyzer (VNA) calibrated for 75-Ω impedance (see Chapter 11, Sections 3.5.2.3 and 3.5.2.4) to drive the circuit (the terminal which in normal use is the "output" terminal), and check one by one the outputs of the combiner box. The relative voltage outputs at these ports should be the same as current values shown in the table which is part of Fig 7-58.

If a VNA is not available (…get one…), an antenna analyzer such as the MFJ-259B (or even your transmitter running QRP) can be used as a signal generator, while using a dual (or multiple) trace oscilloscope to check the various phases and RF voltages at all the "output" branches. The precision of this method is, of course, not comparable with what can be obtained using a VNA.

Switching directions is done using an 8 × 8 relay matrix (yes, 64 small reed relays, but they are inexpensive). The entire array, including the diode matrix can be built on two stacked perf boards measuring 10 cm × 10 cm, one containing 64 small reed relays, the other one 64 diodes.

If we did not use the hybrid combiners (H1 through H4) but had simply paralleled the lines to the "pairs" of elements, we would have come up with an 8 × 4 matrix, but that is *not* the best solution. The design of the matrix is shown in Figs 7-55 and 7-56.

1.34. Grounding the Feed Lines

In all of the feed systems, we have used impedance matching transformers (2:1 or 4:1 impedance ratio) with galvanically separated windings which helps suppress the ingress of common-mode signals that can be present on the feed lines. In Section 2.7.2.8. it is explained that the feed line should *not* be grounded to the ground system of the antenna, but that the feed line can be grounded to a separate ground rod at least 5 meters away from the Beverage antenna ground rod.

In most cases, a simple ground rod serves as antenna ground with a Beverage (assuming a ground rod can be driven into the ground). In arrays using short vertical elements, the ground system is usually much more elaborate, not just to provide a "low" ground resistance (we actually need a high loss resistance which we obtain by using a series connected resistor). Primarily we need to achieve a stable ground resistance (stable during varying seasons). That's why we normally use some kind of radial system on these short loaded vertical antennas.

If we would ground the feed line of a receiving array somewhere near and/or between radials, this would likely be a bad idea. Therefore it is better *not* to ground the feed lines in the area where the radials are buried. Stay at least 5 meters away from the area where the radials are buried. In most cases, especially when you have the feed lines buried (mine are buried 60 cm deep under the radial wires), common-mode signals will be heavily choked to the ground if they are present at all.

At the station it is recommended to bring in all antenna cables and ground them at a single large metal entrance panel that is connected to an excellent RF/lightning ground, and also,

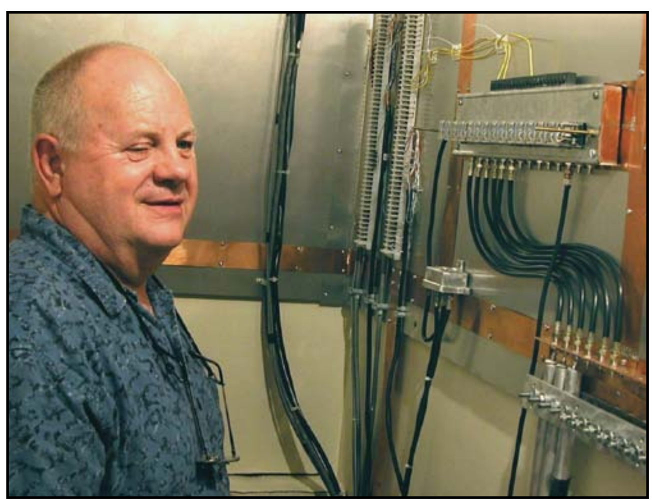


Fig 7-59 — Wally, W8LRL and his antenna cable entrance panel.

for safety reasons, to the electrical safety ground. This SPG (single point ground) should consist of several ground rods in the form of a grounding grid, connected with heavy solid copper strap to the panel. The panel should be made of a solid copper plate, but aluminum can do the job also (see **Fig 7-59**). It is highly recommended to have each coaxial cable go through a suitable surge protector, mounted on the entrance panel.

1.35. Parasitic Receiving Arrays

Why don't we see small parasitic receiving arrays? Can't we make our transmit arrays be excellent receiving arrays? Transmit arrays can be designed to have excellent directivity, but as a rule they are very large. For a transmit antenna, efficiency is a major concern, unlike the case for receiving antennas.

Parasitic arrays require a high Q to work well. Mutual coupling is essential, since the current in the parasitic element is obtained *only* through mutual coupling. Parasitic arrays, whether intended for receiving or transmitting, must be designed to have the lowest possible losses.

Do we need full-sized (λ/4) elements to have low-loss, high-Q elements? The answer depends on how good the ground system is. In other words, with a nearly perfect ground system (perfect in terms of near-field, where return currents are collected) shortened elements can be used. Jim, N7JW, and Al, K7CA, have designed various transmitting parasitic arrays for 160 meters (**Fig 7-60**) that use 16.45-meter long elements, top loaded with two drooping wires (see **Fig 7-61**). This element is self-resonant slightly higher than 1.83 MHz, and has a radiation resistance of about 11.5 Ω. In the N7JW/K7CA transmitting arrays, a "nearly perfect" ground radial system is used, consisting of 120 40-meter long radials. The equivalent loss resistance is on the order of 1 Ω. This brings the feed-point impedance of the element to approximately 12.5 Ω.

In the array the element is tuned exactly to resonance on 1.83 MHz with a small coil at the base (approximately 0.1 μH). All the elements are made identical. Fig 7-61 shows the element configuration. When the relay is energized, the bottom of the

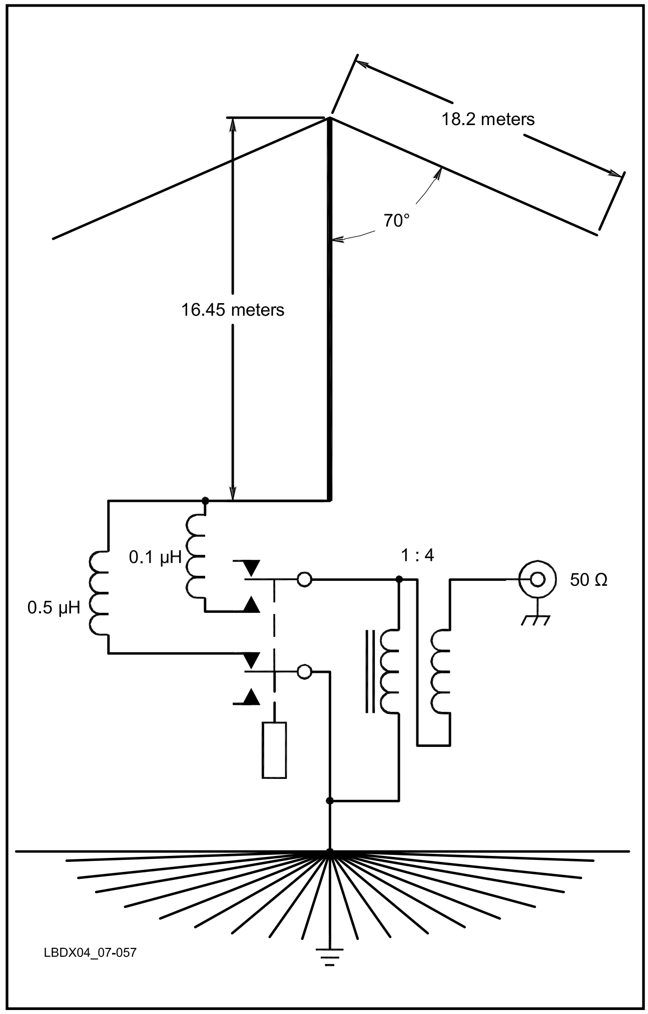


Fig 7-61 Basic element for the N7JW/K7CA parasitic array. If the array will be used for transmit, a regular DPDT relay can be used. For a receive-only array, two small reed relays are recommended to do the switching.

Fig 7-60 — At N7JW's QTH in St George, Utah, we see the eight elements of his Eight Circle array on top of a ridge. This array consists of two side-by-side 2-element parasitic vertical groups, spaced approximately 94 meters. Similar to W8JI's Eight Circle in concept, it is switchable in eight directions. The main difference from W8JI's Eight Circle is that N7JW's antenna is a transmit array as well.

Table 7-17
Performance Data for 2- and 4-Element Parasitic Arrays

Configuration	*Gain (dBi)*	*3 dB Angle (degrees)*	*RDF (dB)*	*DMF (dB)*
2 element	4.61	147°	8.73	14.43
4 element	8.24	59°	12.1	22.5

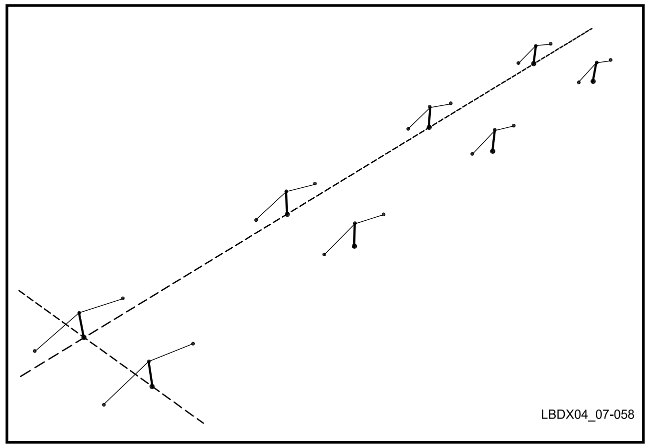


Fig 7-62 — The 8-element parasitic array at N7JW/K7CA consists of four 2-element parasitic cells spaced just over λ/2 apart.

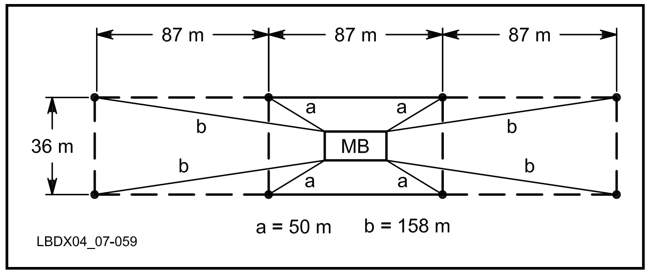


Fig 7-63 — The center elements are fed with cables of equal length to the central matchbox (MB). The feed lines to the outer elements are 1 λ longer (108 meters for RG-213) to keep the four cells fed in phase.

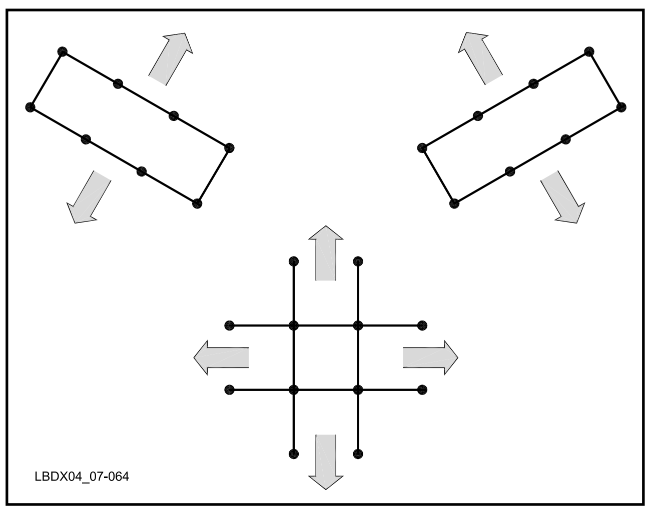


Fig 7-64 — Directions covered by the four 8-element arrays at the remote QTH of N7JW/K7CA.

element is connected to ground via a coil of approximately 0.5 μH, which turns it into a reflector (tuning the element to a self resonance of approximately 1812 kHz). With no voltage applied to the relay, the base of the element is resonated to 1.83 MHz with a coil of approximately 0.1 μH and connected to a 4:1 wide-band transformer.

The basic idea of the transmitting array is the same as the concept of an Eight Circle, where eight elements are on a circle. Just like in the Eight Circle receiving array described in Section 1.32, four elements are active at a time (see Fig 7-54). The circle diameter is 94 meters, and the elements are spaced 36 meters in the circle. This makes for two broadside 2-element arrays, spaced 86.9 meters, while the elements of each cell are spaced 36 meters (0.2 λ).

The parasitic element in each cell is a reflector. The performance data for two cells in a broadside array are given in **Table 7-17** (modeling files *ch7-n7jw-1cel.ez* and *ch7-n7jw-2cel.ez*). How does this N7JW/K7CA Eight Circle (transmitting) array compare to the W8JI Eight Circle receiving array in Section 1.30?

- The directivity (both RDF and DMF) are not quite as good as for the all-driven array but they are still excellent.
- The parasitic array requires a "perfect" ground system.
- The parasitic array requires longer vertical elements (16.5 meters vs 6 meters).
- The parasitic array has a gain of +8.2 dBi vs –8.5 dBi for the all-fed receiving array. This makes it an excellent transmitting array, with 2.5 to 3 dB more gain than the well-known transmitting Four Square.

N7JW and K7CA went one step further and also built 160-meter receiving arrays that use four broadside 2-element cells. Four such 8-element arrays were built in the middle of the desert, at a remote station some 35 miles from their homes just west of Zion National Park, north of St. George, Utah. These are linked to the home QTH via VHF and UHF. These 8-element arrays are mostly used for receiving only, as the remote station in the desert is powered by solar cells. To transmit from there, in a contest for example, a suitable generator has to be brought out.

With a length of 261 meters and a width of 36 meters, the 8-element array has a very large footprint (see **Figs 7-62** through **7-64**)! A 6-element array could also be made that would measure "only" 174 meters wide. And by the way, all these dimensions are without the radials extending 40 meters in all directions! You need half a desert to put up such an antenna. And that is the case at the remote QTH of N7JW (see **Figs 7-65** and **7-66**).

The performance of these receiving arrays is quite spectacular, as shown in **Table 7-18** and **Fig 7-67** (modeling files *ch7-n7jw-3cel.ez* ***and*** *ch7-n7jw-4cel.ez*). With an RDF of 15.12 dB the 8-element array is 1.65 dB better than the all-fed transmitting Eight Circle (see Section 1-30). Note however that this is obtained through an extremely narrow forward lobe, where the –3-dB angle is only 32°.

In the 4-element array, both driven elements are fed with the same current. If you use two equal-length 50-Ω feed lines going to the centrally located matching and switching box, the combined impedance will be 25 Ω. This can be matched to the

Fig 7-65 — Breathtaking aerial view of two of the 8-element arrays at N7JW's remote station.

Fig 7-66 — View of seven of the eight elements of one of the four N7JW/K7CA arrays at their remote antenna farm in Utah desert.

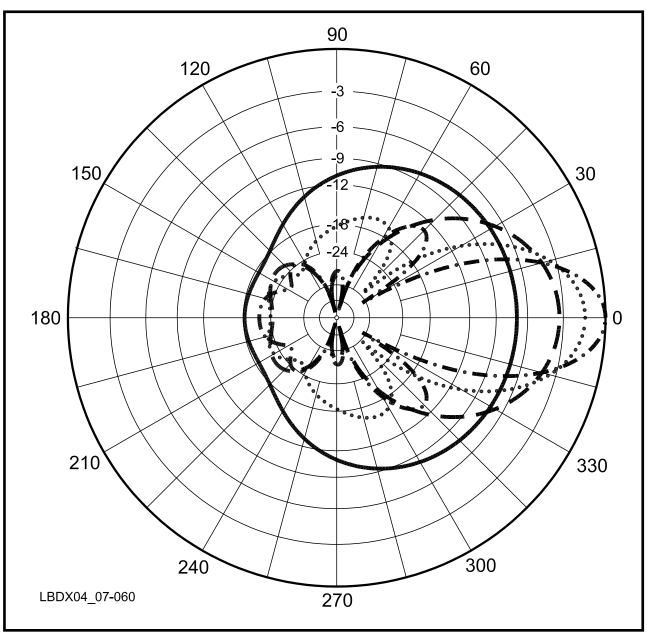


Fig 7-67 — Azimuth patterns for the array in its various configurations. Solid line = 2 elements; dashed line = 4 elements; dotted line = 6 elements; dashed-dotted line = 8 elements. For gain, nose beamwidth and directivity figures, see Table 7-22.

feed line by any convenient means, such as a 2:1 transformer, L-network or a 37.5-Ω quarter-wave transformer.

For the 8-element array, consisting of four cells, the best directivity results from feeding the outer driven element with 70 to 80% of the current value used for the two central driven elements. This is automatically obtained by using a "lossy long line" to these elements. As all four elements need to be fed in-phase, you would feed the outer elements with feed lines that are 1 λ longer than the center driven elements. This is 108 meters of RG-213 (VF = 0.66), which has a loss of approximately 0.9 dB. This means a current ratio of 1:1.11 or 90%, not exactly what we need but close enough. A coax with a little more loss would be even better. The same applies for the 6-element array, where you should feed the two outer

Table 7-18
Performance Data for 6- and 8-Element Parasitic Arrays

Configuration	*Gain (dBi)*	*3 dB Angle (degrees)*	*RDF (dB)*	*DMF (dB)*
6 element	9.65	42°	13.54	25.52
8 element	11.2	30°	15.12	27.80

Fig 7-68 — K7CA at the base of one of the array elements. Note the heavy copper ring to which all 120 radials are soldered. The two loading coils are wound on the black insulator, while the relay and the 1:4 transformer are placed in a plastic enclosure.

Table 7-19
Operational Bandwidth of 8-Element Array

Frequency (kHz)	*Gain (dBi)*	*RDF (dB)*	*DMF (dB)*
1810	10.59	14.86	20.31
1820	11.03	15.16	24.90
1830	11.20	15.12	27.80
1840	11.11	14.84	24.29
1850	10.87	14.47	20.92

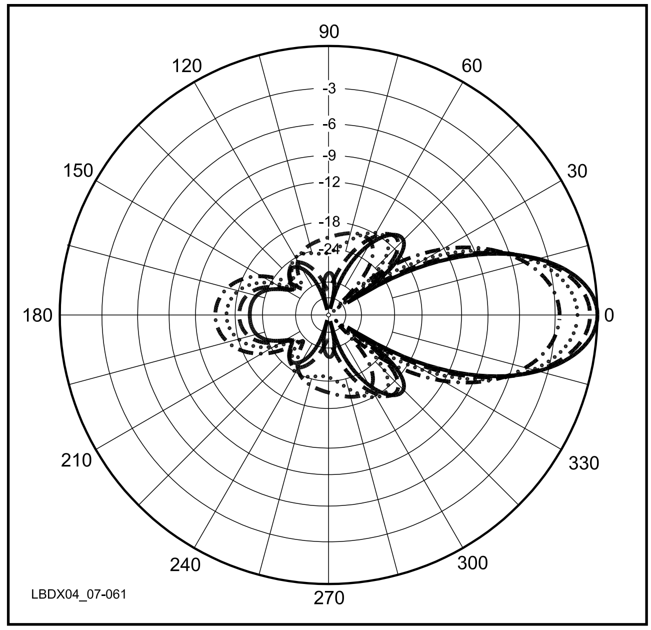


Fig 7-69 — Azimuth patterns for the 8-element receiving array at various elevation angles. Solid line = 10°; dashed line = 20°; dotted line = 30°; dashed-dotted line = 40°.

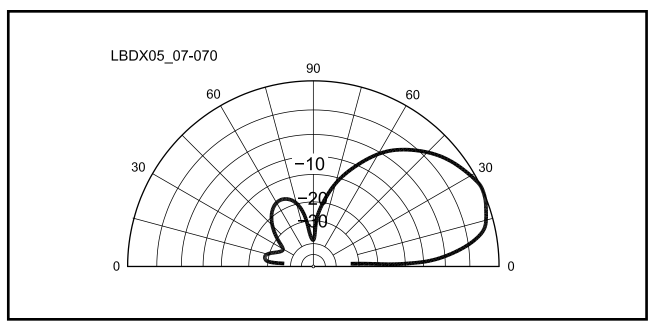


Fig 7-70 — Vertical radiation pattern of the N7JW/K7CA arrays.

elements with approximately 80% of the feed current of the center element.

The 6-element array requires the same current tapering across the cells (1:2:1) for best directivity. In the centrally located box for matching and switching, the feed lines to one side of the array are connected in parallel, as well as those to the other side. A simple relay switches directions by selecting which side to feed. At the same time, the elements on the back of the array are turned into reflectors by energizing the relays at their base (see Fig 7-61 and **Fig 7-68**). A 4:1 wideband transformer (identical to the one used at the feed point of each element) is used to transform the 12.5 (four 50-Ω lines in parallel) back up to 50 Ω.

The directivity of this array is excellent over a wide range of elevation angles, as shown in **Fig 7-69**. Contrary to what you might expect, the operational bandwidth is also excellent, as shown in **Table 7-19**. The SWR bandwidth is very flat from just below 1820 to well over 1850 kHz, and the directivity figures remain remarkably high over about 50 kHz as well.

Jim and Al have Beverages up there as well, but they swear by the 8-element arrays. They claim the arrays have better performance due to the cleaner vertical pattern, and to the fact that the elevation angle (at 25° to 30°) is substantially higher than for long (300-meter minimum) Beverages. See **Fig 7-70**.

The 8-element array gets its directivity from the very narrow forward lobe. From Utah, looking into Europe, the back is California and the Pacific. Very little 160 meter activity originates from there, at least as compared to Europe. In Belgium, I have the entire continent of Central and Eastern Europe behind me, and I need my directivity mainly in the back, a very different situation. This goes to illustrate that what is best in one situation is not necessarily so in another one.

Going against the mainstream of using separate receiving and transmitting antennas, these high performance TX/RX arrays have proven to be a winner for N7JW/K7CA, who are consistently either the loudest or the only stations heard in Europe from that part of the world. And it works on reception too, since I never have to call them twice.

1.36. Vertical Receiving Arrays Compared

Table 7-20 gives an overview of the characteristics of a single element (see Section 1.21) in the various receiving arrays examined so far. To obtain reasonable bandwidth the R_{rad} of an element should be kept low enough compared to the total equivalent series loss resistance in the feed impedance. If we use a 75-Ω system impedance, we generally will require elements with R_{rad} <2 Ω to provide adequate bandwidth. Elements with a higher R_{rad} will need to be loaded with a higher series resistance, as was done in Section 1.29. In **Table 7-21** the DMF and RDF are shown vs the forward lobe at a 20° elevation angle for a variety of arrays.

Note that the Eight Circle and the Nine Circle have similar RDF performances, but quite different DMF figures. The Eight Circle has a narrower forward lobe (due to the wide broadside spacing of the two groups) while the Nine Circle has a much better DMF, which means a better reception of signals in the entire back area of the array.

Table 7-20
Characteristics of a Single Element

Single element gain (at 20° angle) and R_{rad} (over average ground).
Gain calculated with a series resistance totaling 75 Ω impedance

Element Type	*Reference Section*	*80 m Gain (dBi)*	*160 m Gain (dBi)*	*80 m R_{rad} (Ω)*	*160 m R_{rad} (Ω)*
W8JI	1.21.1	–12	–17.4	5.3	1.2
K8BHZ	1.21.2	–10.1	–15.5	7.6	1.7
6 meter, base-loaded	1.21.3	–14.8	–19.9	2.2	0.5
9 meter, base-loaded	1.21.3	–11.1	–16.4	5.1	1.2
12 meter, base-loaded	1.21.3	–8.3	–13.9	9.8	2.2

Table 7-21
DMF and RDF for a Variety of Arrays

Antenna Description	*DMF (dB)*	*RDF (dB)*	*3 dB Angle (degrees)*	*Output* (dBi)*	*Reference Section*
2 el EF λ/4 spacing, ψ = 135°	17.22	9.57	132	–12	1.8
2 el EF λ/8 spacing, ψ = 155°	18.56	10.01	120	–16	1.14
2 el EF λ/16 spacing, ψ = 165°	18.47	9.92	121	–21	1.14
4-square side λ/4, ψ = 120°	25.51	11.49	86	–8	1.26
4-square side λ/8, ψ = 140°	25.12	11.24	86	–15	1.26
4-square side λ/16, ψ = 160°	25.88	11.27	85	–23	1.26
6-el Stone-Hex 305 m diameter	27.4	11.7	79	–19	1.29
8-circle ψ = 120°, diameter 0.594 λ	22.6	12.3	55	–8	1.30
9-circle ψ = 90°, diameter 0.77 λ	31.7	12.6	60	–7.5	1.33
Broadside 2 el bidirectional spacing = 0.61 λ	9.5	9.5	52	–10.5	1.11.1
Broadside 4 el bidirectional spacing = 0.69 λ, 6 dB feed current taper	12.83	12.83	25	–7.9	1.11.2
Broadside / end-fire 2 el λ/4 end-fire spacing, ψ = 105°	23.45	12.4	57	–4.7	1.12
Broadside / end-fire 8 el spacing = 0.69 λ, 6 dB feed current taper	29.35	16.06	24.5	+0.7	1.13

*Gain (dBi) on 1.8 MHz over average ground, using 12 meter long, non-top-loaded elements. Bottom loaded to achieve 75 Ω impedance.

Table 7-22
Performance Data for the N7JW/K7CA Arrays

Antenna description	*DMF (dB)*	*RDF (dB)*	*3-dB Angle (degrees)*	*Gain vs 1 el (dB)*
4-el parasitic array (N7JW style)	22.5	12.1	59	+7.1
8-el parasitic array (N7JW style)	30.0	15.1	30	+10.2

Note: these are also used as transmit antennas. See Section 1.35.

1.36.1. Gain Corrections for Other Types of Array Elements

For other elements, the gain can be corrected using the gain figures from Table 7-20. If you use, for example, 6-meter instead of 12-meter elements, the gain will be 19.9 – 13.9 = 6 dB lower than what is shown in the table. If you use the W8JI-style elements you have to subtract 17.4 – 13.9 = 3.5 dB.

1.36.2. Transmitting Arrays with Outstanding Receiving Performance

While these antennas are not in a strict sense receiving antennas only, I will list their performance figures here in **Table 7-22**. The N7JW/K7CA array is covered in this chapter; the WØUN/K9DX Nine Circle is covered in Chapter 11.

1.37. Using a Noise-Canceling Bridge to Feed a Receiving Array

In Sections 1.4 and 1.5, I briefly touched on the subject of noise-canceling devices. A noise-canceling device is nothing but a phasing/combining system, in which you combine the output of two antennas and adjust the relative amplitude and phase for full cancellation *of a particular signal*.

The feed system for a receiving array can be seen as a noise-canceling bridge. For two signals to produce zero output when combined, they must be the same amplitude and 180° out-of-phase. In the receiving arrays described so far, identical elements were used (same polarization, same directivity pattern), close together (less than 1 or 2 λ apart to avoid space diversity effects). These are prerequisites for signals with identical magnitudes, when you consider signals (QRM, noise, etc) received via skywave. If you use different antennas to feed a noise bridge, the varying polarization and signal strength will make it impossible to get a stable null on skywave signals. As W8JI stated: "Despite folklore... mixing a horizontal antenna with a vertical is real trouble on skywave circuits. Fading, averaged over a short time, actually increases…"

The main difference between a directive array and a noise-canceling setup can be summarized as follows:

- A directional array uses identical antenna elements in the array.
- A directional array is designed to provide directivity, not to null out a specific noise source. (The array has a narrow forward lobe in which we receive, and it suppresses signals as much as possible in all other directions, to increase RDF and DMF.)
- A noise-canceling setup is intended to reduce/eliminate locally generated noise.
- A noise-canceling setup normally uses a small-size noise-sensing antenna located near the noise source.

Instead of designing the feed system of a directive array for a fixed (optimized) phasing angle to produce the best possible DMF or RDF, we can bring the individual feed lines of the end-fire array and do the phasing "in-house" using a phasor-combiner, usually called a *noise bridge*. The MFJ-1025 and 1026 are commercial units that perform very well on the low bands. The advantage of such a setup is that you can move the null of your array continuously around from inside the shack. They are great for nulling out a particular QRM source.

However, noise usually comes from many directions. The DMF or RDF of your receiving setup is really much more important than being able to null out a signal from one specific direction. In addition, moving a null by varying the phase not only moves it in the horizontal plane, but also moves it at the same time in the vertical plane (see Section 1.6).

A noise-bridge is really the best instrument if you want to cancel out QRM received from nearby sources on ground wave. Still better, of course, is to kill the noise source itself. When used for such an application, any small vertically polarized sensing antenna can be used. The noise-sensing and receiving antennas can be very dissimilar, so long as the noise antenna hears the noise well. Ideally, the noise antenna should hear only the unwanted signal. Since both antennas receive the noise by ground wave, there are no variations as we normally experience on skywave signals.

W8JI has covered the MFJ-1025/1026 noise canceller in great detail on his Web page. Tom describes some modifications that can be made to improve its efficiency on 160 and 80 meters (**www.w8ji.com/mfj-1025_1026.htm**). More recently Tom, W8JI designed his own noise canceller/phase controller which is available from DX Engineering (model NCC-1, **Fig 7-71**).

These units can also be used to enhance received signals. W8JI, wrote on his Web site: "Enhancing signals requires both antennas to have similar S/N ratios, and ideally they would have similar patterns. That's true if it is a Beverage and a Four Square, K9AY loops or Flags, two regular (so-called magnetic) loops, or whatever being mixed. Vertically polarized antennas mix well with other vertically polarized antennas. Since my Four Square has a similar pattern to a pair of echelon-staggered Beverages, they mix almost perfectly! Remember if you combine one poor S/N antenna with a good S/N antenna in an effort to enhance signals, you will almost certainly make the good antenna become worse! Mixing in a poorer S/N antenna is great for nulling local noise, but not peaking distant signals. I mix three Beverages with a Four Square, using any combo that works best. Sometimes the improvement is 15 dB or more. Sometimes it won't help, but mixing anything vertically polarized with my dipoles is almost always a real bust for signal enhancement."

Mauri, I4JMY, who made his own noise-canceller, wrote on the Top Band reflector: "The null steerers are top performers to avoid overload by local huge signals on the same or another band, such as a nearby contester running a pileup or someone ragchewing even a few kHz away."

This very true, and if you have a local station (within a few kilometers) on 160 meters running a kW, it's likely he'll be S9 + 40 dB or better, and you may hear him all over the band. The noise canceller can literally kill his signal!

I have successfully used the MFJ noise canceller for getting rid of local ground wave noise. I use my transmitting vertical as the noise antenna, with a attenuator so that only the local noise was heard, while the main antenna was one of the Beverages. This worked very well. The built-in small 1-foot

Fig 7-71 — W8JI developed the NCC-1 noise canceller/ phase controller, which is available from DX Engineering.

long sensing whip is really not much of a sense antenna, and I never could make good use of it.

It takes some understanding of what happens inside the black box before you can adjust the noise canceller properly. Dave, NR1DX, wrote: "I first note the S-meter reading of the noise with the 'main antenna gain' all the way up and the noise antenna gain all the way down. I then turn main control to zero. I turn up the gain of the noise antenna until I get the same S-meter reading. Next I turn the main antenna gain all the way back up and adjust the phasing control for a null. Careful tweaking of the noise gain and the phasing control then tames the beast. If I can't get an equal level through the noise-antenna side, I either live with the noise or try a different antenna as the noise antenna. Successful noise nulling is an art and it takes a good noise antenna plus patience to tune out a noise source as the controls can be quite sharp."

2. AN INTRODUCTION TO BEVERAGE ANTENNAS

2.1. The Beverage Antenna: Some History

The Beverage antenna (named after Harold Beverage, W2BML) made history in 1921. In fact, a Beverage antenna was used in the first transatlantic tests on approximately 1.2 MHz. For many decades, the Beverage antenna wasn't used very much by hams, but in the last 30 years it has gained tremendous popularity with low-band DXers. The early articles on the Beverage antenna (Ref 1200-1204) are excellent reading material for those who want to familiarize themselves with this unique antenna.

A revealing interview with Dr. Harold H. Beverage and H.O. Peterson (interview by Norval Dwyer, done in 1968 and 1973) can be read at: **www.hard-core-dx.com/nordicdx/antenna/wire/beverage/interview1.html**.

2.2. Beverage Antenna Principles

Fig 7-72 shows the basic configuration of the *Beverage* antenna (also called the *wave antenna*). It consists of a long wire (typically 1 to several wavelengths long) erected at a low height above the ground. The Beverage antenna has very useful directional properties for an antenna so close to the ground, but

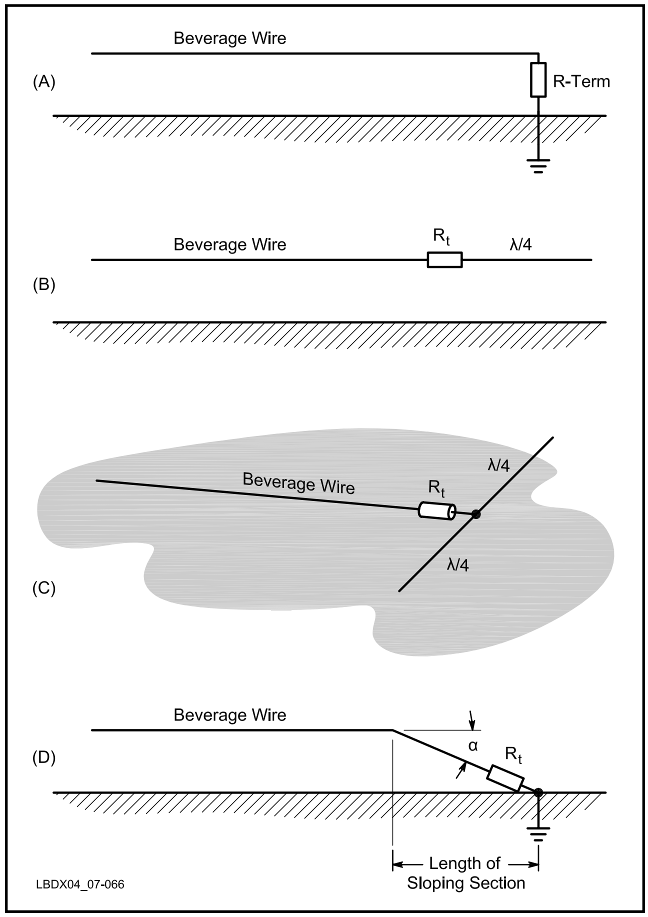


Fig 7-73 — Different terminating systems for Beverage antennas. The version at A suffers from stray pickup because of the vertical down lead. At C, two in-line quarter-wave lines terminate the Beverage, by which the radiation from these lines is effectively canceled. This is not a very practical solution because of its area requirements. This configuration is used throughout the book for modeling Beverage antennas, however. At D, the method most widely used with real-world Beverages. Using a long sloping section is thought by many to reduce omnidirectional pickup.

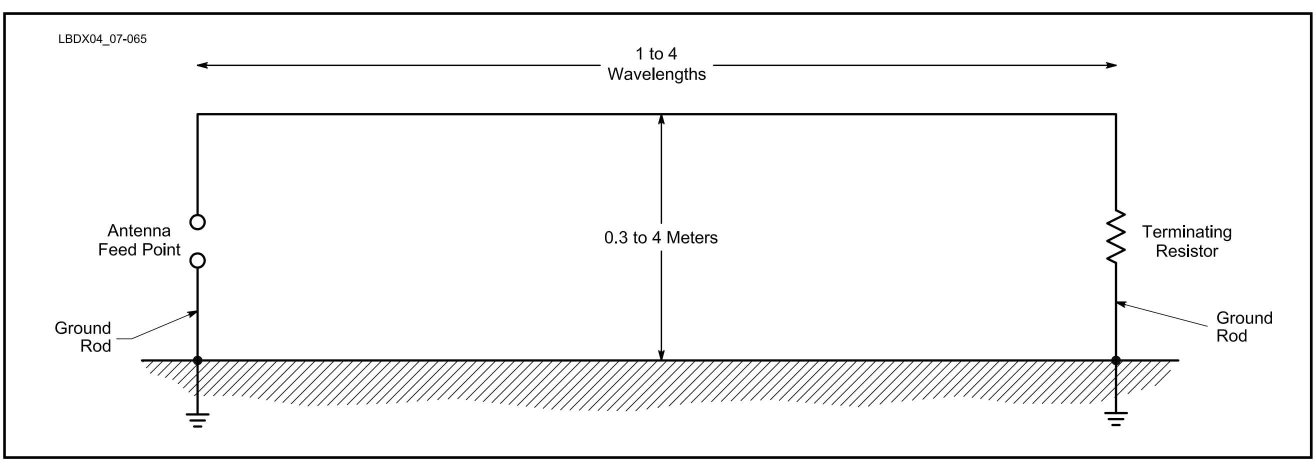


Fig 7-72 — The Beverage antenna is a straight wire, typically 1 to 4 λ long, mounted parallel to the ground at a height of 0.01 to 0.03 λ.

it is relatively inefficient as far as signal output is concerned. This is why the antenna is primarily used only for reception on the amateur low bands. The Beverage receives best off the end where the terminating resistor is located. Looking at Fig 7-72, the Beverage antenna favors signals coming from the right (coming from the end where the terminating resistance is located). Note that with small loop antennas (see Section 3) it is the other way around!

The Beverage antenna can be thought of as an open-wire transmission line with the ground as one conductor and the antenna wire as the other. To achieve a unidirectional pattern, the antenna must be terminated at the far end in a resistor equal to the surge impedance of the antenna. See **Fig 7-73**. So-called *bidirectional Beverages* are covered in Section 2-13.

If the Beverage antenna is to be used on VLF (where it was originally used), the velocity of propagation in the two wires (one is the antenna conductor, the other one is its image in the earth) has to be different, so that the arriving wave front (at a 0° elevation angle for ground wave VLF signals) inclines onto the wire and induces an EMF in the wire. Therefore, the ground under the antenna must have rather poor conductivity for best performance.

A radio wave travels in the air with a velocity factor of 100%. In the antenna wire it travels at a slower speed, depending on its height above ground and the quality of the ground. As we make the wire antenna longer, an increasingly large phase difference exists along the wire between the wave in the air and the wave in the wire. When the difference becomes 90°, the EMF induced by the radio wave will start subtracting from the traveling wave in the wire instead of adding. This mechanism limits the length for maximum gain in this antenna.

On the amateur low bands (MF and low HF), the situation is different, because signals do not come in at 0° elevation angle. The elevation angle is typically above 0° and below 30° for DX signals on 160 and 80 meters. Here too we have to look at the phase difference between the wave in the air and the induced EMF wave in the antenna wire. It's clear that in the horizontal plane through the wire, the projected wavelength is now the wavelength in the air multiplied with the cosine of the elevation angle. The wave in the wire is now the faster one, at least for high enough elevation wave angles. For a given height and ground quality there is an elevation angle at which the velocity factor (VF) and the elevation angle compensate one another and where both remain in phase all the time. Below this elevation angle the wave in the antenna is the slower one, and above this angle it is the slower one. This angle is also the elevation angle of the antenna, or the angle at which the gain is greatest. For signals coming in at higher elevation angles (and long enough antenna wires) the gain drops, goes through a minimum (180° phase difference), goes up again in a cyclical way.

This is the mechanism that forms the radiation pattern of the Beverage. See **Fig 7-74**. Notice the secondary lobes. The longer the antenna, the more nulls there are, as this phenomenon of adding/subtracting keeps repeating along the wire.

For the main elevation wave angle (the angle at which the main lobe peaks) there is a length beyond which the gain will drop. From Fig 7-74 you can deduce the theoretical maximum, as a function of practical parameters, such as the antenna height and ground quality, both of which together determine the velocity factor of the antenna. However, the velocity factor of the antenna is not the mechanism that limits the useful maximum length of a Beverage antenna. In our approach above, we assume that the wave in the air is homogeneous and linear in space, which in reality is not the case (because of space diversity). The true mechanism that limits Beverage length is discussed in Section 2.4.1.

The velocity factor of a Beverage will vary typically from about 90% on 160 meters to 95% on 40 meters. These figures are for a height of 3.0 to 3.5 meters. At a 1-meter height the velocity factor can be significantly lower, depending on ground quality. BOGs (Beverages on ground, covered in Section 2.12) may have a velocity factor around 60%, which means that very short, very low Beverages can exhibit radiation patterns similar to higher Beverages that are much longer. A drawback of "on the ground" Beverages is that their output is much lower. But that may not be our main worry. After all, people trip on them, deer get tangled in them and critters eat them! The good news is that, if you use Teflon insulated wire, rodents may stay away and you can actually bury the BOG Beverages a few millimeters below the ground.

2.3. Modeling Beverage Antennas

2.3.1. Modeling with *MININEC* Based Programs

I don't advise modeling Beverages with *MININEC* based programs (see also Chapter 4, Section 1.2) because *MININEC* assumes a perfect ground under the antenna. This has a number of consequences such as:

- *MININEC* works with a 100% velocity factor (because of the perfect ground under the antenna). The gain will increase with length, without reaching a maximum, which is not correct.
- *MININEC* will show an antenna impedance that is typically 10 to 20% lower than the actual impedance over real ground.
- *MININEC* patterns will show deep nulls in between the different lobes. This is not correct. The various lobes merge into one another due to real-ground conditions.
- Gains reported with *MININEC* are too high.

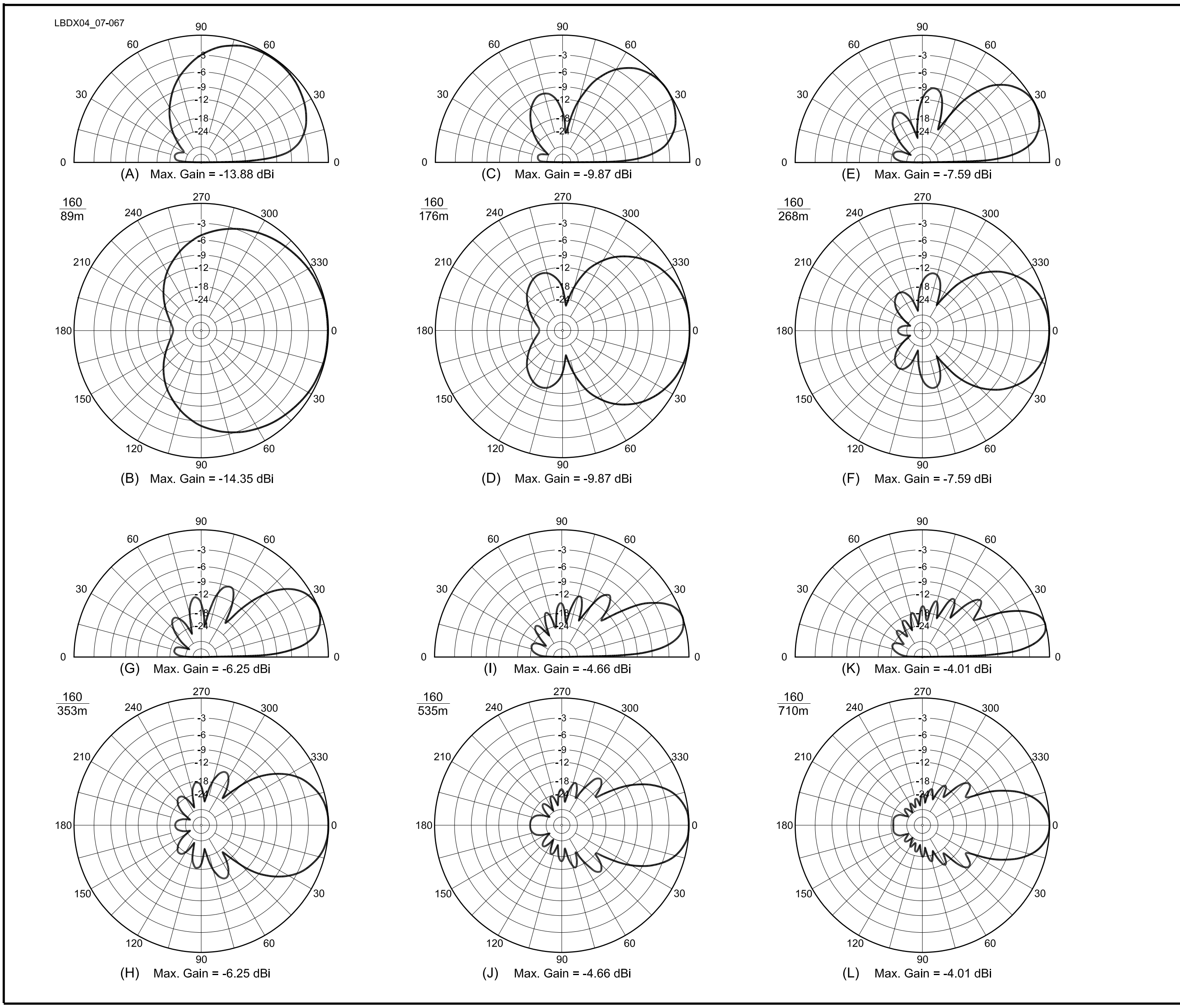
2.3.2. Modeling With *NEC*

We can do much more accurate modeling using modeling programs based on *NEC-2* (or *NEC-4*). See Chapter 4, Section 1.3. Using a *NEC-2* based program, the Beverage antenna's velocity factor is taken into consideration. Modeling various Beverage lengths will show the gain going through a maximum at about 5 to 8 λ, depending on height and ground quality.

NEC-2 is, however, well-known for predicting excessive gains for wires close to the ground. It underestimates loss along the antenna, which distorts the pattern. W8JI wrote: "I measure as much as 70% current loss in 500 feet, which is 3-dB loss." As gain is no issue with receiving antennas, this should not be a problem. But, we are much more concerned with the directivity of this antenna than with its output (gain), which is rather irrelevant for use as a receiving antenna on the low bands.

With *NEC-2* you cannot connect any part of the antenna to the ground. With *NEC-4* you can do this, but *NEC-4* is not released to the general public and a *NEC-4* license is still very expensive (see also Chapter 4, Section 1.4).

There is a simple way though to overcome this problem with *NEC-2*: We terminate the Beverage in our computer model



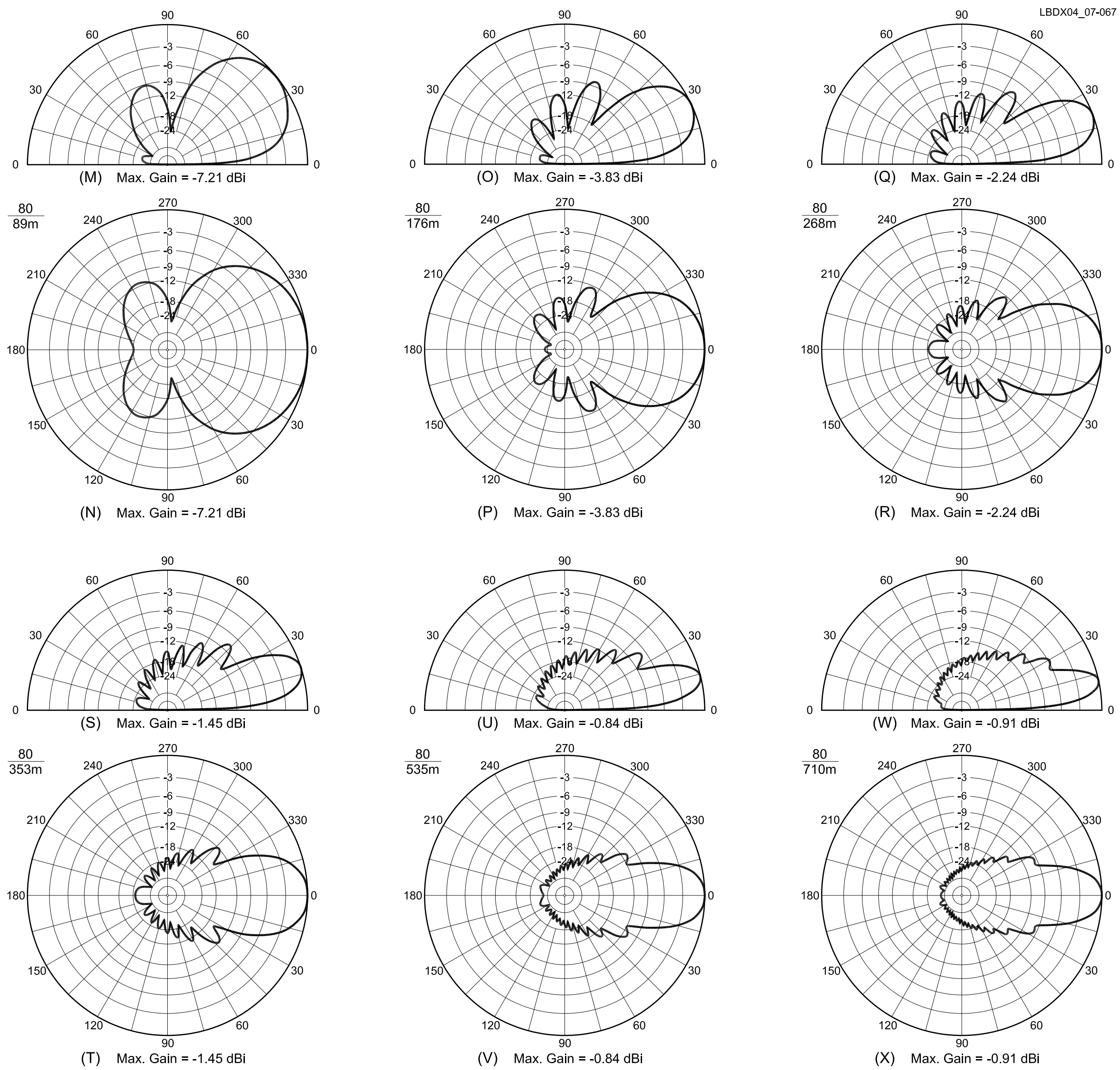
using quarter-wave wires as ground terminations. If you use such a scheme in reality, you will have a single-band Beverage, since the termination wires are only λ/4, a dead short, on one frequency and its odd multiples.

The correct configuration for modeling Beverages uses two λ/4 termination wires (in-line) at each end of the Beverage (see Fig 7-73). Considering the current distribution, these wires do not radiate in the far field, just like symmetrical radials or top-loading wires with verticals (see Chapter 9). We should not automatically extrapolate this to the real world. In the real world, in close proximity to a ground that does not exhibit constant characteristics all along these wires, near-field losses will be different in different places, resulting in an imperfect cancellation of the radiation from the two λ/4 wires in the far field. A similar problem is discussed in Chapter 9, where we deal with verticals using just a few elevated radials.

In the real world, quarter-wave terminations are rarely used. If they are used, it would mostly be as a single wire, in-line with the antenna wire. Directivity is slightly better with these models than in the real world, where we use a vertical or a sloping downlead at both ends. This vertical downlead (or the vertical component of the sloping lead) is responsible for some omnidirectional signal pick up. The impact of the pick up from the down leads is further discussed in Section 2.8.

Most Beverage patterns described in this chapter were modeled using the *EZNEC* program (which uses the *NEC-2* computing core) and using the real "High Accuracy" ground option, employing the Sommerfeld ground method. These *EZNEC* models were terminated using two in-line quarter-wave terminations at each end (Fig 7-73B).

When modeling arrays of Beverages (Section 2.3), I often use a simplified model. In this model all Beverages are in an inverted-V fashion, 2 meters high at the center and sloping down to ground level at both ends. Since this configuration includes a ground connection, modeling was done using a "Real-*MININEC*" type of ground in *EZNEC*. This is not really a disadvantage, so long as I do not compare gain figures obtained using the two different methods.



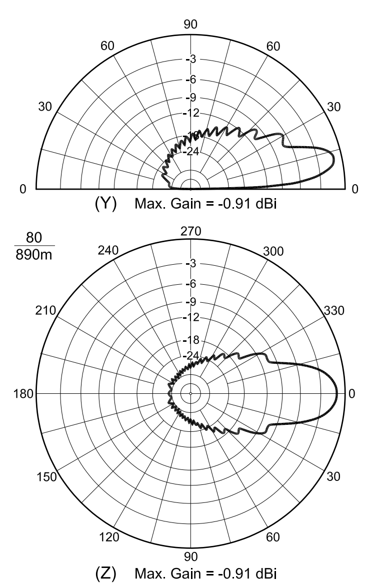


Fig 7-74 — Vertical and horizontal radiation pattern of 2-meter high Beverage antennas over good ground, for different "cone of silence" antenna lengths for 80 and 160 meters. In these patterns you can think of the Beverage feed point being located at the center of the pattern and running to the right along the X axis.

2.4. Directional Characteristics and Gain

Directivity is the name of the game with any receiving antenna. Forget about gain — preamplifiers can do wonders! I analyzed a series of Beverage antennas (at 2 meters height) for 160 and 80 meters. I modeled these antennas over good ground (see Table 5-2 in Chapter 5), terminating them in a 600-Ω resistance. (See Fig 7-73 and also Section 2.5.6.)

2.4.1. Influence of Length

The lengths of the modeled Beverage antennas varied from 89 to 890 meters. The choice of lengths was indicated by the fact that these lengths happen to be the ideal target

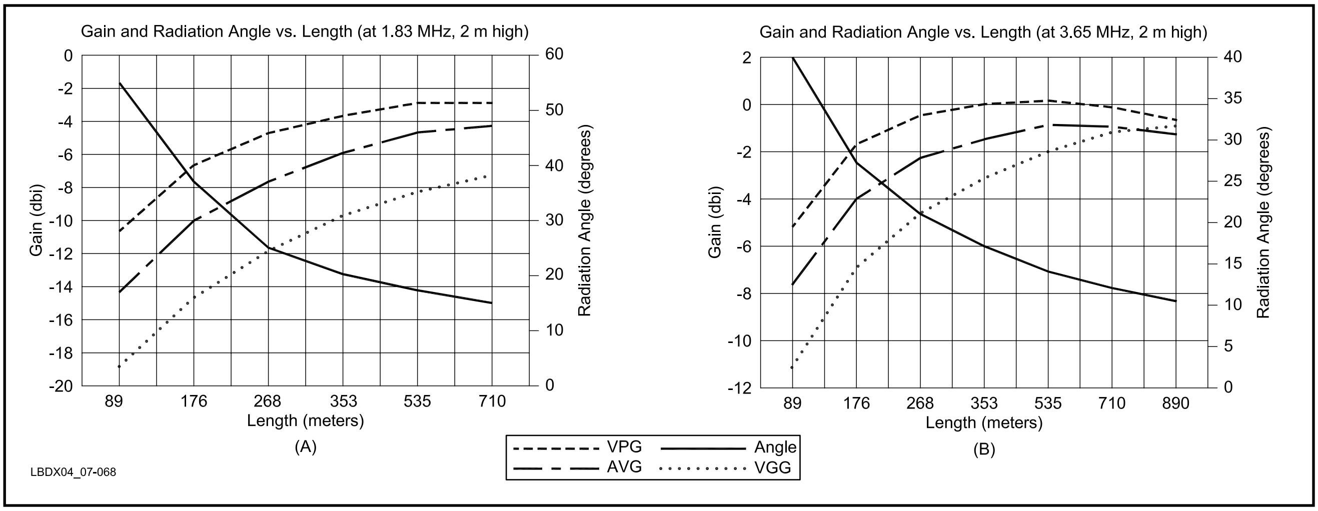


Fig 7-75 — Gain and elevation angle for a 2-meter high Beverage antenna for 160 meters, as a function of the antenna length. Three curves are shown: over Very Poor Ground (VPG), over Average Ground (AVG), and over Very Good Ground (VGG). The radiation angle is computed for Average Ground. This angle only changes marginally between Very Poor and Very Good ground.

Table 7-23

80-meter Beverages

Length Meters	*Gain dBi*	*–3-dB Angle Degrees*	*DMF dB*	*RDF dB*
89	–8.6	90	15.4	9.3
176	–4.1	59	20.6	13.1
268	–2.2	43	23.6	15.0
353	–1.6	35	24.6	15.8
535	–2.3	26	23.8	15.4
160-meter Beverages				
89	–15.9	122	11.7	6.5
176	–10.6	86	16.6	10.1
268	–7.8	66	21.3	12.2
353	–36.3	55	21.8	13.6
535	–4.8	40	24.5	15.3
710	–4.6	32	24.2	15.6

lengths for obtaining best F/B. The influence of ground quality is discussed in Section 2.4.4.

The patterns: Fig 7-74 shows the horizontal and the vertical radiation pattern for different Beverage antenna lengths. The horizontal radiation patterns are calculated for the maximum elevation angle.

The elevation angle: The radiation angle only changes marginally (a few degrees) between very poor ground and very good ground. The elevation angles shown in Fig 7-74 are for Average Ground. Note the large difference in elevation angle between long and short Beverages: 17° for a 3-λ long antenna and approximately 40° for a 1-λ long antenna!

Gain: **Fig 7-75** shows gain curves vs length, over very poor ground, over average ground and over very good ground. As indicated above, the gain figures may be a little optimistic, a known flaw with *NEC-2* for antennas close to the ground. From these curves it may seem that maximum usable length is determined by the way gain diminishes beyond a certain length. This is not important because gain is not an important parameter with receiving antennas.

Directivity: We learned in Sections 1.8 and 1.9 that the most important parameters for receiving antennas are DMF (Directivity Merit Figure) and RDF (Receiving Directivity Factor), as well as the –3 dB main-lobe angle. **Table 7-23** lists the DMF and RDF vs 20°-elevation-angle forward lobe and the 3-dB beamwidth angle shown in Fig 7-74 for 80 and 160 meters. Note that these are modeled values that assume there is no space diversity effect involved.

Is longer really better?

It appears that you can build very long Beverages (4 to 5 λ long) and get really superb directivity. Here again, models and reality may not always be the same. There is such a thing as *space diversity*, which means that wave characteristics change with place. As long as you stay within a radius of approximately two wavelengths, this usually does not cause any problems. This is the reason why very large arrays and very long Beverages, may actually behave differently from what the model tells us. "*Longer is better*" does not hold true for Beverages (as for any large receiving array).

In real life the limit is not imposed by the velocity factor of the antenna (see Section 2.2), but by the space diversity. In modeling with *NEC*-based programs, losses are definitely underestimated, as all Top Banders who have actually measured losses can confirm. But again, loss by itself is not really an issue.

Beyond a length of three wavelengths, little seems to be gained, as I have learned from real-life experience. If you look at the gain in DMF and RDF beyond three wavelengths, there is little to be gained there as well. And since we don't go by gain with receive antennas, we can safely conclude that three wavelengths is the maximum we want to use.

If you want to improve your receiving antenna beyond this point, you can go to staggered end-fire phased Beverages or broadside phased pairs (see Section 2.16). If you have enough space I would recommend a 268-meter long antenna as a best compromise for the two bands. This is 1.5 λ on 160 and 3.0 λ on 80 meters. However, a 176-meter long Beverage (1 λ on 160, 2 λ on 80 meters) is very effective as well.

These exact length figures are more symbolic than anything else. We will see in Section 2.4.2 that there are no such things

Table 7-24
Performance for 320-m Beverage at Various Heights

1.83 MHz				
Height	*1 m*	*2 m*	*4 m*	*6 m*
DMF (dB)	21.6	21.6	20.7	19.4
RDF (dB)	13.6	13.1	12.4	11.7
Opt Rterm	450 Ω	500 Ω	525 Ω	550 Ω
3.65 MHz				
DMF (dB)	23.3	24.2	23.6	22.2
RDF (dB)	15.2	15.5	15.2	14.7
Opt Rterm	450 Ω	500 Ω	525 Ω	550 Ω

as *magic* Beverage lengths. In a nutshell: A length between 160 to 270 meters seems to be optimal for Top Band.

If you choose to use 300-meter long Beverages to cover all directions, and you want to use them on 80 and 160, you have to take into account the HPBW (half-power or 3-dB beamwidth) of 40° on 80 meters. This means you need to use at least nine Beverages to cover all directions equally well. At my QTH I have Beverages ranging from 170 to 300 meters long, and I have 12 of these spread out every 30°.

2.4.2. The "Cone-of-Silence" Length

Authors looking for F/B optimization as a function of antenna length developed the so-called "cone of silence" length. We now know that F/B does not mean much unless you would want to null out a specific local noise source. What we need to evaluate is DMF or RDF.

The concept of the cone of silence resulted in "sacred" Beverage antenna lengths, lengths that were supposedly better than others. It appears that the front-to-back ratio (geometric F/B) goes through maximum values for lengths that are a multiple of electrical half-wavelengths. This is logical, since it is pure trigonometry. However, the geometric F/B is not very relevant, as explained in Section 1.7.

If you assess the overall directivity performance of a Beverage, you come to the conclusion that there are no "special" lengths, provided the Beverage is properly terminated. I evaluated the DMF and RDF for Beverage antennas with lengths varying from 140 to 300 meters, using two different termination models. The "perfect model" uses two T-shaped quarter-wave wire terminations (Fig 7-73C). In the "sloping" model, both antenna halves slope down to the ground level from the middle.

The results for 1.83 and 3.65 MHz are shown in **Table 7-24** and are valid for both sloping wire and T-termination models. When properly terminated for best F/B, DMF and RDF both increase monolithically with length, without appreciable bumps. RDF is a fairly linear curve, mostly determined by the forward lobe. The terminations varied between 425 and 525 Ω.

In the DMF curve there seems to be some kind of "wave" superimposed on the curve, probably generated by the effect of the geometric F/B, which is largely undone in the RDF curve because of the impact of the forward lobe. We see that the wave tops out at about 160 meters and 300 meters, which are the so-called "cone-of-silence" lengths.

2.4.3. Influence of Antenna Height

The general rule is as follows:

- Higher Beverages produce higher output
- Higher Beverages have larger sidelobes
- Higher Beverages have a higher elevation angle
- Higher Beverages have a wider 3-dB forward lobe

Fig 7-76 shows the elevation and azimuth patterns of a 320-meter long Beverage for 160 meters, at various heights (1, 2, 4 and 6 meters) over average ground. The horizontal pattern was calculated for a 20° elevation angle. The DMF and RDF figures for 160 meters are listed in Table 7-24. The variation in gain between 1 and 6 meters height is less than 3 dB.

What you see in the tables is what you'd expect. The secondary lobes become more outspoken at greater heights, which reduces both the RDF and DMF. The secondary lobes are present in the front half of the radiation pattern as well as in the back half.

On 160 meters, a height of 4 meters seems to be still

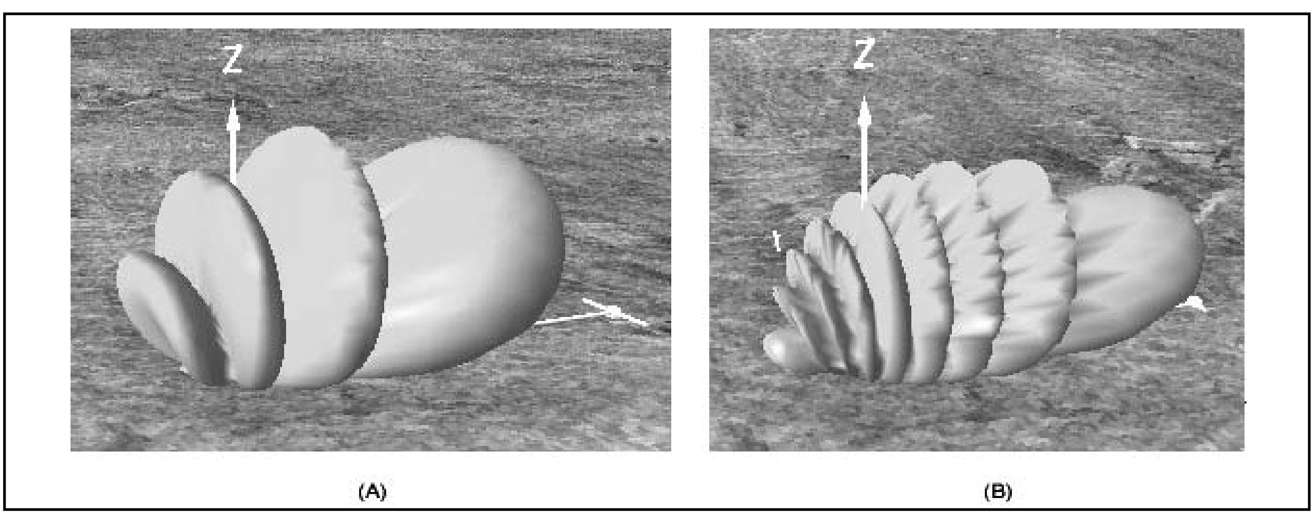


Fig 7-76 — Three-dimensional radiation patterns for a 320-meter long Beverage antenna. Top: the pattern for 160 meters, where the antenna is 2 λ long. Below: the pattern for the same antenna on 80 meters (4 λ). (Patterns generated with *4Nec2*.)

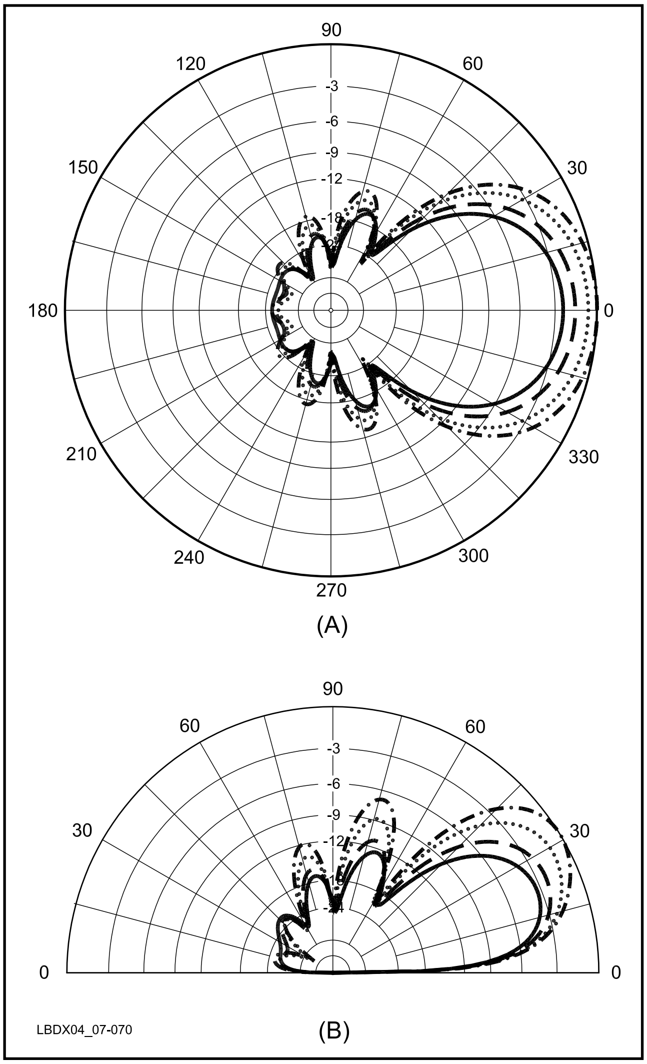


Fig 7-77 — Elevation and azimuth radiation patterns for 320-meter long Beverage antenna on 160 meters over average ground, for various heights. Solid line = 1 meter; dashed line = 2 meters; dotted line = 4 meters; dashed-dotted line = 6 meters. See text for comments. (Although the patterns for the different heights are shown together, again I want to emphasize the differences in patterns, since gain is not important for receiving on the low bands.)

very good, and even 6 meters does not sacrifice much (see **Fig 7-77**). What about using the same antenna on 80 meters? **Fig 7-78** shows the story. Amazingly enough even at 4 meters the secondary lobes are down almost as much as they are at 1 meter in height. Even 6 meters, which is generally considered as being way too high for 80 meters, is still a very good Beverage antenna! The directivity figures for 80 meters are also given in Table 7-24.

You have to be careful about extending these model findings to real life. The model used in the configuration shown in Fig 7-73C uses two $\lambda/4$ wires in-line as terminations, which means there is no influence from omnidirectional pick-up from a vertical or sloping down lead (see Section 2.8). The high-angle lobes that appear with higher Beverages are due to the increasing horizontally polarized radiation component. **Fig 7-79** shows the azimuth patterns (both vertically and horizontally

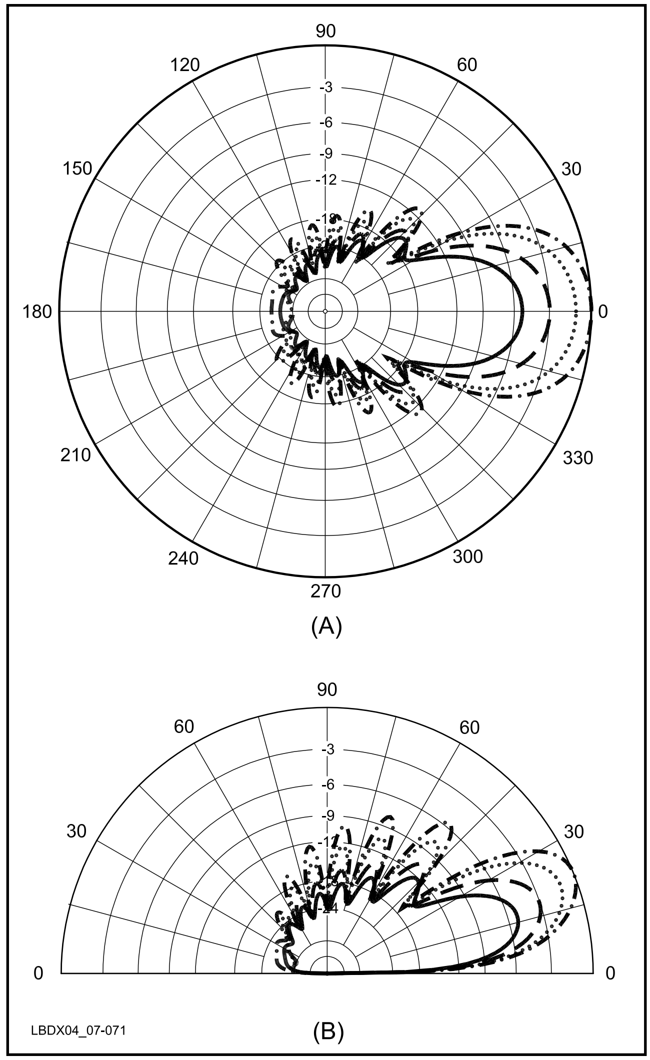


Fig 7-78 — Elevation and azimuth radiation patterns for 320-meter long Beverage antenna on 80 meters over average ground, for various heights: solid line = 1 meter; dashed line = 2 meters; dotted line = 4 meters; dashed dotted line = 6 meters.

polarized components, plus total pattern) for the 320-meter long Beverage at 3.65 MHz for heights of 6 and 1 meters. The horizontal component is significantly more important at 6 meters than at 1 meter. **Fig 7-80** shows the whole situation in 3D, with the horizontally polarized component on the right of the total pattern for 80 meters at the top, and 160 meters at the bottom.

When elevated even higher, the Beverage will start behaving like a terminated long wire, not like a Beverage antenna. This increased high-angle response of a high Beverage is often used by those who don't believe in important path skewing, to explain "apparent" path skewing (see Chapter 1). Although I don't deny that reception of high-angle sidelobes may cause some confusion at times, the existence of direction skewing has been confirmed repeatedly through the use of other directive antennas, such as phased arrays, which do not have such high-angle secondary lobes.

If you suspect you are receiving signals from such high-angle sidelobes, you can usually verify this by switching to a high-angle antenna, such as a low dipole. As explained in the

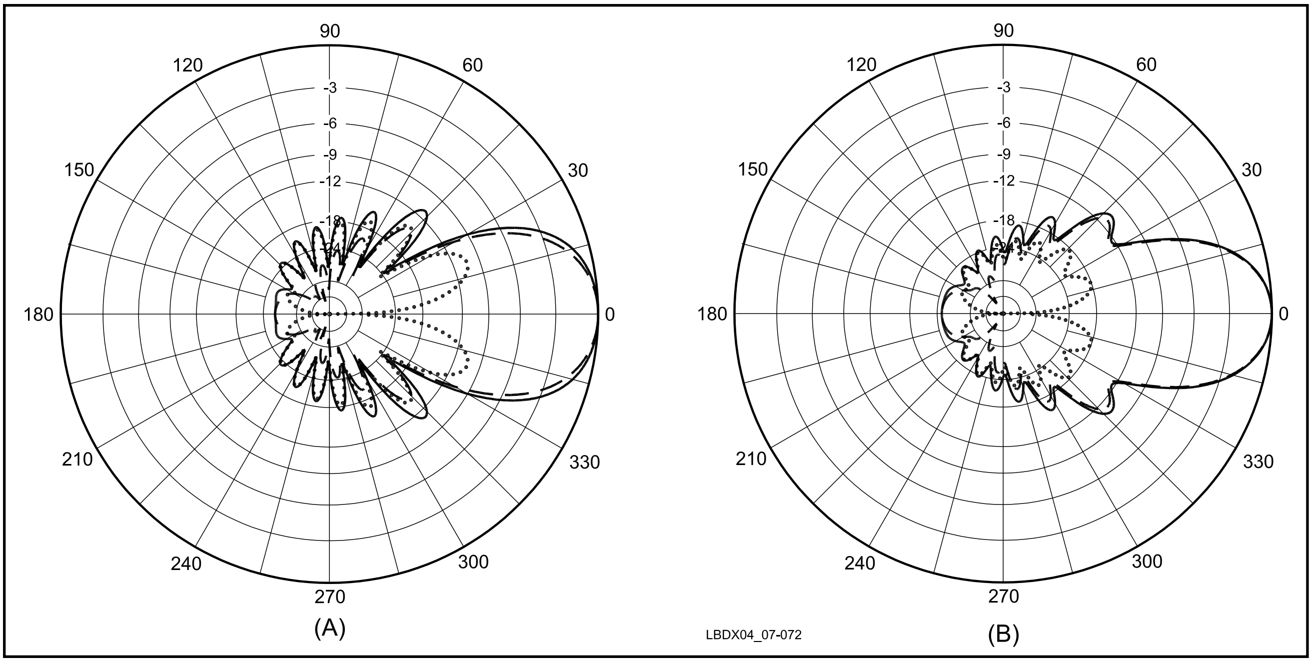


Fig 7-79 — Azimuth radiation patterns for 6-meter high (A) and 1-meter high (B), 320-meter long Beverage at 3.65 MHz. The patterns show the total fields (solid lines) as well as the horizontal and vertical components (dotted and dashed lines). At the 6-meter height, the azimuth component is in certain directions approximately 10 dB stronger than at 1 meter high. This broadens the forward lobe.

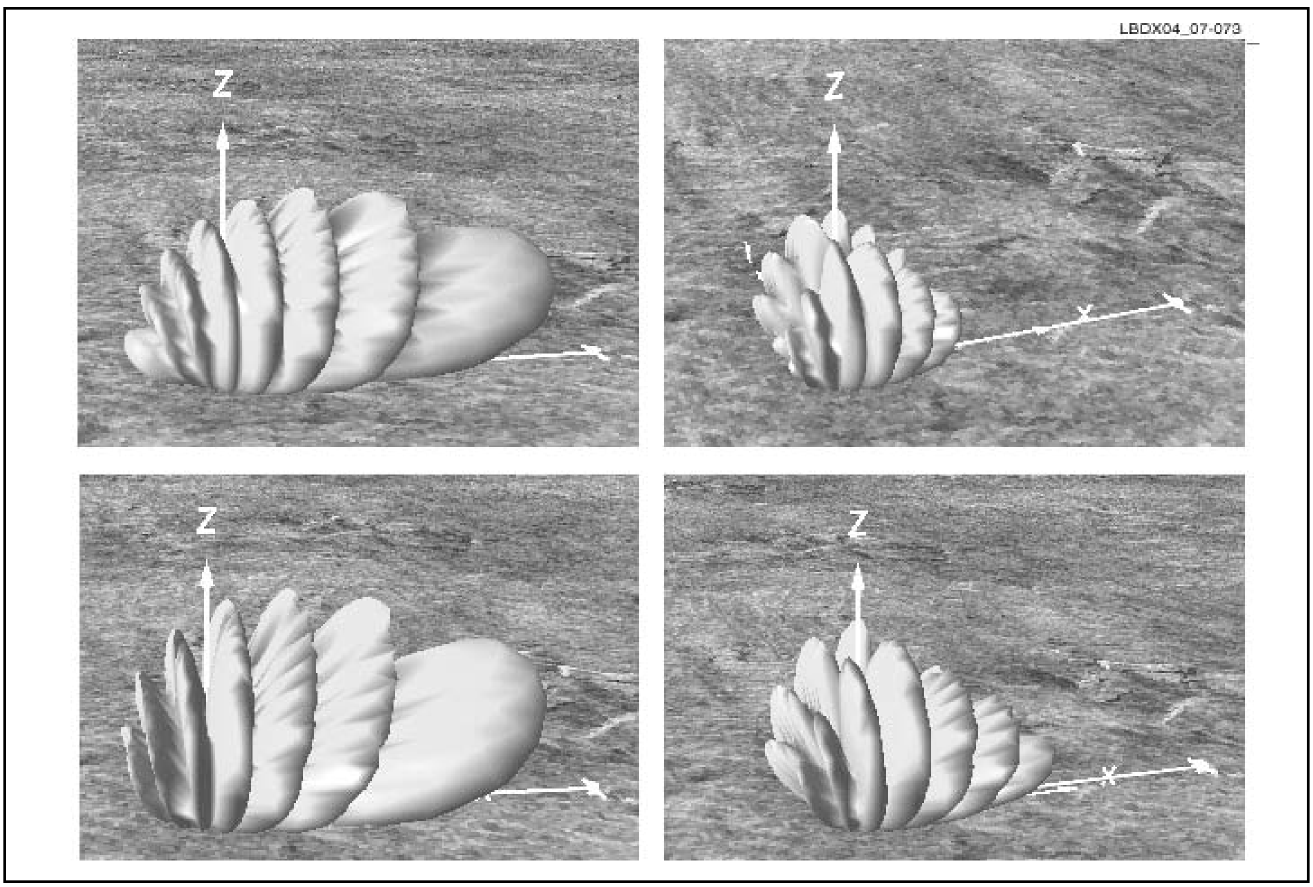


Fig 7-80 — Top left: 3D pattern for 320-meter long Beverage on 80 meters. Top right: the horizontally polarized component. Note this is a typical radiation pattern of the many lobes perpendicular to the wire, which we know from (high) "long" longwire antennas. Left bottom: 3D pattern for the same Beverage at 6 meters height. Note the slightly fatter and higher forward lobe, and the more outspoken secondary lobes. Right bottom: the horizontal component for the same 6-meter high Beverage. Note this component is much more important. (Patterns by *4Nec2*.)

chapter on propagation, this often can occur at sunrise or sunset (gray-line propagation) or during very disturbed conditions.

2.4.4.1. Height of Beverage Antennas: Conclusion

The height is not all that critical. Below 2 meters, Beverages can be a hazard for people and animals. If you must cross a driveway or small street, you can put your Beverage up to 6 meters high and still have a working Beverage. You can also slope the Beverage gently up from 2 meters to 6 meters to cross the obstacle without much harm at all, except for the fact that the surge impedance of the Beverage antenna will not be as flat as if it were perfectly horizontal and parallel to a horizontal ground. Tom, W8JI, writes on his Web page (**www.w8ji.com**): "I've found very little performance difference with height, unless the Beverage is more than 0.05 λ high."

If the Beverage can be constructed on terrain that is inaccessible to people, deer and other animals, then you can consider using Beverages at a height of 0.5 or 1 meter for added high-angle discrimination and reduced omnidirectional pick up. All of mine are about 2.2 meters high at the support post. Since I use fairly thin wire, mine sag quite a bit between the supports (to a height of about 2 meters), but I do seem to hear well nevertheless.

In Section 2.12 we will also cover BOGs (on-the-ground Beverages), and Beverages have even been reported working very well when buried a few millimeters below the earth surface (invisible "anti-trip" Beverages).

2.4.4. Influence of Ground Quality

The general mechanism is:

- The better the ground, the lower the output from the antenna (around 6 to 8 dB difference between very poor ground and very good ground). But even over very good ground Beverages have more than enough gain (signal output level)
- The peak elevation angle changes only slightly with ground quality. For example, a 300-meter long Beverage peaks at 27° over Very Poor Ground. The response at 10° elevation is down 3.4 dB from the peak. Over Very Good ground, the lobe peaks at 29°, and the response at a 10° elevation angle is down 2.8 dB from the peak response.
- The poorer the ground quality, the less pronounced the nulls will be between the different lobes. This is similar to what we notice with horizontally polarized antennas over real ground.
- The directivity factors (DMF and RDF) of a Beverage antenna remain almost constant for grounds ranging from Very Good to Very Poor.
- The Beverage does not work at all *over* saltwater. Its output is down 15 dB compared to the same antenna over poor ground and the main elevation angle is at 45°. This confirms the observations made by Ben Moeller, OZ8BV, that his Beverages near the sea never worked well at all. The Beverages at VKØIR, erected over a saltwater marsh never worked either.
- This does *not* mean they do not work *near* saltwater! You can construct a Beverage running alongside the saltwater (on the beach), if the Beverage is separated from the saltwater at least 3 to 5 times its height above ground. Another condition is that the ground under the Beverage is not soaked with saltwater; it must have relatively poor conductivity. It is only the area right under and very close to the Beverage wire that needs to have relatively poor conductivity (much poorer than the conductivity of saltwater).
- With good ground, the vertical ends do become much more important than over poor ground (see also Section 2.8).

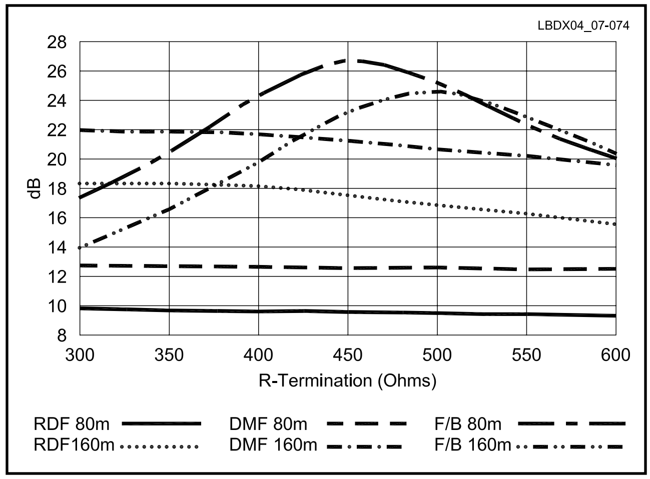


Fig 7-81 — F/B, DMF and RDF for a 160-meter long Beverage antenna (2 meters high, #12 AWG conductor, over AVG ground) terminated in a resistance between 300 and 600 Ω. Values shown for 80 and 160 meters.

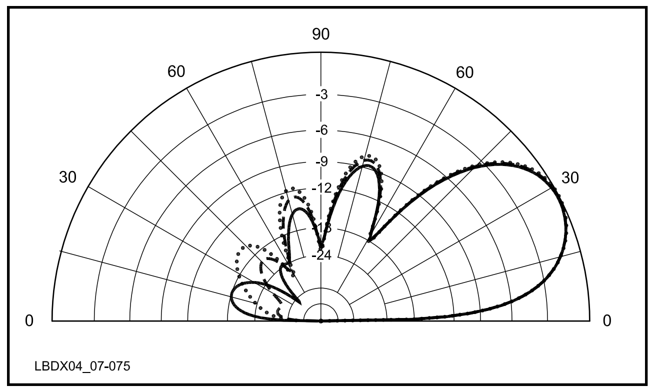


Fig 7-82 — Changing the termination resistance of 160-meter long Beverage from 300 Ω (solid line) to 450 Ω (dashed line) to 600 Ω (dotted line) on 3.65 MHz merely changes the place of the nulls in the back of the antenna, and hardly impacts the total directivity (DMF and RDF) of the antenna.

2.5. Terminating the Beverage Antenna

I have calculated the directivity patterns for a 160-meter long Beverage (over average ground, 2 meters high, wire: #12 AWG) on both 160 and 80 meters. While the F/B changes for a given elevation angle, the RDF and DMF figures remain relatively constant, as shown in **Fig 7-81**, which shows that the 3.7-MHz geometric F/B peaks for a termination value between 400 and 500 Ω. For thinner wire (#20 AWG) these values will be somewhat higher. Unless you need to null out a local noise or QRM source right off the back of the antenna, the exact value of the termination resistance is far from critical. Varying the termination resistance merely changes the position of the notches in the back of the Beverage. See **Fig 7-82**.

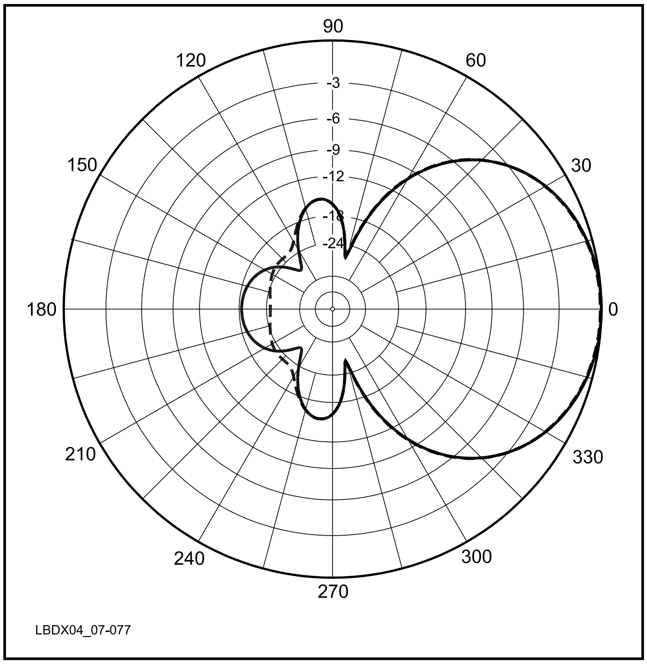


Fig 7-83 — A complex termination impedance can be used to move the notch in the back of the antenna around (dashed line). It hardly affects the overall directivity (DMF and RDF) however.

Table 7-25
DMF and RDF for Resistive and Inductive Terminations

Termination	*DMF*	*RDF*	*F/B*
525 + *j* 0 Ω	17.5 dB	10.7 dB	20 dB
475 + *j* 125 Ω	17.2 dB	10.6 dB	44 dB

2.5.1. Inductive and Capacitive Load Terminations

Take the example of a 200-meter long Beverage (2 meters high, over AVG ground, #20 AWG wire — 0.8 mm diameter) terminated in a resistor giving the best F/B: 525 Ω yields a 20-dB F/B. If you terminate the Beverage in a complex impedance of $475 + j\,125\ \Omega$, you obtain a higher F/B, as **Fig 7-83** and **Table 7-25** show. Let's analyze the DMF and RDF for both cases:

Looking only at the F/B clearly leads us to the wrong — or rather to an incomplete — conclusion. All you do by changing from a resistive value to a complex load, is to move the nulls around in the back of the antenna, but that does not significantly influence the global directivity (RDF or DMF) of the antenna.

Unless you want to use your Beverage for nulling out one specific noise source in a very specific direction, the complex termination impedance has little or no added value. When you use such a complex termination the Beverage becomes a single-band antenna, and you will have to switch loads for different bands. The only sensible application for inductive terminations is with very short Beverages less than 0.5 λ long (see Section 2.13).

Unless you want to null out a specific noise source that is present all the time, it makes no sense to "tune" a Beverage antenna. By adjusting the termination impedance (either as a pure resistor or as a complex impedance), you can put a null right in the direction of the noise source. The adjustment of

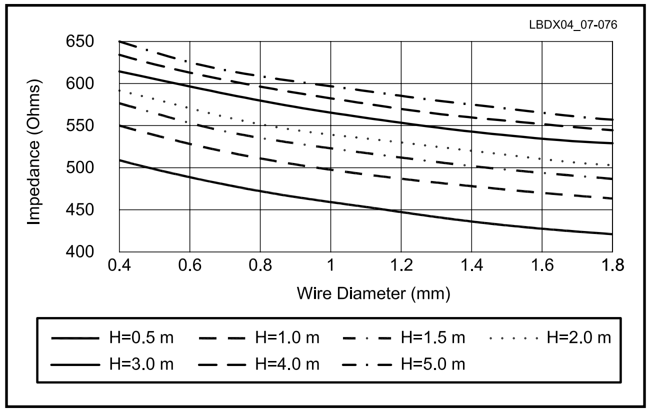


Fig 7-84 — Characteristic impedance of a single-wire Beverage antenna for different conductor diameters and different antenna heights. The values are calculated for the single-wire feed line equivalent. In practice the values can be 10 to 30% higher, depending on the ground quality. See text for details.

the terminating impedance can be done using a small signal generator with a small whip antenna. This should be placed outside the near field of the antenna, a minimum of 2 λ from the far end of the Beverage.

I must say, however, that using a second noise-sampling antenna and a noise canceller (as described in Section 1.37) is usually a simpler and better solution to the problem of eliminating a noise from a fixed source in the neighborhood. You will only eliminate the local noise — and not all the signals coming from the direction of the noise — because the small pick-up antenna will not hear them well enough.

2.5.2. Beverage Impedance (Surge Impedance)

Over perfect ground the single wire Beverage impedance can be calculated using a simplified model using the formula of the single-wire transmission line over ground:

$$Z = 138 \log \left(\frac{4h}{d} \right) \qquad \text{(Eq 7-1)}$$

where

h = height of wire

d = wire diameter (in the same units).

The theoretical impedance values listed in **Fig 7-84** are calculated over perfect ground. They are useful for estimating the terminating resistor for a single-wire Beverage and for designing matching transformers and networks. Note that the impedance does not change drastically with height or wire size.

Use these values to convert AWG to wire diameter in mm: #10 AWG = 2.6 mm; #12 AWG = 2.0 mm; #14 AWG = 1.6 mm; #18 AWG = 1.0 mm; #20 AWG = 0.8 mm; and #22 AWG = 0.65 mm.

Over real ground the surge impedance appears to be higher than over perfect ground. This is because the electrical ground is not the surface of the ground, but a little deeper. The following correction figures can be used as compared to the impedance over perfect ground in Fig 7-84.

- Good ground: +12%
- Average ground: +20%
- Very poor ground: +30%

These termination values are for highest F/B at low elevation angle (450 Ω in Fig 7-81 on 80 meters). Very low Beverages (BOGs, see Section 2.12) do *not* have a *very low* impedance as is sometimes claimed. A BOG on the grass shows about a 300-Ω surge impedance because the actual ground is deeper than the surface of the soil. That's also why we can "bury" a BOG a few millimeters in the ground and make it a trip-free BOG that still works! 2.5.3. Determining the Best Termination Resistance

With the correct terminating resistance, the antenna current along the Beverage shows little or no sinusoidal pattern but rather it shows an exponential decrease toward the termination (due to the attenuation). There are different ways to determine the best termination resistance value. The principle with all of them is to vary the resistance value for minimum standing waves on the antenna.

- Couple your antenna analyzer to the Beverage transformer and tune across the spectrum (for example, from 1 to 10 MHz). When you do this with the far end open-ended (or short circuited) you will see large swings in impedance or SWR. The proper terminating resistance is the value for which the variation in impedance or SWR is least when tuning across the entire spectrum over which you want to use the Beverage. In this exercise the termination should *not* be adjusted to achieve the best SWR, but rather the *flattest SWR curve*.
- Excite the antenna with a small signal, and measure the current along the antenna with a clamp-on RF current meter or RF voltage meter. Adjust the termination resistance until the voltage or current has a uniformly smooth taper toward the far end (typically 25% to 50% depending on ground quality and antenna length).
- Using your antenna analyzer, measure the feed-point impedance (using the method described above) across a range of frequencies (perhaps 1.5 to 3 MHz) and note the lowest (Z_{min}) and highest (Z_{max}) impedance values. The Beverage surge impedance is then given by

$$Z_{Bev} = \sqrt{Z_{max} \times Z_{min}}$$

- Measurement methods making use of a field-strength meter are useless unless extreme care is taken because pretty much all measurements are done in the induction (near) field.

Whichever method you pick, the impedance (or SWR) should remain fairly constant as frequency is varied. That means the antenna is properly terminated. A very small impedance change with frequency results in the best F/B at low elevation angles. Do not forget, however, that RDF and DMF are generally much more important than F/B and are hardly influenced by the value of termination resistance (see Fig 7-81).

2.5.4. Measuring the Surge Impedance of the Beverage Antenna

In various literature describing one of the above test methods, we inevitably find expressions such as "fairly flat," "fairly constant," "small variations," and so forth, but very seldom do we see any numbers given.

What follows is the procedure that Roger, ON6WU and I use to check my Beverages and to find their source impedance and the best value of the terminating resistance for each one.

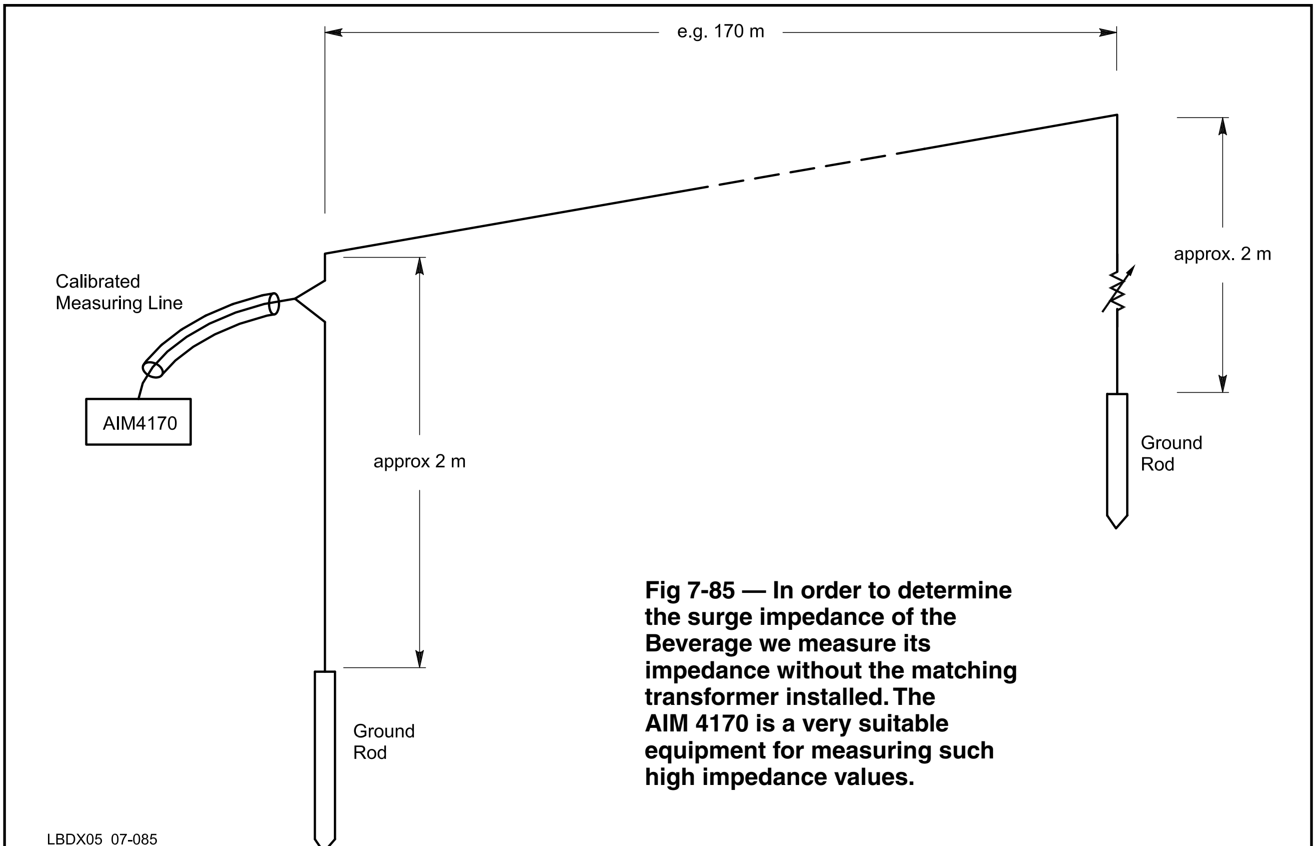


Fig 7-85 — In order to determine the surge impedance of the Beverage we measure its impedance without the matching transformer installed. The AIM 4170 is a very suitable equipment for measuring such high impedance values.

See **Figs 7-85** through **7-87**.

The test requires an impedance analyzer and a calibrated variable resistor (0 to 1000 Ω). Of the various antenna analyzers I have used, I find the AIM 4170 (available from Array Solutions) to be the most suitable tool for analyzing and setting up Beverage antennas.

Let us assume you want find out the Z_{surge} of your new Beverage, in order to know the best termination resistor value for that antenna. How shall you proceed?

1) Make sure the AIM 4170 analyzer is calibrated *including* a length of coaxial cable, long enough for you to set up the analyzer and laptop PC comfortably near the Beverage termination. In my case, where all receiving antennas are fed using 75-Ω coax, I use a 10 meter length of high quality 75-Ω cable. This means you need to connect your calibrating terminations (open, short and 75 Ω calibrating resistor) at the end of that coax cable and not at the analyzer.
2) Next, connect the end of the cable between the end of the Beverage and ground (see Fig 7-85). The connection can be made at antenna height, at ground level or anywhere in between. It does not make any difference.
3) Set up the AIM 4170 software as follows:
   - Set up Z_0 reference at a value of between 500 and 600 Ω (the value of Z_{surge} you expect to find). Let us start with 500 Ω.
   - Limits: 1.7 to 7.3 MHz, 0.01 MHz step.
   - Plot parameters: SWR (and any other parameter you might want to know).
   - Scales: SWR: 2 (maybe also Z_{max}: 1000 Ω, return loss: 30 dB).
4) Connect the variable termination resistor between the end of the Beverage and the ground rod. You can put the resistor at ground level or at antenna level (or anywhere in between). Adjust the value to 500 Ω total load resistance (potentiometer value plus ground resistance). If you use a single ground rod in average to good ground, you can estimate the ground resistance to be between 100 and 200 Ω.
5) Make a plot of the impedance and SWR.
6) Adjust the termination resistance and change the Z_0 reference of the AIM 4170 software to that same value (for example, 100 Ω ground resistance, 525 Ω

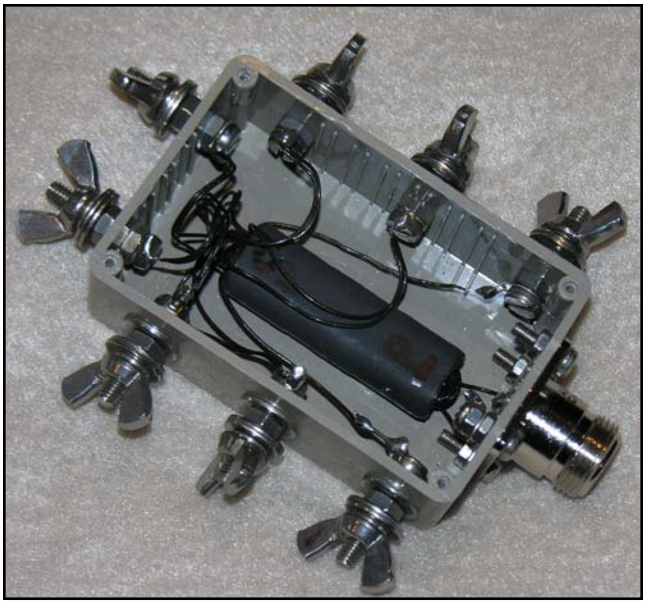


Fig 7-86 — A triple core multi-tap transformer yielding secondary impedances going from 200-612 Ω for a 50-Ω primary, or 300-920 Ω for a 75-Ω primary. This unit was specially built for determining the surge impedance of Beverage antennas.

Fig 7-87 — A high-quality, linear 1-kΩ potentiometer is a valuable instrument for testing Beverages and Beverage transformers.

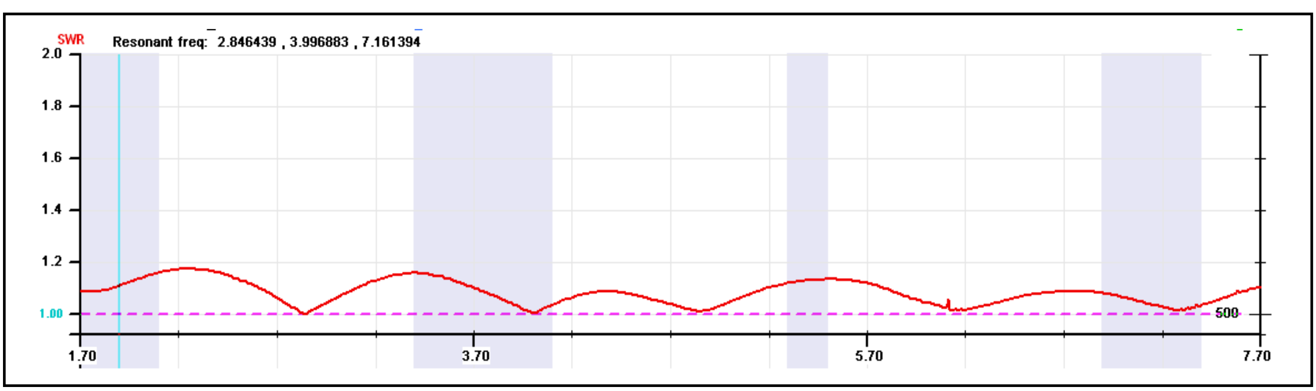


Fig 7-88 — The fact that the SWR line is not perfectly flat is due to the fact that the surge impedance of the Beverage is not a pure resistance. Inside the three bands (160, 80 and 40 meters) the SWR is less than 1.1:1 (return loss >25 dB).

potentiometer value and 525 Ω as Z_0 reference value for the AIM 4170 impedance analyzer). Make another plot.

7) By changing the value of R_{term} and Z_0 reference in steps, you will eventually find a value where the SWR line is "fairly" flat and where it comes down to a 1:1 value at a number of frequencies. **Fig 7-88** shows such a situation where the SWR, at any frequency between 1.7 and 7.3 MHz is lower than 1.2:1 (>21 dB return loss). In my opinion this is a very acceptable case of a "fairly" flat SWR curve for a Beverage antenna. The curve shown in Fig 7-88 is for a 160-meter long Beverage where $Z_{term} \times (R + R_{ground}) = Z_0$ reference = 575 Ω.

8) It is important we understand that we do this test *without* the matching transformer in line. Matching the surge impedance of the Beverage to the feed line impedance is a separate step.

Conclusions

At my QTH for my 0° Beverage, the Z_{surge} came out to be approximately 625 Ω, and that includes an estimated ground rod resistance of 50 Ω (very rich ground and a long, large diameter copper ground rod). The Beverage is made of 1 mm diameter wire at 2 meters height. From Fig 7-84, which was obtained via modeling, we anticipated a Z_{surge} of approximately 540 Ω + 12% = 605 Ω. We can conclude that the measured value and the value predicted through modeling match very well.

Again, we should not forget that the performance of the Beverage is not very dependent on a perfect match between the terminating resistor and the antenna's surge impedance. Changing the value of the termination resistance by as much as 20% merely moves around the notches in the back, but it hardly changes the DMF and RDF at all. Also, if you are not sure what the termination value should be, it is better to have it too low in value than too high as witnessed from the curves shown in Fig 7-81.

I am really not sure if it is worth all the effort to really measure the Z_{surge}. If we had simply gone by the values predicted in Fig 7-84, we would have terminated the antenna with a resistance of 555 Ω (605 – 50 Ω ground rod resistance) which would have been within a few percent of the value we measured, and which, for all practical purposes, would have been just as good. But ham radio is a technical hobby, and we want to "know." And we know that "measuring is knowing."

2.5.5. Terminating Resistors

The termination resistor must be a "low-inductance" resistor, which means that it cannot be a wire-wound resistor. In principle, any wattage will do but if you have a Beverage close to a transmitting antenna (which is bad), you may want to use a few 1 or 2-W resistors in parallel. I use a single 2-W resistor and have never seen one discolored or burned up. If they do burn up, you either have your Beverages much too close to your transmit antenna or you run excessive power — or lightning has been causing such problems.

Any type of resistor can be killed by a long-term overload. In some areas where lightning is frequent, inappropriate types of terminating resistors easily get blown out by lightning. Metal-film or carbon-film resistors are not good in that respect since they cannot withstand short overloads. Old type carbon-composition resistors are excellent, but they are difficult to find nowadays. Standard Ohmite OX/OY series ceramic-composition resistors will do the job as well. AB stopped manufacturing carbon composition resistor in 1997, but, although not cheap, these resistors are still available. One source is **www.hificollective.co.uk/components/allen_bradley_resistors.html**. John, WØUN recommends using Globar resistors (**www.globar.com**). K9DX uses a large fuse and clip for quick replacement when his terminating resistors are destroyed by lightning. See **Fig 7-89**.

If you are not sure the resistor has low enough inductance, just check it on your impedance analyzer. We should, however, not be overly critical about some reactive component in the resistor impedance, as we have learned that some reactance in the load resistor can actually improve the F/B (see Section 2.5). It will however, hardly change the RDF or DMF figure of your Beverage antenna. Just make sure you are not using a wire-wound resistor!

2.5.6. Protecting the Termination Resistor Against Lightning

If you live in an area with a high thunderstorm occurrence, the use of carbon-composition or metallic-composition resistors is a must. You can add further protection with small gas-discharge tubes connected across the resistor. These tubes are made by Bourns. Use the lowest voltage type, which is 90 V. This is the Bourns # 2035-09-B, which is available from Mouser as their # 652-2035-09-B. Bill, KØHA reported using Taiwan Semiconductor's SRYH-90L gas tube surge voltage protectors (the CATV model, with "high current capability") for the same purpose.

You could even put a pair of homemade small air-gap

Fig 7-89 — K9DX uses a fuse clip and parallel carbon resistors soldered across a blown fuse. This makes replacement, in case of lightning strike destruction, very easy.

electrodes across the resistor. Riki, 4X4NJ, described such a homemade spark gap: "The spark gap consists of heavy solid wires, about #10 AWG, soldered to teardrop terminals that are placed under the screws going to the antenna and ground. The ends of the wires are cut with side cutters leaving a nice knife edge, and I position the two edges very close to each other. A piece of paper makes a nice feeler gauge for this purpose. It is very effective, most simple, and negligible cost."

2.6. Termination Ground Requirements

What are the requirements our grounds should meet? Are they the same for the termination end as for the receiving end?

2.6.1. Ground Requirements at the Far End (Termination End)

If you have real soil (not rock), the ground system can consist of one or more ground rods. The RF resistance of the ground system at the far end (termination end) of the Beverage does not have to be very low, since even high ground resistance is effectively in series with the terminating resistance. A 1.5 meter (⅝-inch OD) copper or copper clad steel rod will have an RF ground resistance ranging from ~50 to ~300 Ω depending on ground quality. We have seen that the termination resistor value is not very critical (see Section 2.5). Let's assume you need a 425 Ω total termination resistance and that your ground rod is 100 Ω. You would require 325 Ω as a termination resistor. In most situations where you use a single ground rod, the actual resistance value of the termination will likely be between 250 and 500 Ω depending on the ground quality (conductivity). At my QTH, a single 1-meter long ground rod requires a resistor of approximately 470 Ω for Beverages that are approximately 2 meters high. This means ground conductivity is pretty good!

Do you want to know the RF ground resistance for the single ground rod at your particular QTH? Drive one rod in the ground, plus three more at about 2 meters distance around the first rod. Attach eight 25-meter long equally spaced wire radials to the ground system. Now you have a pretty decent ground of 20 Ω or better. Apply one of the procedures outlined in Section 2.5.2 to determine the optimum termination resistor value. Note this value (say, 415 Ω). Remove (not just disconnect) the radials and the extra three ground rods, and repeat the same procedure. Note the new value (for example, 330 Ω). The single ground rod RF ground resistance is thus: 415 + 20 – 330 = 105 Ω.

Where ground conductivity is really bad (Ref 1260) or where you cannot drive in a sufficiently long ground rod, the resistance of a single ground rod may actually be higher than the required terminating resistance. In that case you can install multiple ground rods, combined with a number of short radials, to bring the resistance down to an acceptable value. Do not use one or two long radials, but rather a large number of short radials forming a capacitance to ground (a ground screen). This was confirmed on the Internet by Greg, ZS5K, who wrote: "With my 800 ft Beverage, I have found it desirable to have more than one rod, and use more inserted resistance, so that the total termination resistance is less dependent on the weather. I have found that with only one rod, things change quite noticeably with the seasons."

At the receiving end of the Beverage you should pay a little more attention to the ground system.

2.6.2. Ground System at the Beverage Receiving End

Requirements are a little different at the receiving end of a Beverage antenna. The Beverage is in essence a asymmetric (unbalanced) system, one terminal being the antenna wire, and the other being the ground. This second terminal is what we call the *antenna ground*. This ground can be a good ground (low loss, low resistance) or a bad ground.

The ground rod resistance is in series with the high-Z (secondary) winding of the transformer. Assume we use just a short ground rod in poor ground. The RF ground resistance can be as high as 300 Ω. This ground resistance is in series with the secondary (high-Z; eg, 450 Ω) winding of the matching transformer (see **Fig 7-90**). In this case the voltage loss would be: 20 log (450/450 + 300) = –4.4 dB.

I've always said that gain is not the real issue with a receiving antenna, but that only holds true if you try to obtain gain at the sacrifice of the only important receiving antenna parameter, which is directivity. Using a better ground will not upset the antenna directivity.

We will see in Section 2.7.2.8 that it *is* important to have a low loss ground with Beverages, not for loss reasons. It's important to prevent unwanted common-mode signals from entering the Beverage feed system and deteriorating its performance through poorer S/N ratio.

How good a ground?

In Section 2.6.1. I explained how to assess the equivalent ground resistance of a single ground rod. I would suggest that at both ends of your Beverage antenna you try for a ground resistance that is less than 100 Ω. With very good ground, a single ground rod might do the trick. But you might need multiple ground rods, spaced at least the length of the ground rods, to lower the ground resistance. When you install rods too close together their effective *fields of influence* overlap and the end result is not much better than the first rod by itself.

Where you cannot use ground rods because of rocky ground, you can use a large ground mat made of large strips of chicken wire, made into the shape of a cross measuring approximately 6 × 6 meters. Or you could use a large number of

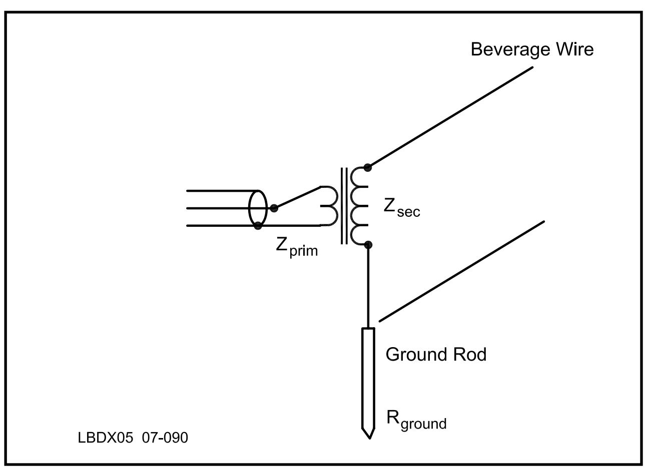


Fig 7-90 — In this example we see how the signal delivered by the Beverage antenna is split across the secondary of the transformer and the ground resistance which is effectively in series with the secondary.

short interconnected radials laid on the ground. Do not use a small number of long radials, as they could upset the directivity of the Beverage.

2.6.3. Which Kind of Ground Rods?

Many years ago, I used hot dip galvanized steel ground rods, having a cross section like a cross (to maximize the surface having contact with the ground). At my particular QTH this kind of ground rod completely rusted up within 10 years. Brian, K8BHZ, of Stone-HEX fame, uses ⅝ inch heavy galvanized steel rods to his satisfaction (see Fig 7-51).

Later I started using ½ inch copper-clad steel rods, but the ones I could buy here have a terribly thin copper flash with lots of pinholes, so that after a few years the steel rods also rusted away.

I have now switched to pure copper tubing. For the far end terminations I use a 1.25-meter long copper tube with a diameter of 20 mm. It is maybe a little more expensive, but I now know I have reliable and lasting ground rods. At my QTH (good conductivity ground), such a single ground rod appears to have an RF ground resistance of less than 100 Ω.

If you cannot use ground rods because of rocky terrain, make a ground screen consisting of 10 to 20 short radials (approximately 5 meters long) going in all directions. Do not use a single long radial, as you do not want this counterpoise to be resonant.

2.7. Matching the Feed Line to the Antenna

In Section 2.5.1 I discussed the surge impedance of a Beverage. The Beverage feed system transforms the antenna impedance to the transmission-line impedance (usually 75 or 50 Ω) and transports the received signals to the receiver. The entire feed-system can be broken up into these parts:

- The feed line.
- The feed-point transformer.

The technical issues involved are:

- Correct impedance matching.
- SWR on the feed line and its consequences.
- Common-mode suppression.

2.7.1. Which Feed Line to Use?

It is important that you pay as much attention to the feed line as to the Beverage antenna itself. Bad feed-line practice can completely annihilate the directive properties of the antenna. The Beverage principles explained here apply equally well to all other low-signal receiving antennas described in this chapter. Issues to consider are discussed in the following sections.

2.7.1.1. Coax Impedance

This is definitely the least important issue: It is totally irrelevant whether you use 75 or 50-Ω coaxial cable. If you have to buy cable, buy 75-Ω cable, as 75-Ω cable has, for a given diameter, less attenuation than 50-Ω cable. That is why 75-Ω cable is universally used in CATV systems. As it is so widely used, this cable is often available at bargain prices. If you already have plenty of 50-Ω coax, by all means use it!

2.7.1.2. Attenuation

As your Beverage antennas will most likely be operated on relatively low frequencies, you need not use a feed line with the lowest possible loss except for cases where the very longest feed line runs are being considered. For runs up to 100 meters, flooded (water-blocked) RG-6 coax will likely be a good choice.

The loss of common types of coaxial cable is a function of the frequency. **Table 7-26** shows the typical losses for common 75-Ω coaxial cables used for feeding Beverages on 1.8, 3.5 and 7 MHz. From this table we learn that up to approximately 100 meters of feed line length RG-6 will usually do the job. If you need longer lengths, ½ inch or even thicker cable may be advised. You can of course always use a good quality preamp to make up for the losses. Loss characteristics for many cable types may be found in Fig 6-2 in Chapter 6, the *TLW* software that comes with the *ARRL Antenna Book*, or on manufacturer's Web sites.

But how much attenuation can we really accept? Under the quietest circumstances (daytime) and using the narrowest receiver bandwidth we must be able to hear the antenna (band) noise over the internal receiver noise. If you are in a very quiet environment (way out in the country, no nearby power lines, or in the middle of the ocean on an uninhabited island) the surplus sensitivity that all modern receivers (transceivers) have on the low bands may not be enough. In that case you will need a preamplifier. It will practically never be necessary to put the preamplifier at the antenna though.

Table 7-26
Attenuation (dB/100 m) for Two Common Feed Lines

Coax Type	*----- RG-6 -----*		*--- 1/2 inch, 75 Ω ----*	
	SWR 1:1	*SWR 2:1*	*SWR 1:1*	*SWR 2:1*
1.8 MHz	1.1 dB	1.2 dB	0.3 dB	0.3 dB
3.5 MHz	1.6 dB	1.9 dB	0.4 dB	0.5 dB
7 MHz	2.3 dB	2.6 dB	0.6 dB	0.8 dB

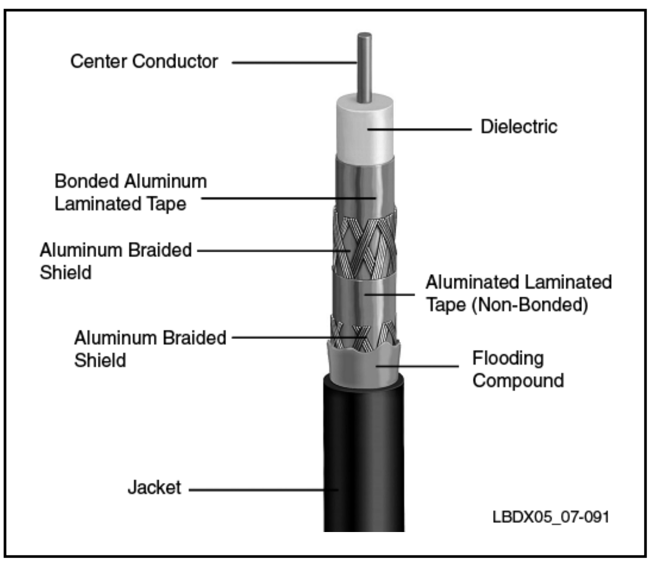


Fig 7-91 — Construction of the popular quad-shielded and flooded RG-6 coaxial cable. The flooding compound prevents water from entering the cable should the outer jacket be damaged.

2.7.1.3. Mechanical Properties

Running standard coax on the ground can be a problem for some. Many low banders report damage from animal bites to small or medium-sized RG-type or CATV drop-type feed lines. Small bites will usually not open or short the feed line, but will do enough damage to cause moisture migration and corrosion. There are three ways to prevent this from happening:

- CATV Hardline (½- or ⅝-inch stuff) that employs solid aluminum or corrugated solid copper tubing as the shield (outer conductor) covered by a high density PE jacket, is very sturdy and difficult to damage.
- Use quad-shielded and flooded RG-6 type coax (see **Fig 7-91**). It appears that the critters don't like the flooding compound.
- Bury the cables in a closed cable duct underground.

Also from a common-mode point of view (the outer shield acting like an antenna) it is always better to bury the coax in the ground.

2.7.1.4. Availability

For short runs any coax you can buy at the local flea market will do, provided the shield or the inner conductor is not corroded (green and black for copper shields, dull and white for aluminum shields) from moisture ingress. Often, 75-Ω CATV-type coax leftovers (often lengths up to 100 meters!) can be found at reasonable prices from the local cable TV company. The flexible coax used for drop lines is good for runs up to typically 50 meters. Quad-shielded and flooded RG-6 type coax is the common-sense choice up to 100 meters in length and is available new at attractive prices. Hardline (½ inch or thicker) is the ultimate choice for long runs, since it is the sturdiest (mechanically) and in addition offers the lowest attenuation.

George, K8GG, uses ½-inch 75-Ω CATV Hardline from his shack to the Beverage antenna park "head end," which is 1200 meters (yes, 4000 feet) from the shack. From the head end he uses RG-59 flooded cable for all the connection runs to the remote antenna selector.

I use ⅝-inch 75-Ω CATV coax to several remote head ends (two at 200 meters from the shack, another one at 350 meters), with ½ inch Hardline running from these head ends to the Beverage feed points. None of that stuff was bought new. It pays to have friends at the local CATV company.

2.7.1.5. Shielding Effectiveness

It is important to use well-shielded coax to achieve a quiet feed system under all circumstances and on all frequencies. Some of the very cheap (non Mil-standard) coax has a very poor shield coverage factor (50% or so). This cable should *not* be used.

Quad-shielded RG-6 cable has very good shielding effectiveness.

Hardline is the best choice, provided the solid shield is not broken. In many European countries, CATV companies use so-called figure-8 Hardline. Instead of lashing the coax to a messenger wire, figure-8 Hardline structurally incorporates a support cable. Overhead Hardline, especially the variety that does not employ the corrugated shield structure, sometimes develops cracks and eventually breaks in the solid shield due to work hardening as they swing in the wind. In a CATV trunking network these breaks are responsible for radiating signals, which many of us have experienced. Watch out when buying *used* CATV Hardline, and always check the cable for visual shield damage as well as for electrical shield continuity.

2.7.2. The Beverage Feed-Point Transformer

The easiest way to match the Beverage (typically 450 to 600 Ω) to common coaxial cable (50 or 75 Ω) is to use a wideband transformer. Such transformers are usually wound on magnetic-material cores and the most common shapes of the cores used are the *toroidal core* and more recently the *binocular core*. Such transformers are commercially available from various sources but can easily be homemade for a fraction of the price of commercial units.

2.7.2.1. Core Shape and Material

A number of types of core materials and different shapes of cores can be used for the job, but as we will see further, one type has particularly attractive properties. First of all, don't use just any core you find in your scrap box. Transformation ratio or SWR is not the only issue. I have seen transformers that showed a perfect SWR when terminated with the intended load impedance but exhibited a prohibitive loss of about 5 dB!

You can generally distinguish between two types of core material used at RF: *powdered iron* and *ferrite*. Ferrites have much higher permeability (up to 10,000) than powdered-iron materials (only up to 10), but ferrites are less stable at higher frequencies and saturate more easily. For wideband transformers used for receiving Beverages, ferrite cores are the most logical approach, as the application involves neither resonance nor high power levels. But there are literally hundreds of different types of ferrite materials, the majority of which are far from ideal for the application we have here.

Another distinction you have to make is between a transmission-line type transformer and a regular transformer. Core materials that are excellent for transmission-line transformers are not necessarily the best for a regular transformer.

What is a *transmission-line transformer*? The transformers where we wind two, three or more wires in parallel to make sections of a transformer are transmission-line transformers. With transmission-line transformers, the core material loss tangent isn't critical. All we want is a high impedance.

Regular transformers are transformers where the primary and secondary are usually wound as separate windings on the core (autotransformers are one exception). They can be heavily coupled (one winding on top of the other) or wound to have minimum capacitive coupling by winding them on opposite sides of a toroidal core. With this type of transformer, loss tangent does become a factor. Regular transformers rely on magnetic coupling, while transmission-line transformers do not.

The more isolated the windings, the more critical the loss tangent becomes. This means that the loss-tangent issue is especially important with transformers where primary and secondary are separated to obtain minimum capacitive coupling.

Various ferrite-core materials have been used for Beverage transformers: In the past type-77, 75 and even 43 mix have been described as being suitable for this job. Personally I have never used those materials but for more than 25 years I have used the Ceramic Magnetics MN8CX high-permeability cores and I have not had a single failure or complaint. More recently I have become an enthusiast for the binocular-core transformers recommended by W8JI. These use type-73 material and are much easier to wind than toroidal transformers.

2.7.2.2. Cores

Until quite recently most Beverage transformers were probably wound on *toroidal* cores, also called donut shaped cores. It was W8JI who broke the trend with the *multi-aperture* or *binocular* cores he recommends on his Web site.

The popular Amidon ferrite material toroidal cores using the 75 and the 77 mix (core sizes FT-50, FT-82 and FT-114), were tested, next to the toroidal core type MN8CX from Magnetic Materials (available from Misek Antenna Research, W1WCR), and the now popular binocular core manufactured by Fair-Rite (type # 2873000202).

The test results for the transformers using toroidal cores with 75 and 77 material turned out to be significantly inferior to what was obtained with the other two cores. I cannot recommend toroidal cores using these materials at all for making non transmission line type transformers for Beverages (and loop receiving antennas).

Of all tested cores, the binocular core gave the best results for designing and making the type of transformer we need for receiving antennas for the low bands. These binocular cores are available from different sources: Amidon code # BN 73-202, Ameritron code # 412-5250 and CWS Bytemark code #B-202-73.

The toroidal MN8CX cores — which I used for many years — do approach the results obtained with the Fair-Rite binocular core. While properly designed and wound transformers on the binocular core (mentioned above) can yield losses between 0.2 and 0.5 dB, a transformer using MN8CX cores will typically exhibit a few tenths of a dB more loss.

I have switched 100% to the abovementioned binocular cores because of two reasons: their performance is (slightly) better, and *they are so much easier to wind*. These are the reasons why the designs of the receiving antenna transformers in this chapter are all based on the use of the binocular core.

2.7.2.3. How Many Turns?

The basic formula that determines the *turns ratio* in a transformer is

$$\frac{Z_{prim}}{Z_{sec}} = \frac{N_{prim}^{\;2}}{N_{sec}^{\;2}}$$

There is also a minimum number of turns required. The low-Z winding has an impedance consisting of a resistive and a reactive part (R + *j*X). The parallel equivalent impedance is in parallel with the system source impedance (usually 75 or 50 Ω). Both parts play a role of their own in attenuating the signal and degrading the SWR (return loss). The magnitude of both parts (the reactive and the resistive part) differ a lot from one magnetic material to another (high-Q materials have a low series resistive part, low-Q materials a high resistive part). What is true for one type of magnetic material does not necessarily hold for another material.

It is a general "belief" that if we are using a *very low loss* core, the loss will be less than 0.1 dB if the magnetic impedance (the reactive part) of the winding is at least *five times* the impedance of the source or load connected to that winding.

Robye, W1MK, pointed out to me that this "five times" rule however does *not* apply to low-Q materials such as the 73 mix used for the Fair-Rite binocular core. He developed the mathematics to calculate the loss for a low band transformer (applicable for 160 and 80, and with some reservations to 40 meters), given the measured impedance of single turn on the core. An Excel file program for doing these calculations is available on the CD that comes with this book (*W1MK-TRF-insertion-loss.xls*). If we apply the popular "five times" rule to this material, we will end up with a loss of 0.9 dB loss! In that case the "five times" becomes "17 times" for an insertion loss of ≤0.2 dB, or "35 times" for an insertion loss of ≤0.1 dB. If the ratio is "only" 10, the insertion loss becomes approximately 0.35 dB.

Table 7-27 lists the calculated and the measured loss for a transformer using two, three or four turns as low-Z winding, and that for a single core, a stack of two and a stack of three cores. Stacking cores increases the A_L factor (inductance / turns ratio), which means that you can achieve a higher inductance with fewer turns, which is nice as the holes in the binocular are not very big.

Note that according to the calculations (using simplified mathematics) the attenuation keeps getting smaller as you increase the number of turns, which is not the case in real life as the higher number of turns creates higher capacitance and increased attenuation due to this increased capacitance. This is most evident on 7 MHz, of course.

Measurements were done using the AIM 4170 impedance analyzer, as well as on the N2PK vector network analyzer (**Fig 7-92**). A three stack of binocular cores *with just one turn* exhibits a transformation loss of approximately 1.1 dB on 160 meters (in a 75-Ω system), which is really not acceptable. Two turns is a minimum for a low-Z winding, also with a three stack of binocular cores.

Going by the measured insertion loss data, the lowest attenuation seems to be obtained with a four-turn low-Z winding on a single core, or a three-turn winding on a dual stack core. Watch out, in many cases we are talking about tenths of

Table 7-27
Transformer Attenuation Calculations vs Measurements
Attenuation for transformers made on a 73-mix binocular core (Fair-Rite #2873000202). The first figure was calculated while the second figure was obtained by measurement. All values for a 75 Ω impedance system.

	Freq (MHz)	*Loss (dB) 2 turns*	*Loss (dB) 3 turns*	*Loss (dB) 4 turns*
Single core	1.8	0.78 – 0.70	0.35 – 0.35	0.20 – 0.15
	3.5	0.75 – 0.70	0.34 – 0.35	0.19 – 0.20
	7	0.70 – 0.70	0.32 – 0.40	0.18 – 0.40
Dual core	1.8	0.39 – 0.30	0.18 – 0.20	0.10 – 0.10
	3.5	0.38 – 0.40	0.17 – 0.20	0.10 – 0.20
	7	0.36 – 0.60	0.16 – 0.40	0.09 – 0.40
Triple core	1.8	0.26 – 0.50	0.12 – 0.20	0.07 –0.10
	3.5	0.25 – 0.50	0.11 – 0.30	0.06 – 0.20
	7	0.24 – 0.50	0.11 – 0.60	0.06 – 0.40

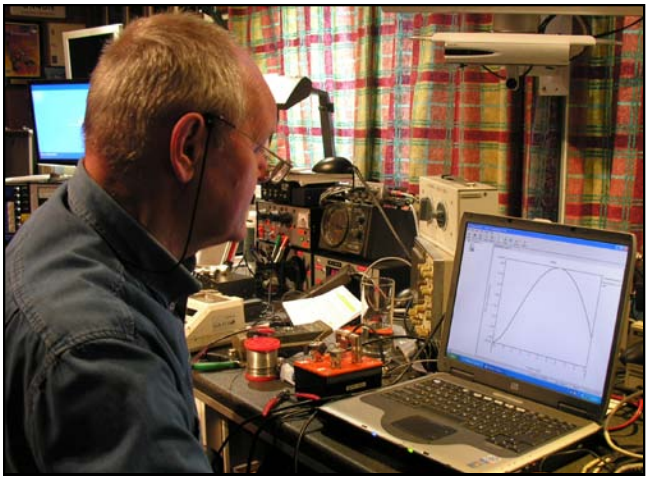


Fig 7-92 — Roger, ON6WU, testing a wide range of transformers suitable for feeding Beverage antennas. To measure insertion loss, two back-to-back connected transformers are tested on the N2PK vector network analyzer using the *Exeter* software by W8WWV.

Table 7-28
Transformer Turns Data

This table lists the secondary impedance for Z_{prim} = 50 Ω and Z_{prim} = 75 Ω as a function the primary and secondary turns. Using the binocular cores only 2, 3 and 4 turns can be used on the primary. The figures for 5 and 6 turns on the primary are for the MN8CX cores.

N_{prim}	N_{sec}	*Z ratio*	*50 Ω*	*75 Ω*
3	6	4.00	200	300
3	7	5.44	272	408
3	8	7.11	356	533
3	9	9.00	450	675
3	10	11.11	556	833
3	11	13.44	672	1008
4	8	4.00	200	300
4	9	5.06	253	380
4	10	6.25	313	469
4	11	7.56	378	567
4	12	9.00	450	675
4	13	10.56	528	792
4	14	12.25	613	919
4	15	14.06	703	1055
5	10	4.00	200	300
5	11	4.84	242	363
5	12	5.76	288	432
5	13	6.76	338	507
5	14	7.84	392	588
5	15	9.00	450	675
5	16	10.24	512	768
5	17	11.56	578	867
5	18	12.96	648	972
5	19	14.44	722	1083
6	12	4.00	200	300
6	13	4.69	235	352
6	14	5.44	272	408
6	15	6.25	313	469
6	16	7.11	356	533
6	17	8.03	401	602
6	18	9.00	450	675
6	19	10.03	501	752
6	20	11.11	556	833
6	21	12.25	613	919
6	22	13.44	672	1008

a dB of difference, which — from a practical point of view — is totally irrelevant. However, if you are happy with a couple of tenths of loss here and there, you may quickly find yourself confronted with a couple of dB of loss. It still is not really dramatic on the low bands, where it is not signal strength but signal-to-noise ratio that is important. But an engineer, if he can avoid losses, will do so.

From **Table 7-28** one can easily obtain the required number of primary and secondary turns for a given primary and secondary impedance. Data for both a 50-Ω and a 75-Ω primary impedance are given. The higher number of turns on the primary will result in a somewhat better performance at the lowest frequencies. Using a higher number can also have the advantage of allowing a more suitable impedance transformation ratio. Watch out: with a binocular (multi-aperture) core one turn equals two passes (the wire going through both holes)!

2.7.2.4. Measured Transformer Characteristics

You can read a lot on Beverage transformers, but almost nothing about their insertion loss! I made several transformers using the Fair-Rite 2873000202 type binocular cores (and others) and thoroughly tested them for insertion loss and SWR on the three frequencies of interest (1.8, 3.5 and 7 MHz). See **Table 7-29**.

Transformers were wound on single cores and on stacks of two and three such cores, and a range of transformation ratios were tested in a 75-Ω characteristic impedance test setup.

The SWR measurements were done using the AIM 4170 impedance analyzer, scanning from 1.7 to 7.3 MHz. In this test the transformer was terminated in its calculated load impedance, and the SWR values were registered for a range of transformation ratios and transformer designs.

The insertion loss was measured using the N2PK Vector Network Analyzer, whereby two transformers were connected back to back. In this setup the attenuation of the transformer is half the measured value.

Half Turns

I have — on several occasions — explained that, with a binocular transformer, one turn equals two passes (the wire going back and forth between the adjacent holes of the core). My curiosity however drove me to test transformers with "half turns" on the secondary, the goal being to have a more extended choice of impedance transformation ratios. As far as achieving more ratios, it works. The SWR is excellent with these "half turn" designs, but you pay for it with increased loss. The loss can increase by a factor of three to seven as compared with the transformer with half a turn more (or less) on the secondary. Whereas typical loss for a transformer with "full turns" ranges between 0.2 and 0.5 dB, the transformer with half turns on the secondary exhibits losses ranging from 0.8 to as high as 2 dB. Why this increased loss? The coupling from primary to secondary is not uniform. If the primary has 1 turn and secondary 2½ turns, part of the transformer has one wire coupling to three wires, and in the other part it is one wire coupling to two wires. The fewer the total turns on the secondary, the more outspoken this imbalance effect due to the half turn on the secondary.

Table 7-29
Insertion Loss and SWR for Transformers Wound on a Single Binocular Core

Freq (MHz)	*Primary Turns*	*Secondary Turns*	*Turns Ratio*	*Z Ratio*	*Z_{sec} for 75 Ω Z_{prim} (Ω)*	*Attenuation (dB)*	*Return Loss (dB)*	*SWR*
1.8	2	4	2.00	4.00	300	0.7	20	1.22
1.8	2	5	2.50	6.25	469	0.6	23	1.15
1.8	2	6	3.00	9.00	675	0.5	20	1.22
1.8	3	6	2.00	4.00	300	0.3	26	1.11
1.8	3	6.5*	2.17	4.69	352	0.9	26	1.11
1.8	3	7	2.33	5.44	408	0.2	27	1.09
1.8	3	7.5*	2.50	6.25	469	0.6	26	1.11
1.8	3	8	2.67	7.11	533	0.3	27	1.09
1.8	3	8.5*	2.83	8.03	602	0.6	27	1.09
1.8	3	9	3.00	9.00	675	0.5	28	1.08
1.8	3	10	3.33	11.11	833	0.6	28	1.08
1.8	4	8	2.00	4.00	300	0.2	30	1.07
1.8	4	9	2.25	5.06	380	0.2	32	1.05
1.8	4	10	2.50	6.25	469	0.2	32	1.05
1.8	4	11	2.75	7.56	567	0.2	33	1.05
1.8	4	12	3.00	9.00	675	0.2	33	1.05

*Note how the transformers wound with "half turn" secondaries exhibit more attenuation.

Single Core Transformers

Transformers were wound with two, three and four turns for the primary (low Z) and full turns for the secondary. All show losses between 0.2 and 0.7 dB (worst case). These worst cases are transformers using only two turns on the primary. These also exhibit the highest SWR. Three or four turns on the primary really does not make any significant difference in the frequency range of interest (1.8 to 7 MHz). Four turns gives the advantage of a more flexible secondary impedance range. In practice, after having measured the surge impedance of the Beverage antenna (Section 2.5.4) you can select the closest impedance from the series of transformers using either three or four turns on the primary (Table 7-28).

Dual Core Transformers

What do we gain with a dual core (two cores glued together)? Looking at 160 and 80 meters, single cores have approximately 0.2 dB less attenuation, but the difference is certainly not noticeable in practice. Whereas with a single core the SWR is as "high" as 1.09:1 (27 dB return loss), with a dual core it is 1.05:1 maximum. Again this is an insignificant difference. On 40 meters, insertion loss and SWR are almost exactly the same for a single core or dual core. All in all, the dual cores do not really offer a substantial advantage over the single core transformers.

Triple Core Transformers

A triple core does not score better than a dual core transformer under any circumstance. Triple cores have been advocated for use with just one turn on the primary (and three turns on the secondary) in order to achieve the lowest possible inter-winding capacitance. I measured this capacitance to be 4 pF, while a dual core with two turns on the primary and six turns on the secondary has less capacitance (3.4 pF, see Table 7-41 later in this chapter) and much less loss (0.2 dB vs 0.9 dB) and also a better SWR (1.05:1 vs 1.33:1). Conclusion: there is no valid reason for using a triple core binocular transformer.

Compensating the Transformers

In general one can obtain an even lower SWR than shown in Table 7-29 by putting a small capacitor (between 18 and 56 pF, depending on the type of transformer) on the primary, but this is certainly not meaningful. An exception might be on transformers used in an end-fire phased array system where you need the lowest possible SWR on the feed lines, part of which are used to obtain a given phase delay (see Section 2.16.5).

Conclusion

A transformer wound on a single binocular core (Fair-Rite Products # 2873000202) with either three or four turns on the primary (low Z) will give you a wide choice of secondary impedances to match (see Table 7-30 in the next section). A dual core transformer may yield a little better lab test results (0.1 dB less insertion loss and 5 dB better return loss), but in real life the difference is certainly not meaningful.

Don't forget that for the same number of turns a dual core still has almost double the inter-winding capacitance of the single core transformer (see Table 7-41 later in this chapter).

2.7.2.5. How to Wind Your Transformer

Considering the excellent results obtained with binocular cores, and the ease of winding this type of transformer, we will only give construction details for transformers wound on binocular cores.

In view of the possible problems associated with common-mode coupling (see Section 2.7.2.8), we shall only use transformers with galvanically separated primary and secondary windings.

If we use enameled wire for winding binocular cores, we must insert insulation tubes (preferably Teflon) in the holes to prevent the walls of the holes from scraping away insulation when winding the core. This will of course reduce the available hole diameter for feeding the wire through. Much better is to leave out the protection tube and use Teflon insulated wire #26 or #27 AWG wire which is readily available from many sources.

If you wind your own cores (toroidal or binocular) it is important to know what constitutes "one turn." With toroidal cores you count one turn each time the wire goes through the core opening. A binocular core has two holes. A wire though one hole is one pass. One complete turn is thus two passes through a binocular core. If you thread the wire in through one hole and bring it back through the other hole, you have one turn.

A binocular core does not leave you a choice of winding techniques. The binocular core transformer, shown in **Fig 7-93,** uses separate primary and secondary windings, but it is not possible to wind them in a specific way to minimize inter-winding capacitance. Tom, W8JI, wrote on this subject: "As for spacing the windings or using a Faraday shield, neither are necessary or useful. I have typically measured about 10 pF or less of capacitance with one winding laid directly over the other. That is more than 9 kΩ of leakage reactance, and that would easily put any common-mode well into the Beverage's noise floor." W8JI addresses the issue of reception of noise and spurious signals due to common-mode problems in detail on his Web site (**www.w8ji.com**). Further ways to reduce common-mode problems are discussed in detail in Section 2.7.2.8.

The transformer can be mounted in a small plastic box, located *by preference* at ground level. As far as the signal pick up by the vertical drop wire going to the horizontal Beverage wire, it does not make any difference whether the transformer is located near ground level (with a relatively long antenna drop wire) or at the top of the support pole near the antenna, in which case we have a ground wire that is equally as long and picks up the same amount of signal from all directions (see also Section 2.8). But it is preferable to keep the coaxial feed line on (even better: "in") the ground to maximize the choking effect of the nearby ground on the common-mode currents on the outside of the cable. If the coax runs up the 2-meter long pole to the transformer box located at antenna height, the outside of the coax shield acts like a 2-meter long vertical antenna!

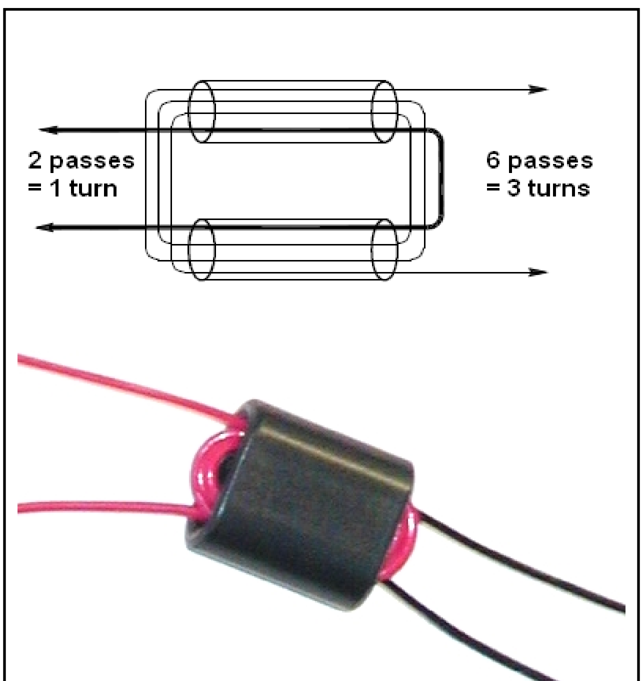


Fig 7-93 — Single core binocular transformer wound using Teflon insulated wire.

2.7.2.6. How Many Turns?

How accurate must the transformation ratio be? **Table 7-30** shows increments in load resistance of approximately 50 to 70 Ω. Let's assume the surge impedance of your Beverage is 500 Ω and your feed line has a characteristic impedance of 75 Ω. In such case you have the choice between a 4:10 turns ratio or a 3:8 ratio. In both cases your load impedance is about 35 Ω off. That means that the misalignment causes an SWR of 500/469 or 533/500 which, in both case causes an SWR of approximately 1.07:1 (30 dB return loss). This is totally negligible.

Table 7-30 gives the winding data for the most common impedance transformations, in both a 75-Ω and a 50-Ω system.

SWR on the feed line to a Beverage has two effects:

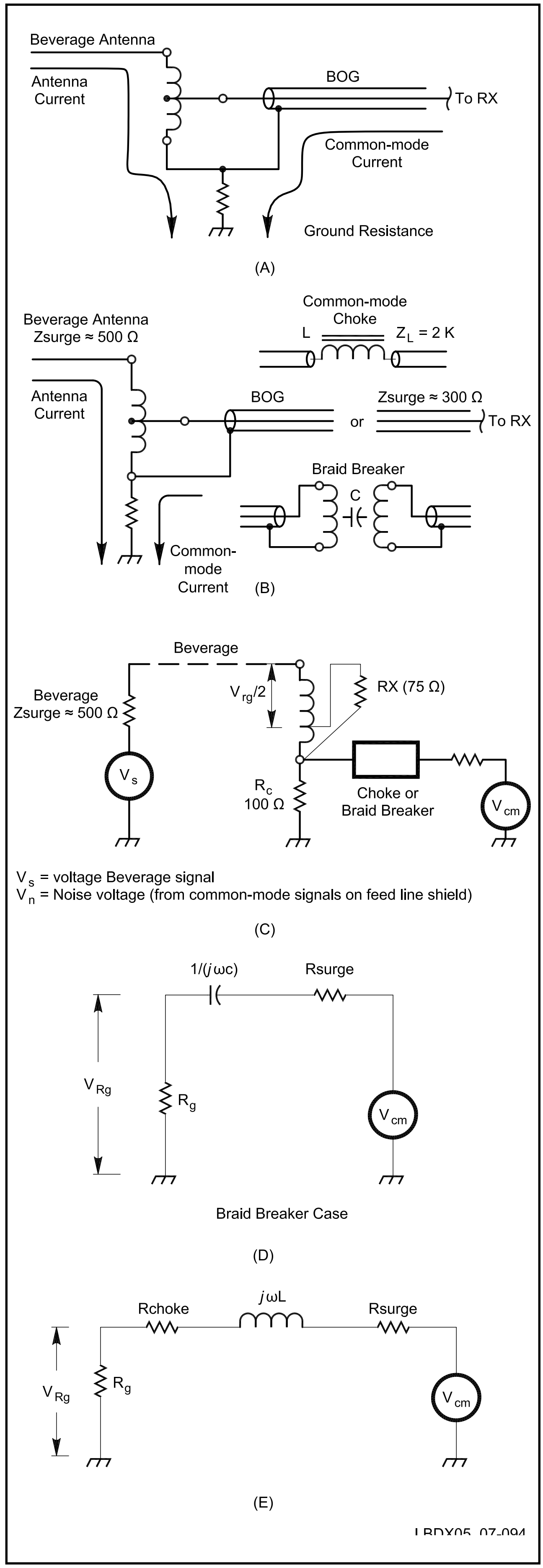
1) **Increased losses**: As long as the SWR is not more than 2:1, the losses will not be noticeable. The additional loss due to SWR on a 300-meter long run of ½-inch Hardline is approximately 0.5 dB, and that is unnoticeable. In short: You can live with an SWR up to 2:1 on a Beverage feed line, but an engineer would never tolerate that. I, as an engineer, would not want to see the SWR higher than 1.5:1 anyhow.
2) **Incorrect phasing delay**: If a length of feed line is used to achieve a certain phase shift (eg when using end-fire phased Beverages), a high SWR could severely upset the correct phase delay. The phase shift obtained from a length of coax only equals the coax length (both expressed in degrees) if the SWR is 1:1. See Table 7-7, where we have listed the phase shift versus cable length for the case of a load measuring 75 – j8 Ω (in a 75-Ω system), which represents an SWR of 1.1:1.

Table 7-30
Transformer Winding Data

Required passes for a transformer using a single Fair-Rite # 2873000202 binocular core (remember 1 turn = 2 passes).

Z_{prim} (Ω)	Z_{sec} (Ω)	*Primary Turns*	*Secondary Turns*
75	300	4	8
75	380	4	9
75	408	3	7
75	469	4	10
75	533	3	8
75	567	4	11
75	675	4	12
50	312	4	10
50	356	3	8
50	378	4	11
50	450	4	12
50	528	4	13
50	555	3	10
50	612	4	14
50	672	3	11

Fig 7-94 — A: Feeding the beverage with a common ground transformer, where common-mode signals coming from the shield of the feed line are added to signals from the Beverage in the common ground resistance. B: Inserting a braid breaker or common-mode choke reduces the level of the unwanted signals. C shows how the equivalent circuits (D and E) are generated. V_{rg} is calculated by taking into account the voltage divider made by $R_g/R_{surge} + 1/(j\omega C)$ for a braid breaker. For a common-mode choke, the voltage divider is made by $R_g/(R_{surge} + R_{coil} + j\omega L)$. See text for details.



Note that the phase angle is not off all that much and we can tolerate quite a few degrees of deviation in phase angle and variation in amplitude. All it will do is move the lobes (maxima and minima) around a little bit in the back of the antenna, but it will hardly influence the RDF and DMF. Conclusion: I would recommend to try to bring the SWR on feed lines going to elements of end-fire phased arrays down to 1.2:1 or better.

2.7.2.7. Checking Your Transformer's Performance

After winding a transformer it's a good idea to check its performance. Connect a noninductive load resistor (a small metal film will do) across the secondary. The resistor has the value of load impedance you made the transformer for. Check the impedance or SWR using a noise bridge or an antenna analyzer (or better yet, a network analyzer). With a well-made transformer the SWR should be less than 1.1:1 from 1.5 to >10 MHz.

W8JI describes another testing procedure: "Terminate the transformer with a resistor twice the normal secondary resistance and measure the SWR. Repeat with ½ the normal resistance. Ideally the SWR should be 2.0:1 in both cases. Multiply the two SWR reading together and take the square root. If it comes out close to 2, you have a pretty good transformer."

You can easily evaluate the insertion loss of a homemade transformer. Build two identical transformers and connect them back-to-back. Insert the back-to-back configuration in the feed line to your receiver. You should not be able to detect any change in signal strength, as two well-built transformers should exhibit less than 1 dB insertion loss, which is just about immeasurable without special test equipment. If you have access to professional test equipment you can do an actual insertion loss measurement in a 50-Ω or 75-Ω system. A network analyzer will also give you accurate results.

2.7.2.8. Connecting the Transformer to the Beverage and the Coax

It is important that *only* the signals coming from the Beverage antenna wire are transported by the feed line going to the receiver. If the Beverage were perfectly grounded, that would mean there would be no possibility for common-mode coupling. Since the ground connection is not perfect (read: "is not a zero ohm ground"), any wire connected to that ground (physically or by capacitive or inductive coupling) will, through the presence of common-mode currents on that wire, induce RF voltages across the ground resistance. Under such circumstances the ground rod is a common path through which not only the antenna current flows, but also common-mode currents caused by other paths connected to that same ground. The greater the earth resistance (the poorer the ground), the higher will be voltages induced through the "parasitic" paths. In **Fig 7-94C**, V_{rg} is the voltage that results from current flowing on the feed line. $V_{rg}/2$ is the voltage then across the transformer secondary which is directly then sent to the receiver input. Thus lowering V_{rg} will lower the level of these undesired signals into the receiver.

The outside of the coaxial feed line acts as a very low (on the ground) lossy long-wire or BOG (Beverage on Ground), receiving signals and noise from directions other than those you are interested in. This is why you should never "hang" your Beverage feed line *in the air* as this would make it a better antenna. You should lay it *on the ground* or better yet, bury it *in the ground*. This helps to choke RF currents to the surrounding ground.

Using a transmission line type transformer (or auto-transformer)

The common-mode signals flowing on the outside of the feed line can be fed into the inside of the feed line through coupling via the common ground resistance as shown in Fig 7-94A. This is a common cause of spurious signals and noise reception, especially if long unburied feed lines are involved.

One way to reduce the unwanted signal ingress is to reduce the equivalent ground resistance (R_g), which is not always easy to do. Using a ground system with R_g = 100 Ω yields an attenuation of approximately 12 dB (that's not very much) for common-mode signals. Improving the ground system so that R_g = 25 Ω (that's a really good ground for a receiving antenna) still only yields 22 dB suppression of the unwanted signals. A more efficient way to reduce the level of the unwanted signals is to insert a *common-mode choke* (see Section 2.7.2.9) or a *braid breaker* (see Section 2.7.2.10) in the feed line near the antenna transformer (see Fig 7-94B). Common-mode chokes are widely used on transmit antenna feed lines (where they are commonly called "current baluns") as well as on feed lines to receiving antennas, where they are called "common-mode chokes").

Fig 7-94C shows how we come to the equivalent circuits shown in D and E, which let us easily calculate the performance of the circuit. Using the braid breaker (transformer), and assuming R_g = 100 Ω, R_{surge} (snake) = 300 Ω, and using a transformer with an inter-winding capacitance of C = 10 pF (easily achievable, see Table 7-41 later in this chapter) which results in $X_C = 10^6/2\varpi fC$ = 9 kΩ on 1.8 MHz, the attenuation achieved is 20 log (100/100+300+9000) = 39 dB. Using a common mode choke with R_{coil} (loss resistance) + $j\Omega L$ = 2 kΩ (that is a very good choke!), the attenuation is: 20 log (100/100+2000) = 26 dB**.** It is clear that in this situation the braid breaker (1:1 transformer) provides much better attenuation.

Using a transformer with galvanically separated windings

A transformer with galvanically separated primary and secondary already includes a braid breaker. The transformer now serves to match impedances and at the same time serves as a braid breaker as described above (see **Fig 7-95**).

Common-mode signals on the outside of the coax now travel through the inter-winding capacitance C of the transformer toward the ground rod. As is the case with the braid breaker transformer, there is a voltage divider involved, consisting of X_C (the reactance of the inter-winding capacitance) and R_{g1} (the antenna ground system resistance). In this case the use of the split-winding transformer achieves an attenuation of 39 dB, the same figure as calculated above for the braid-breaker alternative (assuming the same input data).

We can get more common-mode signal attenuation by adding a second ground rod, which now grounds the coax

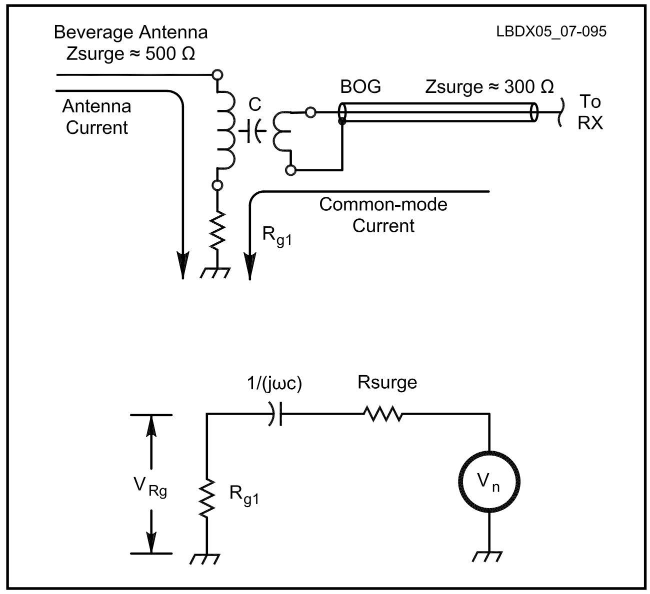


Fig 7-95 — Principle schematic and equivalent circuit for a setup using a transformer with galvanically separated windings.

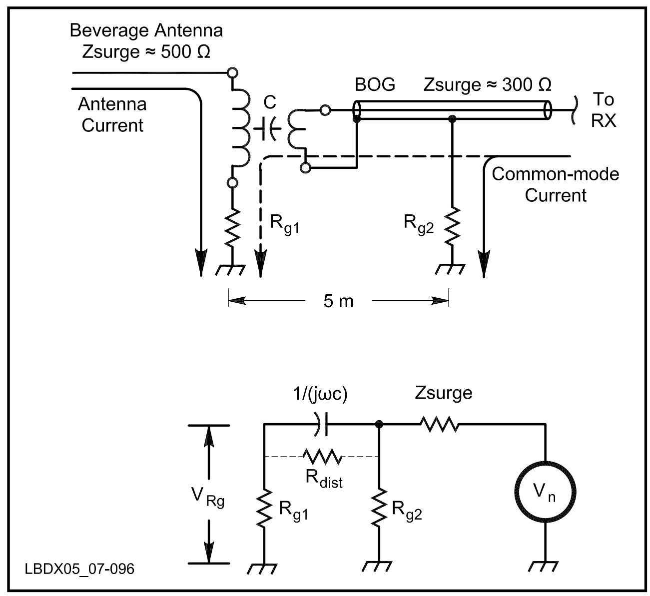


Fig 7-96 — The addition of a separate ground rod for the feed line shield is an easy way to improve the rejection. Make sure the antenna ground rod and the feed line earthling ground rod are separated enough to prevent R_{dist} from becoming too low.

shield as shown in **Fig 7-96**. In order to prevent ground currents from this second ground rod from being coupled into the antenna ground rod (R_{g1}), it is advisable to keep the two ground rods separated by approximately 5 meters. In this case we have two voltage attenuators in series, one consisting of Z_{surge} (BOG) with R_{g2}, and the second one consisting of X_C (= $1/j\Omega C$) and R_{g1}. Let us assume R_{g1} and R_{g2} to be 100 Ω and X_C = 9 kΩ. The total attenuation is now 20 log (100/ 9100) + 20 log (100/100+300) = 51 dB.

If this is not enough, we can go one step further and use a common-mode choke (see Section 2.7.2.9) installed in the feed line at the feed line ground. Make sure the common-mode choke is installed beyond the feed line ground looking toward the receiver.

Fig 7-97 shows such a setup. The total attenuation becomes: 20 log (100 / (100 + 300 + 2000)) + 20 log (100 / 9000) = 67 dB (assuming Z_{surge} = 300 Ω, R_{g1} = R_{g2} = 100 Ω, X_C = 9000 Ω and $R_{coil} + j\Omega L$ = 2000 Ω).

Another alternative is shown in **Fig 7-98**, where we use a 1:1 impedance ratio transformer to act as a so-called *braid breaker* at the second ground rod grounding the transmission line. As the leakage impedance caused by the inter-winding capacitance C (typically >5 kΩ) is even higher than the total series impedance for a common-mode choke (typically ≤2 kΩ) the total attenuation is now even higher: 20 log (100 / 100 + 300 + 9000) + 20 log (100 / 9000 + 300) = 78 dB (assuming X_C = 9000 Ω).

I would suggest *not* using a common-mode choke in conjunction with a transformer with galvanically separated windings *if there is no ground rod in between the two*. It could happen that $j\Omega L = 1/(j\Omega C)$, in which case the common mode choke coil and the capacitor (inter-winding capacitance in the transformer) are series resonant, which represents a short, resulting in *zero* attenuation of the common-mode signals.

The numbers representing the attenuation of the common-mode signals in the various examples are based on a simplified model, and may not represent real life under all circumstances. They should give a fairly good idea what can be done, and what degree of improvement can be achieved by using various techniques.

Conclusion

I would suggest always using a transformer with galvanically separated primary and secondary windings. If that is not enough to suppress common-mode signal ingress, ground your feed line some distance (≥5 meters) from the Beverage antenna ground rod. If you still need more, try to insert a common-mode choke or better yet, a braid-breaker near the feed line grounding rod.

2.7.2.9. Common Mode Chokes

You can make a common mode choke in different ways. For the application we have in mind we need the choke that has impedance of at least 1 kΩ on the operating frequency.

One normally uses ferrite material to make such chokes. The exact ferrite core material is not critical for this purpose, but a high permeability is required. A low-Q situation is preferred (lots of resistance in the impedance number) to avoid resonance effects, but if the coax lays on the ground resonance effects are really excluded.

One way is to use a stack of small beads on the coax. For example, 100 FB73-2401 beads on a piece of RG-58 or RG-59 yields an impedance of approximately 1200 + *j*950 Ω (equivalent to Z ~ 1.5 kΩ) on 160 meters. Z = 1.9 kΩ on 80 meters and Z = 1.8 kΩ on 40 meters. One hundred such small beads represent a stack almost two feet long, and will set you back approximately $20 (**www.thewireman.com**).

From the Amidon catalog it appears that seven turns of miniature coax through a FT-150A-F core (μ = 3000 and A_L = 5000) achieves an impedance of 3 kΩ on 160 meters,

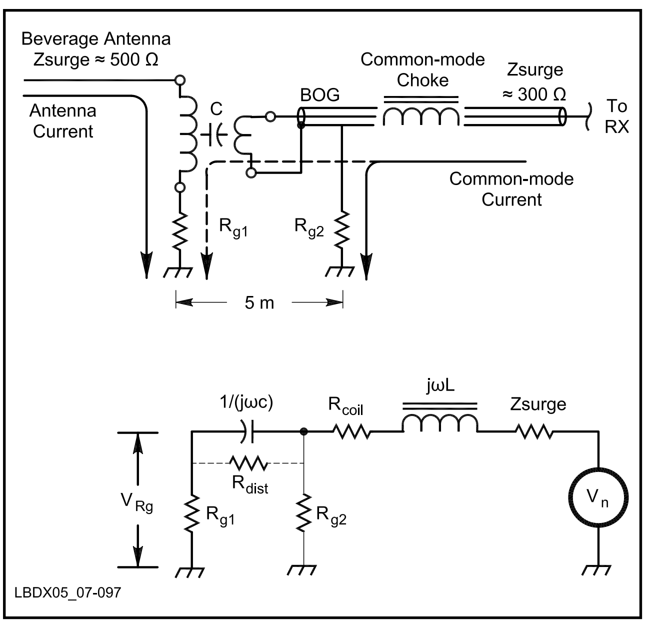


Fig 7-97 — Adding a common-mode choke next to the feed line ground rod is a common way to further improve common-mode signal rejection. See text for details.

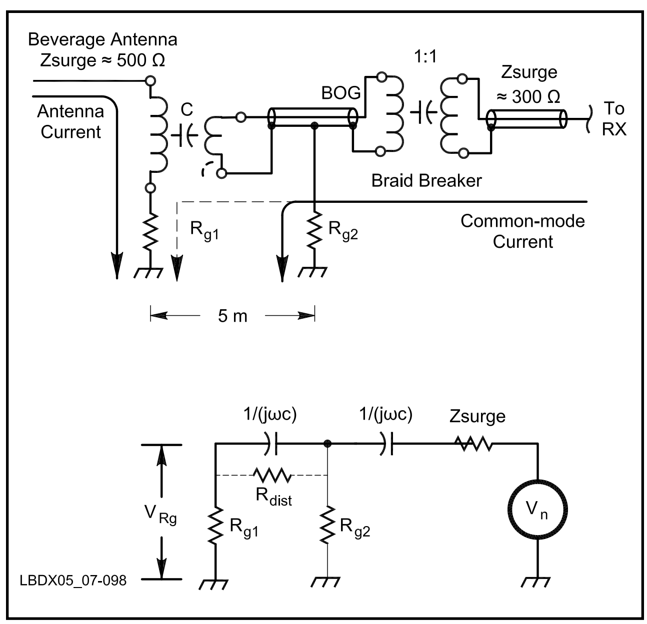


Fig 7-98 — Instead of using a common-mode choke, a braid breaker can be used to insert a high impedance for the common-mode signals. See text for details.

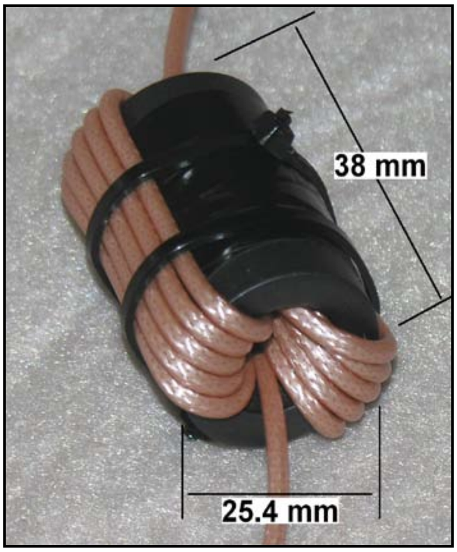


Fig 7-99 — A high performance RFC common-mode filter as used throughout the Beverage antenna installation at ON4UN. See text for details

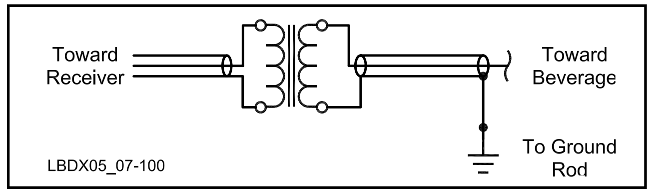


Fig 7-100 — The braid breaker box has two coax receptacles and a terminal to connect to the ground rod. Make sure that the coax receptacle accepting the coaxial cable leading to the receiver is insulated from the box if the braid breaker is mounted in a metal box. See text for transformer winding data.

which would be excellent.

I made my common-mode chokes by winding a 120 cm (4 ft) length of RG-179 (75 Ω, 2.5 mm OD) Teflon insulated coaxial cable on a stack of three 1-inch (25.4 mm) ferrites (unknown source) wound in a split-winding fashion (see **Fig 7-99**). The impedance on 1.8 MHz is 4 kΩ, with a resistive part of 1.5 kΩ, On 3.5 MHz Z_{tot} = 9 kΩ (R part = 7kΩ). On 7 MHz Z_{tot} = 2.5 kΩ (data measured using the AIM 4170 antenna analyzer). This makes it a good common-mode choke.

I have made about 25 such common-mode chokes that are used together with a ground rod approximately 5 meters from each Beverage feed point and at each head end or distribution box containing relays.

2.7.2.10. Braid Breakers

A *braid breaker* is a 1:1 transformer with galvanically separated windings (**Fig 7-100**). Important performance parameters are *loss* which should be kept to a minimum, and *inter-winding capacitance* which should be kept as low as possible.

I found that such a transformer wound on a stack of two Fair-Rite # 2873000202 binocular cores gives very good results. If we use two turns for primary and secondary, the insertion loss is 0.17 dB on 1.8 MHz, 0.2 dB on 3.5 MHz and 0.28 dB on 7 MHz. Using three turns on both windings, the loss is 0.1 dB on 160, 0.15 dB on 80 and 0.25 dB on 7 MHz. The differences are not very meaningful.

We can apply a few assembly tricks in order to minimize the capacitance between the two identical windings. I first insert two small (2.4 mm OD) Teflon insulating tubes through each of the two holes of the stacked binocular cores. Next I wind the primary and the secondary in a separate pair of Teflon tubes. Using #26 AWG Teflon insulated wires, the assembly goes smoothly as the Teflon insulated wires and the Teflon tubes take care of lubrication. Having the two windings in separate channels minimizes the inter-winding capacitance. For a 2 × 2 windings transformer I measured 8 pF (equivalent to 11 kΩ on 160 meters), for a 3 × 3 windings 16 pF (Z = 5.5 kΩ on 160 meters).

DX Engineering (**www.dxengineering.com**) sells common-mode chokes (model RFCC-1). These are the braid breaker type, using a 1:1 impedance transformer on a binocular core, just as explained in the text above.

2.7.2.11. What Else Can We Do to Suppress Feed Line Common-Mode Currents?

Assume you have made the right transformer, installed the correct ground system as well as taken all precautions to get rid of common-mode signals (using a common-mode choke plus ground rod or bread-breaker plus ground rod). What else should we be aware of to build a top notch Beverage antenna system?

Designing an Optimized Feed and Distribution System

Measuring and checking for ingress of common-mode and other noise signals into the feed line system is very difficult. The best thing is to design the feed system from the start to follow all possible rules of good engineering and to include all possible techniques to minimize the problem.

Over the many years my Beverage antenna system has evolved from a system that did not include any such sound engineering, to a system that, to my knowledge, now is state-of-the-art. The results of these efforts have been awesome. Together with using end-fire phased Beverages (see Section 2.16.3), building a "silent" and reliable distribution and feed system for my 12 Beverages has yielded me a very important improvement in signal-to-noise ratio. I now can dig much deeper into the noise to work the really weak stations.

Receiver Ground

Make sure the receiver itself has a good RF ground and is not just "grounded" through the shield of one or more coaxial feed lines. Make sure the coax is well-grounded where it enters the shack. Just relying on grounding at the receiver chassis is bad practice. It is quite common that the power mains feeding your receiver are carrying conducted RFI. This RFI is bypassed to the receiver chassis through any mains decoupling capacitors and the shield of the cable becomes the new path for this unwanted noise to leak via the cable shield all the way back to the feed point of the Beverage.

It is essential to feed the equipment at the shack through high-quality mains filters and ground these to an excellent RF ground. The bottom side of the operating table (which is 8 meters long) in my shack is completely covered with aluminum sheet. This represents a lot of capacitance and virtually zero inductance, which is just what you want! Quality mains filters are bolted directly to those sheets and the mains outlet to which the equipment is connected as well. The ground plane is connected with very short (less than 0.5 meter) low-inductance wide straps to long copper ground rods. Ground the feed lines to *another* good quality ground system where they leave the house of the shack. Ground rods for this ground should be a minimum of 5 meters from the ground rods grounding your equipment.

Different Beverage Antennas from One Hub

Never run different Beverage antennas from a single spot, using a single ground rod (or ground system). In such a case you will, via the common ground rod resistance, cross-couple part of the signals from one Beverage into another Beverage.

If you want to run different Beverages from one point, run the master feed line to a switching box (head end of the master feed line), and fan out to various feed points, each of which is fed via a separate feed line, in such a way that the ground rods of these feed points are separated at least 5 meters from one another.

Feed line Switching and Distribution

Fig 7-101 shows a feed line switching and distribution box designed to reduce common-mode coupling to a minimum. Five hermetically sealed relays (see Section 1.23) are used to switch the main feed line to one of five shorter feed lines going to Beverage feed boxes in the vicinity of the distribution box. These feed boxes are separated at least 5 meters from one another. Small vacuum relays are used in this example, only because they were obtained at a bargain price, but reed relays would do the job as well.

The cable connectors for all coaxial cables (the main feed and the five distribution lines) are insulated from ground and from one another. Via the common-mode chokes, the shields are connected to a common ground. The ground consists of a few ground rods (estimated total ground system resistance is 50 Ω).

Fig 7-101 — Coaxial feed line distribution box for Beverage system. A ⅞-inch main feed line (180 meters from the shack) is fanned out to five smaller (½ inch) feed lines. Notice the small vacuum relays and six common-mode chokes used.

Make sure that the control cable carrying the control voltages for the relays also passes via a common-mode choke.

The Main Feed Line (Trunk)

It is highly recommended that you ground the shield of a long feed line (>50 meters) somewhere near the center of its stretch. If very long feed lines are used, several grounding rods can be used if necessary. Grounding should be done to independent ground rods; that means ground rods that are not connected to anything else. As an additional benefit, of course, lightning paths are disrupted by this method.

Bury your trunk coax. *Never* suspend the feed line to a Beverage antenna in the air! Feed lines on the ground, or even better *in* the ground, will automatically choke off RF from the outside of the shield.

Do not run feed lines parallel to and in close proximity to elevated radials. Otherwise there will be field coupling from induction or radiation fields. Reduce coupling to other antennas (see also Section 2.11.1), towers, power lines, etc by careful placement of the antennas and by judicious feed-line routing.

All of these measures may not be necessary, but in stubborn cases the combination of all of these will guarantee a minimum amount of common-mode RF current on the outside of your feed line. If your feed line is not properly decoupled it can upset the effective directivity pattern of your Beverage and turn a fantastic receiving antenna into a mediocre one.

When all of this is done ...

You should now have a "quiet" feed system to the Beverage(s). Connecting everything to the system except for the Beverage wire itself should yield no signals.

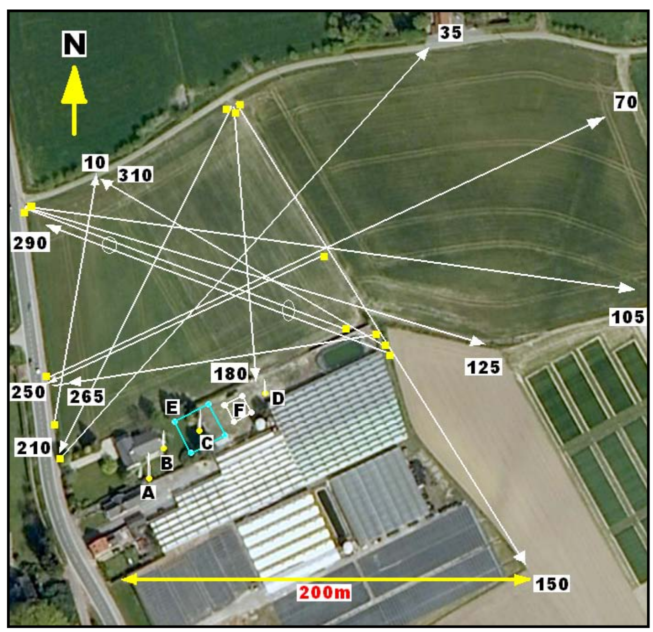


Fig 7-102 — Beverage antennas at ON4UN (from October through March). The square blocks show the Beverage feed points. Further antennas are: A: tower with 5-element 20 meter Yagi and 3-element 40 meter Yagi; B: tower with 6-element 10 meter and 6-element 15 meter Yagi; C: 39 meter tall quarter-wave vertical for 160 meters, also supporting the 80 meter Four Square(E); D: tower with C31XR 10-15-20 meter Yagi, F: 40 meter Four Square.

The Beverages at ON4UN

A friendly neighbor farmer allows the author to use his farming land (approximately 10 acres) to put up a Beverage system consisting of 12 antennas, ranging between 150 and 300 meters long.

Fig 7-102 shows the layout. The 290° antenna (to the USA) consists of two end-fire phase Beverages measuring 160 meters long (see Section 2.16.3).

Implementing a distribution and feed system that takes into account all the sound engineering rules regarding suppression of common-mode signals (see above) has yielded a high performance receiving antenna system. I have learned over the years that Beverages are much more than just a wire and some sort of feed line!

There are a total of five distribution/coax switch boxes. Each one is equipped with relays that have hermetically sealed contacts (vacuum relays or reed relays) in order to provide long lasting low contact resistance. In the schematic shown in **Fig 7-103** these are represented as simple multipole switches to keep the diagram simple. Each port (in and out) is equipped with common-mode filters, either chokes or braid breakers (see Fig 7-99 and Fig 7-100). Beyond these common-mode chokes, the screens are connected together and to a 1.5-meter long ground rod.

Each of the longer coaxial cables runs connecting these distribution boxes has its shield connected to a ground rod approximately in the middle of the run. This provides another path to ground for common-mode noise The shorter cables running from the distribution boxes to the transformer boxes are grounded approximately 5 meters from the antenna transformer box where a common mode filter (either choke or braid breaker) is installed.

Each of the ground symbols shown indicates a separate ground rod that is at least 5 meters from the closest ground rod. This prevents signals from coupling through mutual impedance via a single rod or via rods that are too close together.

2.8. Vertical Down-Lead or Sloping Beverage Terminations

A common way of terminating the single-wire Beverage is to connect the proper terminating resistor between the end of the Beverage and the ground. The termination system using two in-line quarter-wave radials as ground (see Fig 7-66B) is only a good solution for modeling a Beverage, and certainly not a practical solution. In addition, while two such in-line radials offer a perfect nonradiating ground in a computer model, these two quarter-wave wires (in-line), over real (not perfect) ground, with varying ground characteristics along their length,

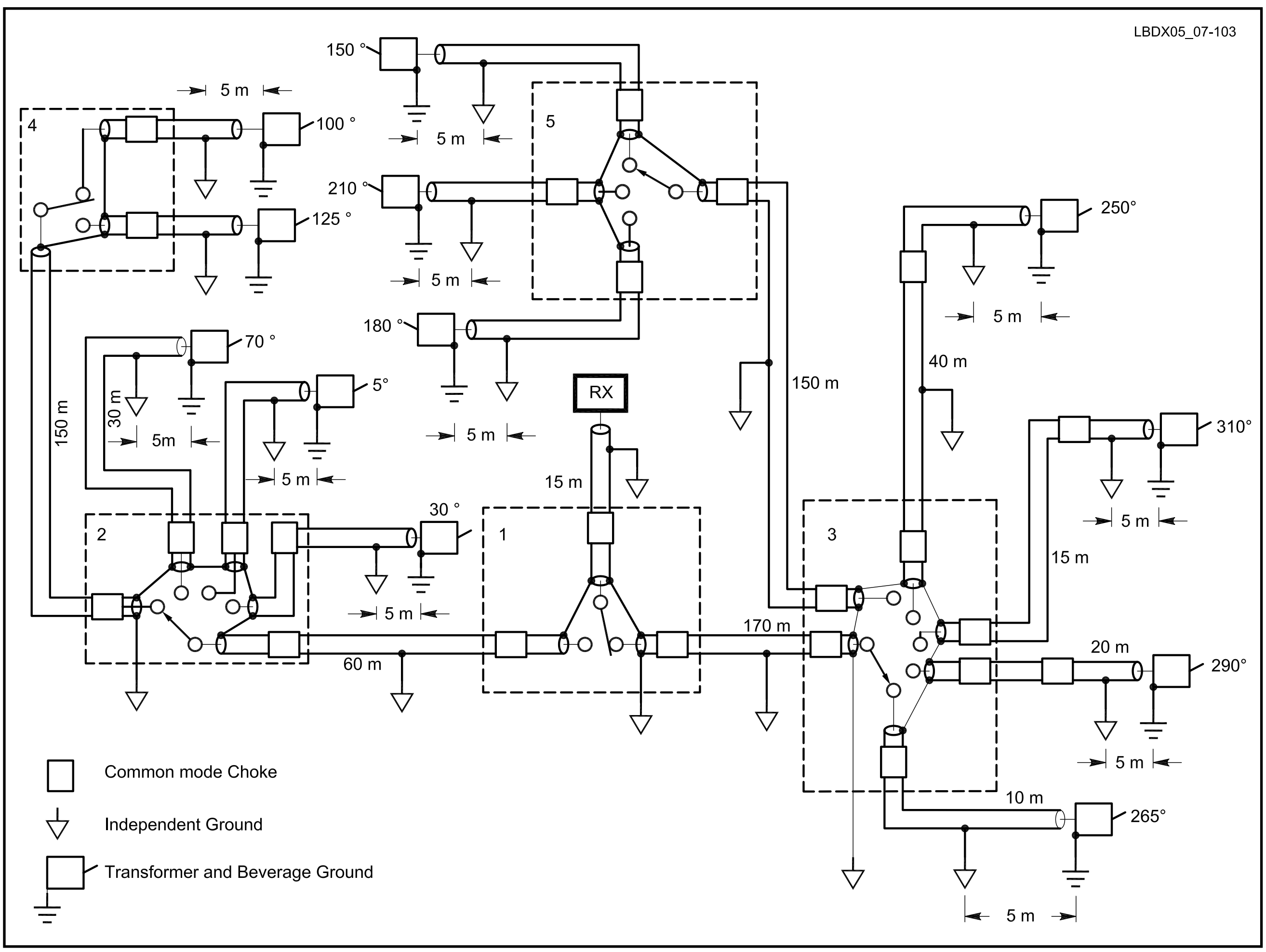


Fig 7-103 — Schematic showing feed points, feed point grounds, cable grounds and common-mode filters for the ON4UN Beverage system. All ground rods are spaced at least 5 meters from one another.

will pick up signals themselves. In a transmitting model they would radiate. The mechanism is similar to the one described in Chapter 9 dealing with low elevated radials and radiation from these radials. Also, such a system is a single frequency system!

Over the years, some have attributed magic properties to sloping terminations (see **Figs 7-104** and **7-105**), since the vertical component of the down lead supposedly *disappears*. Many of us slope the Beverage wire down from its nominal height to a ground stake, where the termination resistor is attached between the sloping antenna end and the ground. The ground stake serves two purposes: it is the electrical ground and the anchoring post for the Beverage wire.

Fig 7-104 — Sloping end section of the 300 meter long Beverage in the direction of Japan at ON4UN. The sloping section is approximately 15 meters long.

There is nothing wrong by doing this, but such a sloping wire does not pick up fewer signals than a vertical wire of the same height. It doesn't matter whether you slope the last 20 meters of the Beverage or continue 18 meters of it horizontal and run 2 meters straight down. Either way you have 2 meters of vertical distance. A vertically-polarized wavefront arriving from the side will see only the 2-meter high vertical component of the sloping end, the same as with a straight vertical wire.

The best proof that a sloping wire works like a vertical wire is the fact that a Pennant antenna (or Delta-shaped loop; see Sections 3.4 and 3.5) works. The Pennant, despite having one end sloped and one end vertical, has nearly identical vertical sensitivity from both ends. This clearly proves that the sloped wire behaves almost exactly like a vertical of the same height, or else the Pennant would have a 0 dB F/B ratio.

Tom, W8JI, explains: "The vertical end-coupling of Beverages is a mostly non-problem that has been over rated, and the 'cures' are mostly non-cures for a non-problem. One way to look at it is that six feet of vertical drop is six feet of vertical drop, no matter how the 'six foot drop' is spread around, and that isn't any big deal when the antenna is several hundreds of feet long. If that isn't reason enough not to worry, the entire antenna responds to vertical signals anyway...especially on ground wave!

"Second, it is physically and electrically impossible to 'shield' the vertical end lead no matter what scheme you employ. The antenna *must* always have the same net common-mode current flowing to ground over the vertical lead distance to ground. The only way possible to prevent that effect is to move the entire ground system up to the element height, and that means work with a bulldozer making a large mound or installing a large counterpoise hanging in the air at antenna height. Only those options can prevent common-mode current from flowing to ground!

"Some Beverage books will give you the idea that a particular scheme does something to 'shield' the vertical end wire, but it does not. On the outside of the tubing, we would measure exactly the same common-mode current as the end-

Fig 7-105 — Sloping down-lead termination for one of the Beverages at ON4UN. A stainless steel pipe supports the feed box that contains the binocular transformer and the common-mode choke. Note the ½ inch 75-Ω feed line and the professional grade connectors used. To the left we see the top of a 15 mm diameter copper rod driven into the ground, to which a tinned copper strap is connected. This provides the ground connection to the transformer.

current in the antenna. The vertical wire carrying current in the center of the tube induces an exactly equal opposing flow on the wall, that spills over at the open end and flows down the outside of the tube. Not that it matters, since that radiation is generally insignificant."

John, K9DX, did a well-controlled test to assess the difference between sloping terminations and terminations with a vertical down lead. He put up two widely separated 2-λ Beverages in parallel. The reference one had two in-line quarter-wave radials (see Fig 7-73C), which in theory precluded any omnidirectional pick up.

The test antennas had either long sloping ends or simple vertical down leads. John concluded from extensive testing that there was absolutely no difference at all between sloping and vertical feeds. Shielding the vertical down lead also does *not* work.

All in all, John came to the conclusion that: "… the vertical feed (fully vertical or sloping), as poor as it is, seems to hear as well as the raised feed. The noise picked up by the Beverage in an omnidirectional noise environment appears dominated by the front lobe. If you want better signal-to-noise, reduce the beamwidth. Going from 15 to 30 dB off the sides or back won't do much." All of this, of course, considers "noise" coming in equal strength from all directions. In very quiet locations, such as at K9DX and W8JI, noise is evenly distributed in all directions and at all angles. Under these circumstances looking at the RDF is much more important than looking at higher noise suppression in specific directions

I would like to argue that narrowing the forward lobe beamwidth is the only way to obtain a better signal-to-noise ratio, as it is the case when noise is evenly distributed in all directions. Living in a place where the noise is *not* evenly distributed in all directions, I have a different experience. That is also why I developed the concept of DMF (see Sections 1.8 and 1.10). I know from my own experience that going from a 19 dB DMF figure (for a single 300-meter long Beverage) to a 34 dB DMF for an array consisting of two such Beverages in an end-fire configuration, improves the S/N ratio at my QTH a great deal (see also Table 7-36 later in this chapter). I can perfectly read weak signals on my end-fire array, signals that I cannot even detect on the single Beverage. That is because in Western Europe, where I live, most of the noise is coming from the back when "looking "at North America. In this case the back is where most of the QRM, splatter and clicks come from. At the front there is only the ocean, and the first man-made noise generators are 6000 km away.

The sloping termination has been a subject of discussions and opinions for a long time. Let me add mine. Whether or not you use a perfectly vertical or a sloping termination wire, these wires pick up vertically polarized waves from all directions. Envision a Beverage with the two vertical termination wires separated by a number of half waves (considering the velocity factor of the antenna). The signals induced on both verticals will be out-of-phase when arriving at the receiving end. This may lead to so-called optimum lengths, where the ill effects of the vertical down leads are partially mitigated. (There still is the loss in the Beverage as a transmission line resulting in incomplete cancellation.)

This only works in one direction, and for the 180° example given, assumes a wave angle coming at a right angle to the Beverage. (The signal arrives at the two verticals in-phase, and will be added out-of-phase because of the delay in the Beverage acting as a transmission line.) So in this particular case the front-to-side ratio would be improved. For other lengths of Beverages, other directions will be nulled out. In other directions the noise may add and the directivity may go down. The canceling mechanism works along the same principles explained in Section 1.6.

Sloping ends can be considered as a number of short verticals, placed in slightly different locations (along the sloping end) so we cannot really talk about a single separation distance between the two sloping ends — it is smeared out. If you now use a Beverage that consists only of sloping ends (one central high post and two sloping wires, each being half the length of the vertical), you will have the disturbing effect of many very small verticals spread all along the Beverage. Depending on the direction of the noise, the contributing EMFs may add or subtract, and in a long Beverage all of this happens at the same time. The net effect is a general, small decrease in the overall directivity in all directions, without creating any specific nulls or maxima for any particular direction. I use this Beverage model (I call it the inverted-V shaped Beverage) when modeling Beverage arrays (Section 2.16) and find that the directivity patterns are very similar to those modeled using the T-terminations with two λ/4 in-line radials (Fig 7-73C).

If you add the above reasoning to the sensitivity analysis done by W8JI, you can conclude that whereas vertical feeds in almost all cases are not a *real* problem, long sloping ends because of the phase distribution are even better. Long sloping terminations can indeed help reduce the ill effects of the vertical distance involved in the down leads.

By using Beverages in a broadside or end-fire array configuration, the effects of vertical down leads can also be greatly compensated (see Section 2.16.3), at least for specific directions. A broadside/end-fire array is even better.

Sloping ends, however, have a negative effect as well because the source impedance of the Beverage is not constant. Whereas with a perfectly "flat" Beverage we can obtain an SWR that is below 2:1 over the three low frequency bands, in the case of a Beverage with sloping ends the wave is more outspoken. The SWR peaks at approximately 1.5:1, which is not a real problem for single Beverages, but which could be annoying in case of end-fire phased Beverages where a feed line impedance exhibiting an SWR of at most 1.2:1 is required to obtain a reasonably correct phase delay in the coaxial feed line (see also Section 1.18).

Conclusion on sloping terminations: If they suit you better, use them, except when the Beverages are part of an end-fire phased array. In that case, a very low and flat SWR curve is required to obtain the correct directivity.

2.9. Beverage Wire

A Beverage is a lossy antenna. There is no point in striving for the lowest possible conductor loss. Any type of wire that is mechanically suitable should do the job. I have used bare copper wire, enameled copper wire and PVC-insulated copper wire, with the copper having a diameter anywhere from 0.6 to 1 mm on my Beverages.

Soft-drawn copper wire cannot be tensioned to any great degree without stretching and it is not suitable for very long unsupported spans of wire. If you don't mind using many supports, however, soft wire may be used for the antenna. Currently

I use mostly soft-drawn single strand copper wire of 1 mm OD (#18 AWG) for my wintertime Beverages, where the bamboo posts are separated by approximately 20 meters.

For permanent Beverages separated by distant posts I use bronze- and copper-clad steel wire (1.6 mm OD, #14 AWG), which makes it possible to cover 100-meter stretches without intermediate posts. I pull this tight with approximately 45-kg tension. Copper-clad steel serves the same purpose, if you can get hold of it.

Insulated wire is potentially better than a bare conductor when the antenna wire touches branches or brushes against leaves in the wind. On the other hand, insulation may hide a broken conductor unless the antenna wire is under significant tension.

Some people have used surplus telephone-pair wire. This wire is often available at hamfests. Don't try to separate the wires, just connect them in parallel at both ends. Using a pair as a Beverage has the advantage that you can check continuity from one end using an ohmmeter. All connections must be soldered, preferably with silver solder. Common electronics solder rots away after years in the outdoors. If you use regular solder, cover the solder joints with liquid rubber or a dollop of petroleum jelly.

Long Beverages made of thin soft-drawn copper wire can stretch in high wind. A solution is to use a well-tensioned bungee cord at the end of the Beverage. Keep the Beverage wire as horizontal as possible; Beverages with lots of sag exhibit a surge impedance that is very unstable, which results in a deep variation in SWR depending on frequency.

Aluminum wire is widely used as electric-fence wire and makes for excellent Beverage wire. It is lightweight and strong and readily available from outlets selling farming equipment. It is commonly available in diameters from 1.15 mm (#17 AWG) to 1.65 mm (#14 AWG) from various sources.

Earl, K6SE (SK), used galvanized-steel wire. Earl claimed that the use of this high-loss resistance wire improves the F/B, since the signal from the reverse (unwanted) direction now travels down the lossy wire twice to reach the receiver — once toward the terminated end and then whatever signal is reflected from the termination must travel down the wire again toward the receiver. Signals coming from the forward (desired) direction must travel down the lossy wire only once.

Tom, W8JI, rightly points out: "Steel fence wire would aggravate losses that already limit the benefits of using long Beverage antennas. In a very long antenna, the small additional loss of steel fence wire might slightly reduce performance."

Please remember that one of the most important considerations in the selection of wire for a Beverage is maintenance!

2.10. Supports, Poles, Insulators and Other Hardware

You can use any convenient support for a permanently installed Beverage: tree trunks (use nail-type electric-fence insulators), wooden poles, metal posts, plastic tubes, etc. Electric fence hardware is cheap and suits the purpose. I always use the in-between posts only to support the wire, letting the wire freely slide through an insulator. This way tension will equalize in all stretches between supports. If at one place the copper wire is stretched by accident, this will smooth out over a longer stretch of wire. If the wire breaks, you will also see it at any point along the antenna. A long piece of wire is also much more difficult to break. Soft copper will typically stretch 25% before breaking! If someone accidentally walks into a taut, short span of wire, he will likely break it and may hurt himself. A long span of wire will stretch without breaking and not hurt the "offender."

I can only have my Beverages up in winter time, so I need an easy, lightweight, flexible and fast-to-deploy system for supporting my Beverage wires. I use 8-foot (240 cm) long bamboo poles that are reasonably cheap if you buy them in large quantities at wholesale garden outlets. The ones I use have a diameter of approximately 10 mm at the top and 20 mm at the bottom. They last about four years before they rot at the bottom. I insert a fence ring insulator from Gallagher in the top, and fix it with tape. (See **Fig 7-106**.) The oblique slit in the insulator lets you place the wire easily. I have however seen the wire "jump" out of the hole during heavy wind. I therefore drill a small hole (2 mm) in the insulator ring, through which I insert a piece of wire, which now prevents this from happening.

You can also slide a 10-cm long piece of suitably sized plastic tube over the top of the bamboo or other stick. Fix it with some electric tape. Make a slit in the tube with a hack saw, and just drop the wire into the slit. The same technique can be used when using slit heavy wall PVC pipe fitted over a rebar driven in the ground. (See **Fig 7-107**.)

Fig 7-106 — A 2.4 meter long bamboo stick can be use to support the Beverage wire. ON4UN uses electric fence insulators on top of the bamboo sticks (see text).

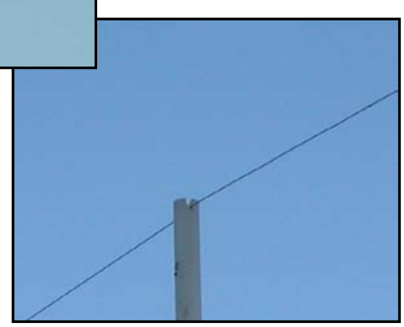


Fig 7-107 — A piece of good quality plastic tube slid over a rebar driven into the ground makes a good and strong support. A slit at the top captures the wire and lets it slide freely (seen at K9DX and W8JI).

Both systems allow the wire to slip freely in the support.

No doubt that with some imagination one can come up with many different solutions for supporting the Beverage wires. One thing is important: never anchor or wrap the Beverage wire around insulators, except at the ends. Always allow the wire to "float" through the insulators

I have a big box with old-fashioned egg-type ceramic insulators, with two holes. These are excellent for end-type insulators. If there is a lot of tension on the wire (in a permanent installation using copper-clad or bronze wire) be careful selecting plastic insulators. Some very thin plastic-compression insulators will actually cold-flow and allow the wire to pass through the insulation. Heavy-walled ceramic egg insulators are much more reliable.

2.11. Terrain and Layout Considerations

- How far should I keep the Beverage from my vertical transmit antenna?
- How close can I run two Beverages in parallel — with one shooting in the opposite direction?
- Can Beverages cross one another?
- How close can they cross?
- How straight must a Beverage wire be?
- What if my terrain slopes up and down?
- Can I run different Beverages from one feed point?

These are all very valid questions. And failing to understand what and why may turn a potentially wonderful antenna into a really lousy performer.

2.11.1. Proximity to Transmit Antennas

When located close to large resonant transmit antennas, receive Beverage antennas pick up noise and signals that are retransmitted by those antennas.

If you have a number of Beverages, and one or two are always noisier than the others, then there is a good chance those are the ones closest to your transmit antenna. Another indication is that a Beverage antenna is not much quieter than your vertical. It should be.

This is caused by mutual coupling between transmit and receive antennas. See **Fig 7-108** for one example. The degree of mutual coupling is one of the factors that determines how much current flows in the receive antenna due to current flowing in the transmit antenna. This mutual coupling under the right situations can produce changes in both the input impedance and pattern of the receive (and transmit) antenna. Even with a high degree of coupling, placing a high impedance in the transmit antenna limits the current flow and thus the interference caused by the transmit antenna. This is often called "detuning the transmit antenna" as this high impedance placed in the transmit antenna is often reactive, and thus a large change in resonant frequency is produced.

How do you make sure that noise is really coming from your transmit antenna? A simple test is to connect a variable impedance such as an antenna tuner to the transmit antenna's feed line and rotate it through all possible inductance and capacitance ranges while listening. By doing so, you will actually change the resonant frequency of the antenna. If you hear any difference in noise level in the receiving antenna while doing this, you can be sure the transmit antenna is coupling noise into the receive antenna.

If this is a problem, the transmit antenna can be detuned during receive, but first check to see whether the transmit antenna actually needs detuning or not. Some dipoles, inverted-Ls and other wire antennas have been effectively detuned by opening the TR relay in the rig or amp. This depends on the antenna impedance, feed line length, tuner, and anything else in line. It's worth checking before starting the detuning work.

If you have the appropriate test equipment, a more thorough approach is to measure the isolation between the two antennas by measuring power on one while feeding power to the other. If you detect any coupling you will need to decouple the transmit antenna, or else move your receive antenna further away. At one time I had a Beverage that passed about 12 meters from my transmit vertical. It was very noisy. Increasing the separation to approximately 40 meters cleared the problem.

The maximum amount of coupling we can live with is given by: 35 dB – receive antenna gain expressed in dBi. For example, if the receive antenna gain is –15 dBi (typical for a Beverage, see Fig 7-75), the maximum allowed coupling is 50 dB, which means that anything *less* than 50 dB is bad news. The decoupling should be at least 50 dB. In the above example signals radiated by transmit antenna are down 35 dB from the desired signals received on the receiving antenna.

The coupling can be measured by transmitting a known signal level on the transmit antenna and measuring the receive level with a band pass filter ahead of a power meter. If you have an antenna analyzer, this can often be used as the signal source (most analyzers have an output power in the 3 to 5 dBm range). Instead of using a BPF and power meter (which not everyone has available) use a receiver with the preamp in the off position. S9 on most well made receivers is ~72 dBm. For example if the receive signal is S9 + 18 dB, and the transmit level is 3 dBm, the coupling can be estimated as (3 – (–72 + 18) = 57 dB. This is a value which really does not require any "detuning" of the transmit antenna.

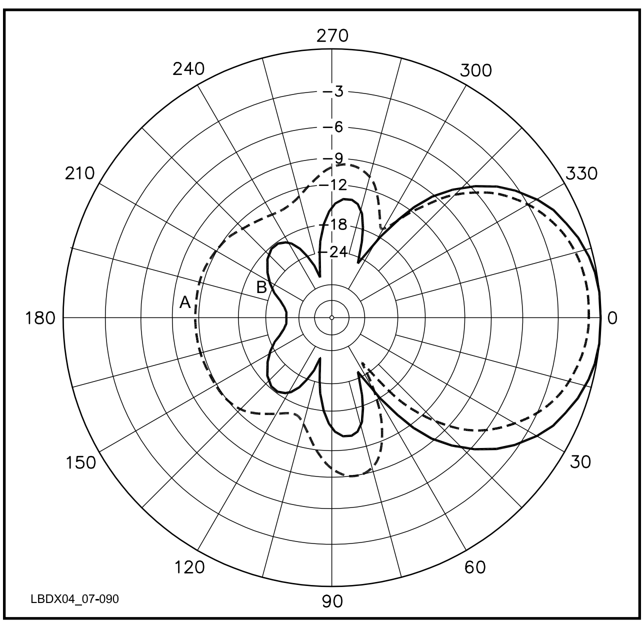


Fig 7-108 — Influence of the $\lambda/4$ 160 meter vertical on the azimuth pattern of a 268-meter long Beverage (dashed line). In this example the Beverage comes within 10 meters of the vertical.

Detuning the transmit antenna by any of the methods shown in **Fig 7-109** will improve the decoupling between the transmit antenna and the receive antenna by approximately 20 dB. This means that, if you measure a coupling of only 30 dB, detuning the transmit vertical will reduce the noise pickup on the receive antenna to an acceptable level.

If you have a series-fed λ/4 ground-mounted vertical, decoupling can be as simple as disconnecting the coax at the antenna, so that the vertical part of the antenna is now "floating." The antenna will now be resonant at twice the frequency and mutual coupling to the Beverage will be minimal. You can use relays at the base of the antenna (see Fig 7-109A), or you can do it at the end of the feed line. If the feed line has an electrical length equal to an odd number of quarter-waves, then the feed line should be short-circuited in the shack during reception (Fig 7-109C). If the feed line length is a multiple of half waves long, then the end of the feed line should be left open in the shack during transmit (Fig 7-109B).

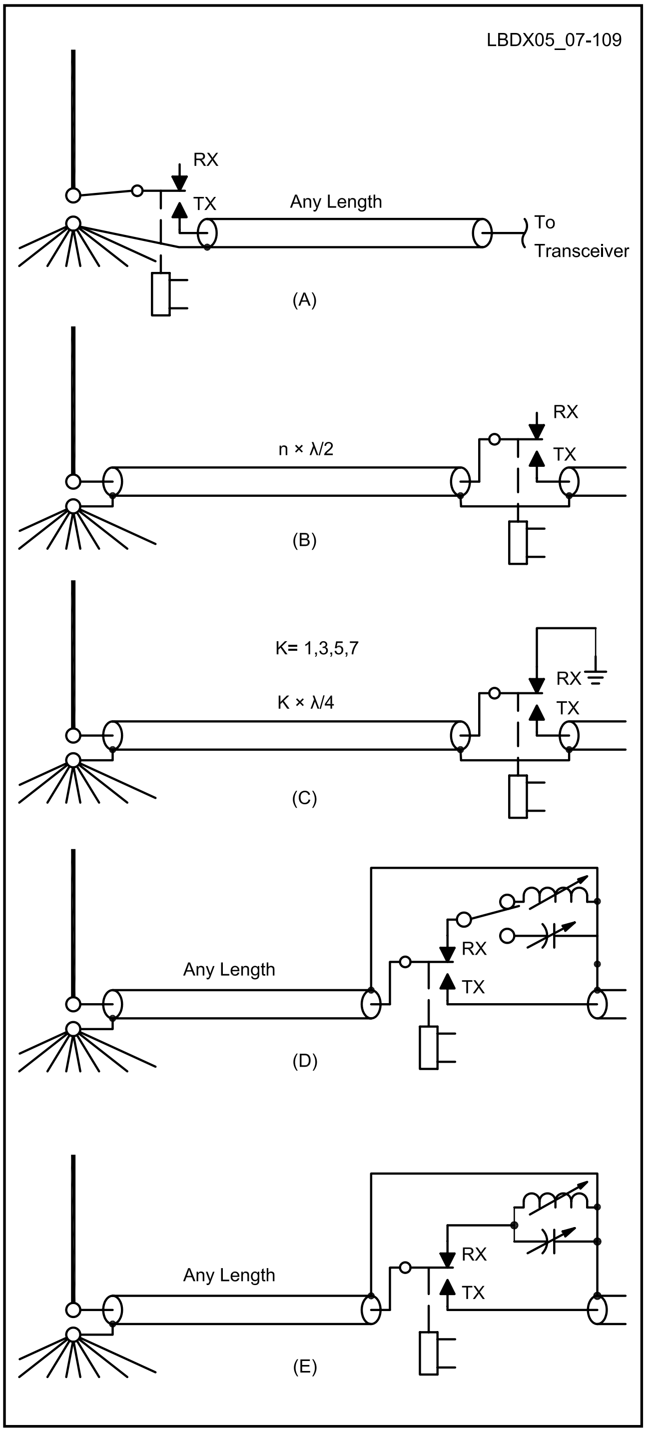


Fig 7-109 — Various methods for decoupling a series-fed vertical. These techniques all place a high impedance at the base of the vertical antenna.

If you don't know the feed line length, just terminate the feed line in a variable capacitor (such as a four-gang broadcast variable) or a variable inductor and tune until you hear the minimum noise level in the Beverage. This capacitor or inductor can then be switched across the end of the feed line during reception with a relay, as shown in Fig 7-109D. You can also use a series-resonant LC circuit with small values of C, or a parallel LC-circuit with large C and low L values, to do your testing. Replace these with a simple L or C once the right value has been determined.

If you have a shunt-fed vertical such as a loaded tower, just disconnecting the feed line may not help! Once you shunt feed a nonresonant structure, it becomes a resonant structure at the frequency to which it is tuned. To detune it you will need to turn a section of the tower into a parallel-resonant circuit, as explained in Section 3.11.

Robye, W1MK, has looked into the problem of elevated radials, and extensive modeling has proven that the coupling with the Beverage antenna is all due to the vertical element. Extensive modeling using *EZNEC* showed that the coupling between the Beverage (which is vertically polarized) and the radials is negligible. Robye modeled a vertical with two elevated radials spaced λ/4 from a 1 λ long Beverage which was pointed directly at the vertical on 160 meters. The isolation obtained was 36 dB. Then the vertical element was removed and the remaining two (in-line) radials were excited as a low (2 meter high) λ/2 dipole. This is not exactly the same as two radials in-line because the current distribution is different (that's why these two in-line radials do not radiate in the far field, as the radiation caused by each half is canceled in the far field). In the near field (radials close to the Beverage antenna) the radiation will be very similar though. The isolation was now in excess of 100 dB, which means that the interfering effect was now approximately zero.

Even if you don't notice an observable noise increase, the proximity of other antennas can hurt the directivity of your receive antenna. This may not always be easily detected just by listening to an antenna. A Beverage may still "out hear" a vertical, even if it is severely compromised by the proximity of another nearby antenna.

Conclusion on Nearby Transmitting Antennas

- Noise picked up by a large transmitting antenna can easily be coupled into a nearby receiving antenna.
- To resolve this problem you can increase the distance between the antennas (usually a minimum of λ/2 depending whether or not the receive antenna is firing into the transmit antenna or not).
- Shift the resonant frequency of the transmit antenna as far away as possible from the operating frequency during receive.

If your Beverage is close to the transmit antenna, and if you have no apparent noise problem, this can be for one of the following reasons:

- Your transmit antenna is right in the null direction of the Beverage.
- You have an extremely quiet location (maybe Heard Island?). If there is no noise being picked up on the vertical, then it would be very hard to discern any change in the Beverage performance.

Under any circumstance it is better to keep antennas separated as far as possible. This is also true for transmit antennas, of course. A situation where one or several wires are stuck randomly in the near field of an antenna ends up with results that are a matter of blind luck.

2.11.2. Parallel Beverages

Beverages are nonresonant and exhibit very little mutual coupling. You can run Beverages in parallel (shooting in opposite directions) with a separation as little as 50 cm! I have two Beverages (70° and 250°) that are separated approximately 70 cm and they work just fine (see **Fig 7-110**). Modeling told us that this is good enough for a decoupling of at least 30 dB, which is just fine.

2.11.3. Beverages Parallel to Fences, Power Lines or Telephone Lines

Here the story can be quite different. Fences, telephone lines and power lines can conduct and radiate a lot of noise that will easily be coupled into a Beverage that runs parallel to it in close proximity. (Power lines probably will.) Stay away from them.

Fig 7-110 — This shot taken on the author's Beverage field shows two parallel running Beverages (receiving from opposite directions) and a Beverage crossing. See text for details.

2.11.4. Crossing Beverages

Look at Fig 7-110 and you will see what the question means. At one point it looks like a busy highway intersection. Yes, Beverage antennas for different directions may cross each other if the wires are separated by 5 to 10 cm (2 to 4 inches). Make sure the crossing wires cannot touch one another (in heavy wind or when one or both wires have been stretched). A little wooden block is affixed to the bamboo stick using two small tie wraps and some plastic tape, and a fence ring insulator from Gallagher is mounted on the block.

2.11.5. How Straight the Wires?

Does a nice clear straight wire, without sags — which looks great — hear better? Deep sags in the wire are a little like sloping terminations (see Section 2.8) and in some way will upset the directivity of the antenna to some degree. It also causes the impedance of the Beverage not to be constant, and as a consequence, the SWR of the antenna to show ups and downs.

If the terrain is irregular, minor ups and downs in height or dips or valleys will not ruin your Beverage performance though. Follow the contour of gradual slopes. Go straight across ditches or narrow ravines without following the contour.

Although it probably is a good idea to keep the wire as straight as possible, it is the *overall* direction and length that is most important because each small area contributes a similar small portion to the overall directivity and signal reception.

If you are in a situation where your terrain goes up and down, a statement from Roger, VE3ZI will probably make you worry a little less about this issue: "I have 12 Beverages, each about 800 feet long at 30° intervals. Some of them are relatively flat, some of them go up and down by about 100 feet, some parts are across rock and some across swamp. I really can detect no difference in performance between any of them. I'm quite sure there is a difference, but my feeling is that it is so small as not to be relevant. Surely the real point is that Beverages work adequately even when gross compromises are made in their construction."

I have one long Beverage where the first 150 meters goes at 150°, the next 65 meters at 130° and the remaining 85 meters again at 150°. To me it sounds like a perfect 150° Beverage.

2.11.6. Multiple Beverages from One Hub

This issue was already covered in Section 2.7.2.9. The answer is: Never bring multiple antennas to one feed point, especially when they share one common ground. The issue is common-mode signal coupling. You can run a master feed line to a hub, where you have a switching box (equipped with common-mode chokes). From the switching box, run a number of short (typically 5 to 10 meter) feed lines to spots separated at least 5 meters from one another, and have *one* Beverage take off from each spot.

2.11.7. Choose the Placement of Receiving and Termination End Judiciously

Keep the near end as well as your far end (termination end) of your Beverage at least 5 (preferably 10) meters away from any grounded tower, such as a tower supporting your HF Yagis. The grounded tower picks up a lot of noise and reradiates it. Moving the near or far end away from such sources can bring the noise level received on the Beverage down by as much as 10 dB or more.

In a small street bordering the terrain that I use for my Beverages, the power lines are overhead bundle cables, supported by concrete poles, which — of course — include steel rebar. Walking down the street with a fox hunting 80 meter or 160 meter receiver, I noticed that every time I came near a pole, the noise went up significantly. So, treat such poles as towers and stay away from them with either the far end or the near end of a Beverage. The same is true for wooden poles with a grounding wire running along the pole.

2.12. BOG (Beverage On Ground) Antenna

The very first Beverage antenna in history was a BOG. The first antenna Harold Beverage used was simply several miles of wire laying on a sandy path on Long Island. Only later was the Beverage antenna refined to be an elevated wire with a resistive termination!

Today's BOG is simply a Beverage antenna that lies on the ground. You can actually bury it a few millimeters under the ground, and it will still work. In such a case you can call it a VLPB (Very Low Profile Beverage) or BUG (Beverage Under Ground). You could actually hide it under the lawn of your neighbor…

What is the difference between a BOG and a regular (elevated) Beverage?

- The output is much lower than from a comparable regular Beverage antenna (typically 10 to 15 dB down from a Beverage 2 meters high), which brings it in the neighborhood of the small loop receiving antennas. What counts is not signal strength but signal-to-noise ratio. We can always increase the strength (of both signal and noise) by using a good preamp, preferably at the antenna if the feed line to the shack is long and lossy (see Section 6).
- The velocity factor of the antenna is *much* lower than for an elevated Beverage (50 to 60% vs 95 to 98%). That is a very interesting property. It means that a 80-meter long BOG has as much directivity as a 150-meter elevated Beverage
- Its directivity is actually *better* than for the equivalent elevated Beverage, as there is no vertical down lead wire causing omnidirectional signal pick up.
- The impedance of a BOG is usually between 200 and 300 Ω. It is that "high" because of the losses of the ground. Over a perfect conductor it would be as low as 50 Ω. Using a 4:1 impedance transformer should give a reasonable match to a 50- or 75-Ω feed line.
- If you put a BOG on the ground, you or animals will trip over it. You can nail it down (using small hooks) or bury it just barely under the surface (which will of course further decrease the signal output).
- The BOG wire must be insulated. It is advisable to use Teflon or similar high quality insulating material that can withstand abrasion (critters have sharp teeth!). Teflon insulated wire is best, but expensive.

Are there special precautions to be taken when using BOGs?

In a BOG situation the pick up on the outer screen of the feed line coax is as important as the pick up by the BOG antenna wire. This means that extra common-mode decoupling is mandatory. In **Fig 7-111** we see that we have installed a common-mode choke in the feed line with a ground stake approximately 10 meters from the antenna feed point. The transformer with galvanically separated windings acts as a braid breaker which also helps in removing common mode signals (see Section 2.7.2.1). Do not connect a braid breaker and a common mode choke without a ground connection in between.

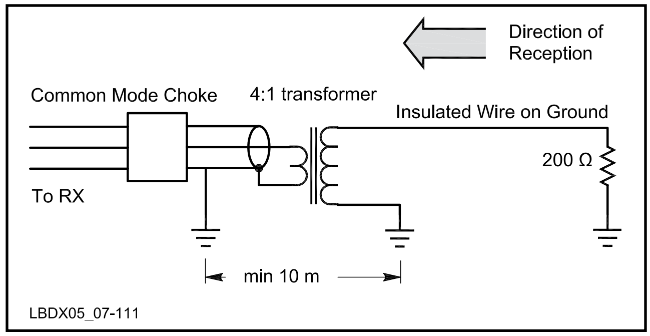


Fig 7-111 — The BOG is a simple Beverage with its antenna wire laying on the ground. See text for details.

The quality of the ground at the receiving end (at the transformer) must be very good in order to prevent further signal loss (attenuation) and common-mode signal ingress. You can, for example, use three ground rods each 1.5 meters long and spaced about 2 meters. If rocky terrain is a problem, use a dense ground screen. Do not use long radials or just a few radials. The counterpoise may not be resonant. It obviously also may *not* lie on top of the BOG wire! If it is OK to use the antenna on a single band you can use one or two in-line ¼ λ wires lying on the ground. Watch out — these wires also have a very low velocity factor. A quarter wave wire for 160 meters lying on the ground may only be 20 to 25 meters long!

Riki, 4X4NJ, is an avid user of BOGs and he wrote: "All of my five Beverages are BOG. I don't believe that this is a compromise antenna for terrain such as mine. My QTH is on a rocky mountaintop with very poor soil conductivity. Of course there is capacitance to ground. Some of my BOGs are unterminated, and others are terminated. Those that are terminated use a 200 Ω resistor to a quarter wave wire just lying on the ground. I use the ¼ wave wire termination because of low ground conductivity and the impossibility to drive in a ground rod — they always hit underground rocks. My experience has been that the unnterminated BOGs have excellent F/B. Receiving from the fed end direction is down at least 2 or 3 S-units. I can't properly explain why this is so, but I expect that it is because of the low impedance and losses. (The usual references predict only about a 3 dB difference.) My gut feeling is that in my particular QTH there is no significant advantage in terminating the BOG. Besides its simplicity, this is a rather stealthy antenna. From my house with a small yard, the BOGs go out into the surrounding public area. Two of them cross roads, so I have buried the wire a few inches beneath the (gravel) road. I've also buried a more extended length — about 100 feet — in order to hide the wire from view."

Guy, K2AV, has thoroughly studied BOGs and done quite a bit of modeling on these antennas. One of the issues he brought up is that "BOGs change significantly with the dampness of the earth which in turn affects velocity factor. If they seem off and the ground is dry, pre-contest soaking the ground (two or three feet) with a garden hose seems to brighten them up. Walking the BOG line while pouring out of a bucket also works." So now you know: you not only have to water the flowers in the garden, but also your BOGs.

A BOG may also be an attractive receiving antenna for DXpeditions. But they do not work in close proximity to saltwater nor over very good, conducting ground (such as a dry salt lake).

Bogs may also be interesting for someone who wants to put up a temporary directive receiving antenna (during summertime, for example, when the farmer is using his land) to be able to catch a new country.

2.13. Bidirectional (Unterminated) Beverage Antenna

When the Beverage antenna is not terminated, the directivity will be more or less bidirectional. In real life the unterminated Beverage is not a true bidirectional antenna. The SWR and the loss on the antenna make it have some F/B. F/B is related to standing waves on the antenna; that is, to the reflection making it back to the receiver.

Fig 7-112 shows the horizontal directivity pattern for a 1-λ long unterminated Beverage antenna. Notice the slight attenuation from the back direction because of the extra loss of the reflected wave in the wire.

A method of switching the Beverage from unidirectional to bidirectional is shown in **Fig 7-113.** A relay at the far end of the antenna wire is powered through the antenna wire via an RFC and the ground return of the feed transformer includes a blocking capacitor. A similar RFC and blocking capacitor circuit is used at the far end of the coax for feeding the dc in the feed line.

Note that the dc return path for feeding the relay is through the ground. Because of the resistive losses in this path, the relay must be powered from a high enough voltage source (up to 24 V dc for safety reasons). Because of the ground loss resistance, the relay requires a must-operate voltage that is substantially lower than 24 V (perhaps 12 or even 5 V).

Because we feed dc through the antenna wire, we need to galvanically ground the coaxial feed line near the antenna. I recommend installing a good quality common-mode choke or braid breaker on the feed line near the ground rod grounding the transmission line (see Sections 2.7.2.8, 2.7.2.9 and 2.7.2.10).

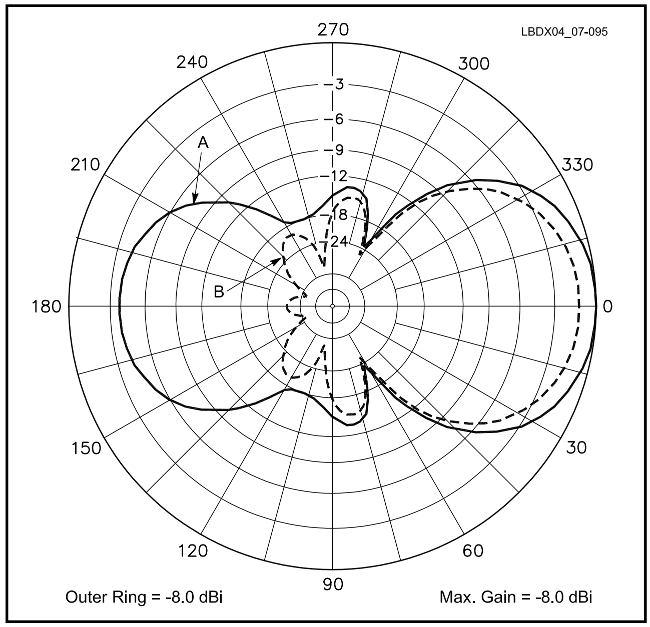


Fig 7-112 — Directivity pattern of a 1-λ long unterminated Beverage antenna. Pattern A is for the unterminated Beverage. Pattern B (dashed line) is the for same Beverage but properly terminated.

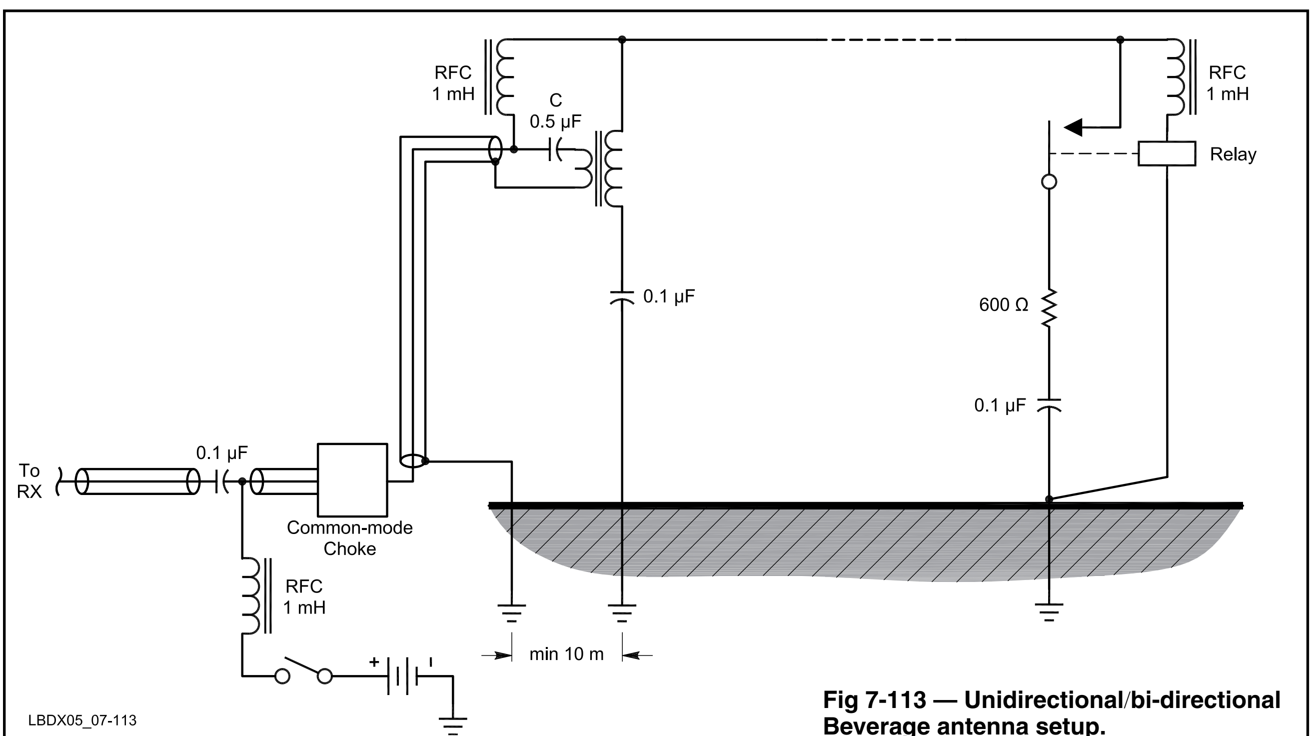


Fig 7-113 — Unidirectional/bi-directional Beverage antenna setup.

2.14. Two Directions From One Wire

Two directions can also be obtained from a single-wire Beverage by feeding coax to both ends of the antenna, with an appropriate matching transformer (**Fig 7-114**). One end is terminated in the shack with an appropriate impedance, while the other is used for reception.

The appropriate terminating impedance can be found as follows. First, determine the surge impedance of the Beverage using the method explained in Section 2.5.4.

Disconnect the antenna wire from T2 and connect an impedance bridge across the high-impedance secondary of the matching transformer. Adjust the terminating impedance at the end of the feed line (inside the shack) until it is the same as the Beverage surge impedance.

In many cases, the impedance will be nonreactive and a simple resistor in the range of 50 to 75 Ω will provide the proper termination. An alternative method is to use a small signal source in the back of the antenna (remember the offset angle, see Section 1.6) and simply adjust a 200-Ω potentiometer at the end of the second feed line for maximum front-to-back ratio. You switch directions by interchanging the coax going to the receiver with the one going to the terminating resistor.

2.15. Two-Wire Switchable-Direction Beverage Antennas

Two-wire reversible direction Beverages were used from the very first days. They are described in detail in the original paper titled "The Wave Antenna — A New Type of Highly Directive Antenna" written by Beverage, Rice and Kellogg for the journal of the American Institute of Electrical Engineers, Volume 42, 1923.

Two-wire switchable Beverage antennas are covered in great detail by V. Misek, W1WCR (Ref 1206). I strongly recommend that you read this book if you want to get serious with open-wire type Beverage antennas.

2.15.1. Switchable Direction Beverage Using Parallel Open Wire Line

Fig 7-115 shows the schematic representation of the two-wire reversible Beverage. Signals coming *from the right* in Fig 7-115 will induce equal in-phase voltages in both wires. Because of the close spacing of the wires, there is no space-diversity effect. The ends of the two wires are connected to a properly designed push-pull transformer (T2), which has the RF from the push-push (in-phase) signals on its center tap. If the transformer is properly balanced, the signals coming from the right will *not* be available at winding n2. The signals, available at the n1 center tap are fed to T1 which transforms these signals to the correct impedance for feeding them into the coaxial feed line going to J1. This means that signals coming from the right are available at J1.

Signals *arriving from the left* in Fig 7-115 arrive in phase at the primary (n1) of the reflection transformer T3. These signals are available at the center tap and are fed to the secondary (n2) of the transformer, which is now inductively coupled, via the n2:n1 windings to the push-pull winding (n1 of T3), which now feeds the balanced transmission line made up by the two parallel antenna wires. These are fed to the push-pull winding (n1) of T2. T2 now transforms these signals to its primary (n2) from where these are routed through the coax feed line to J2. If T2 is properly balanced, push-pull signals arriving on winding n1 will not be available at the center tap, and hence not at J1.

As we can see the two parallel wires act as an in-phase antenna wire and as out-of-phase transmission-line wires. Outputs from both directions are simultaneously available from outputs J1 and J2 of this system.

Preserving proper balance is critical for obtaining good performance: balance in the transformers T2 and T3 as well as proper balance of the parallel open-wire line vs the ground underneath. Therefore the open wire line should be parallel to the ground. If one conductor is closer to ground over the entire length of the line, it unbalances the system. The line should not be constructed in a vertical fashion, using single support poles with one wire under the other.

The closer the wire spacing of the line, the lower the impedance of the line, and the less critical installation-related unbalances become. Using twin-lead 300 or 450 Ω ladder line transmission line is obviously a good choice, since it intrinsically guarantees better balance. When using such lines it is appropriate to twist the line two or three turns per meter. If you use open-wire line, it is also a good idea to "flip" the wires

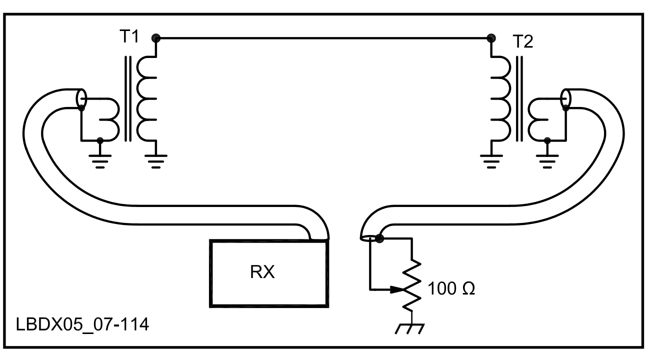


Fig 7-114 — Beverage with individual coaxial feed lines brought to both ends of the antenna.

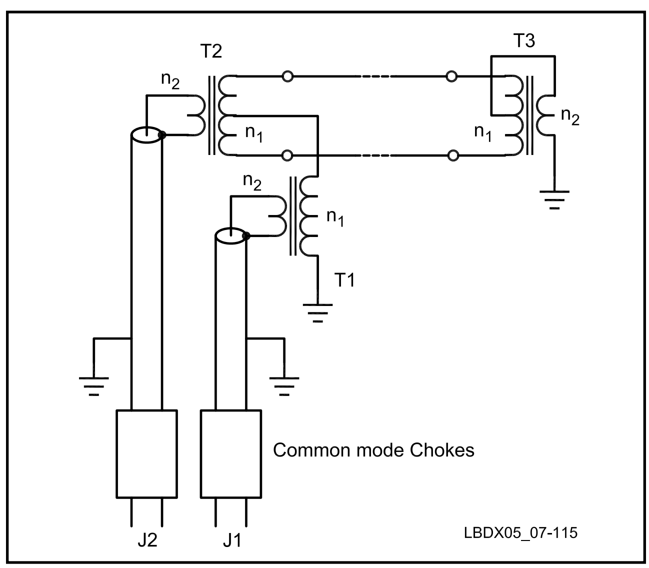


Fig 7-115 — Open-wire two-direction Beverage antenna. This design uses reflecting transformers to obtain both reflections and proper impedance matching (see text for details). The common-mode chokes and the associated grounds should be at least 5 meters from the antenna ground (to which T1 is connected).

every now and then along their run. The only disadvantage of such lines is cost and wind load.

Also, the higher you install the line above ground, the less unbalance any unequal coupling to ground creates. Most two wire Beverages made of open-wire (so-called 600 Ω) line use a wire spacing of approximately 30 cm.

When only one of the feed lines is connected to a receiver, make sure the other feed line is terminated. A variable potentiometer in the shack can be helpful to optimize the F/B under all circumstances.

Reflection transformer T2 in Fig 7-115 must be built to have the correct transformation ratio from the transmission line impedance (for example, 710 Ω balanced for the open-wire line) to the antenna surge impedance (approximately 332 Ω if the two-wire Beverage is 2 meters high).

The same precautions explained in Section 2.7.2.9 apply regarding the intrusion of common-mode currents from the feed lines into your antenna system. Therefore use separate ground rods for the antenna and for the feed lines (separated at least 5 meters), and install common-mode filters and grounds on the feed lines as shown in Fig 7-115.

2.15.2. Impedances Involved

The push-pull impedance of the open-wire line is given by:

$$Z = 276 \log \frac{2S}{d} \qquad \text{(Eq 7-2)}$$

where
S = spacing between conductors
d = diameter of wires (in the same units).

Table 7-31 shows the impedance obtained by combining a number of common wire diameters with wire spacings.

The impedance of the parallel wires over ground is given by:

$$Z = 69 \log \left[\frac{4h}{d} \sqrt{1 + \left(\frac{2h}{S} \right)^2} \right] \qquad \text{(Eq 7-3)}$$

where
S = spacing between wires
d = diameter of wires
h = height of wires above ground (all in the same units)

The surge impedance of the Beverage is higher than the value calculated with this formula (ranging from approximately 10% for very good ground to 30% for poor ground, see Section 2.5.2). **Table 7-32** gives the average values of the surge impedance of a two-wire Beverage made of different types of transmission lines. For the open wire line the Z_{surge} does not change a lot for different spacings going from 10 to 30 cm.

For such a two-wire Beverage system to work properly, great care must be taken to balance everything that needs to be balanced — the open-wire transmission line and the transformer with push-pull windings. With open-wire transmission lines, the balance to ground can be improved by periodically transposing the wires (say every λ/4 or less). Twinlead can be twisted periodically, for example 1 turn per meter. This becomes more critical as the line gets longer. This is also why you should not use a two-wire Beverage with an open-wire transmission line where the conductors are mounted one above the other, although here too frequent conductor transpositions can reduce the imbalance effects (together with a “good” height above ground).

Table 7-31
Open Wire Line Impedance

The impedance of an open-wire transmission line made of parallel conductors. This is the impedance of the two-wire transmission line used to transport the RF back from the far end to the feed point.

Wire Spacing	*For 1.3 mm Wire (16 AWG)*	*For 1.6 mm Wire (14 AWG)*	*For 2 mm Wire (12 AWG)*
10 cm	603 Ω	578 Ω	552 Ω
15 cm	652 Ω	627 Ω	601 Ω
20 cm	687 Ω	662 Ω	635 Ω
25 cm	713 Ω	688 Ω	661 Ω
30 cm	735 Ω	710 Ω	683 Ω

Table 7-32
Beverage Surge Impedance

This table shows the terminating impedance for a Beverage antenna made of two parallel wires (open-wire spaced 30 cm and three types of twinlead).

Height (meters)	*For 30-cm Open-Wire Line (Ω)*	*For 50-Ω Twinlead (Ω)*	*For 300-Ω Twinlead (Ω)*	*For 75-Ω Twinlead (Ω)*
0.5	252	341	365	393
1.0	295	383	407	435
2.0	332	424	449	476
3.0	357	449	473	500
4.0	375	466	490	518

2.15.3. Designing the Transformers for the 2-Wire Reversible Beverage Antenna

Let’s get practical. The transformers will use binocular cores (Fair-Rite # 2873000202, see Section 2.7.2.2). Remember that when we wind binocular cores, that one pass through one hole equals ½ turn. Let us assume we use 75-Ω feed lines.

Transformer T1

Refer to Fig 7-115. Assume we use a 600-Ω open wire line as an antenna (1.5 mm diameter wire, spaced 30 cm). Table 7-32 tells us that the surge impedance is approximately 330 Ω. T1 needs to transform the antenna surge impedance to 75 Ω. A 4:1 impedance transformer (2:1 turns ratio) will do that with an SWR of 1.1:1, which is very acceptable. In Section 2.7.2.3 we learned that we need 3 turns on the low Z side, so the turns for this transformer will be: n1, 6 turns (high impedance side) and n2, 3 turns (low impedance — 75 Ω — side).

Transformer T2

Transformer T2 must transform the impedance of the open wire transmission line (approximately 710 Ω, see Table 7-31) to 75 Ω. A 9:1 impedance ratio transformer comes very close (SWR = 1.05:1). N2 needs to be 3 turns (the 75 Ω side of the transformer), so n1 will be 3 times as much (9:1 impedance ratio means 3:1 turns ratio) or 9 turns. This winding requires a center tap (4.5 turns or 9 passes).

Transformer T3

In Section 2.15.1 we learned that reflection transformer T2 needs to transform the transmission line impedance (in this case 710 Ω) to the antenna surge impedance (332 Ω). This is an impedance ratio of 2.1:1 . The corresponding theoretical turns ratio is $(2.1:1)^{1/2} = 1.35:1$. A 4 to 3 turns ratio comes very close. As we are dealing with nominal impedances that are higher than the low-Z coaxial cable impedances, we shall multiply the last number by 2, which means that n1 will be 8 turns (with center tap) and n2 will be 6 turns (see Section 2.7.2.3).

An identical approach can be followed for a different mix of impedances (eg when using twinlead as antenna/transmission line).

Commercially made transformers (all three) for the two-wire Beverage antenna are available from K1FZ (**www.qsl.net/k1fz**) and from DX Engineering (**www.dxengineering.com**). **Fig 7-116** shows the circuitry at the "near-end" of the two-wire Beverage, where the feed line is connected. **Fig 7-117** shows the far-end reflecting transformer.

2.15.4. Testing the Transformers

The transformers T1, T2 and T3 can easily be tested for SWR by terminating the secondary with a terminating resistor and checking the impedance on the primary side with an antenna analyzer.

The balance of push-pull transformer T4 can be tested by connecting the two secondary outputs together and feeding a small amount of RF from any low-power source between this connection and the center tap. Use an appropriate attenuator at the output of your generator, because, if the transformer is well-balanced, you are actually short-circuiting the output of the generator! A perfectly balanced transformer should yield no output at the low-impedance winding from this configuration. The physical symmetry of the transformer may be adjusted (slightly adjusting turn spacings on the toroid core) while performing this test until the lowest possible signal output is achieved. The balance can be assessed by temporarily disconnecting one of the secondary leads and measuring the signal-strength difference on the receiver. Better than a 40-dB difference should be easily obtainable.

John, W1FV, describes a procedure for tweaking the reflection transformer: "Adjust the turns ratio until the SWR in the reverse direction at the receiver end is flattest. By flattest I mean that SWR exhibits the smallest variations up and down over a wide range of frequencies, and not necessarily the lowest SWR. With an antenna analyzer like the AEA or MFJ, sweep the frequency from 1.8 to as high as 10 MHz and look for the SWR to change the least over this entire range, not just inside the amateur bands. This procedure usually gets you very close to the optimum."

2.15.5. Coaxial Cable as Feed Line and as Antenna Wire

We know that a coaxial feed line can act as a transmission line, where everything happens inside the cable between the center conductor and the inside of the shield. But it can also act as a single fat conductor, when you look at the coaxial cable "from the outside."

Fig 7-118 shows a small coax cable used as a Beverage antenna wire. The coax cable also serves as the feed line for the far end of the array. Assume we use 50-Ω coax such as RG-58 for the antenna (making a "fat" antenna wire), and feed the system using 75-Ω coax. With RG-58 at 2 meters in height the surge impedance of this fat antenna wire is approximately 450 Ω.

Reflection transformer T3 is an easy one: 450 Ω to 50 Ω, which is (3:1 turns ratio). On a Fair-Rite # 2873000202 core, we shall use 3 turns for n2 and 9 turns for n1.

Transformer T1 needs to transform 450 Ω to 75 Ω, which represents a 6:1 impedance ratio or 2.45:1 turns ratio. Using the same binocular core we can wind 3 turns as n2 and 7 turns or 7.5 turns (15 passes) as n1. Try both and see which gives the best SWR.

Fig 7-116 — This near-end receiving board, designed by W8JI, is used in the DX Engineering reversible antenna system model RBS-1P. It contains two transformers, a direction-switching relay and the termination resistor for the unused direction. In this configuration, only one direction can be used at a time.

Fig 7-117 — The far end reflecting transformer with its three terminals: two going to the two-wire Beverage and one going to ground (part of the RBS-1P system).

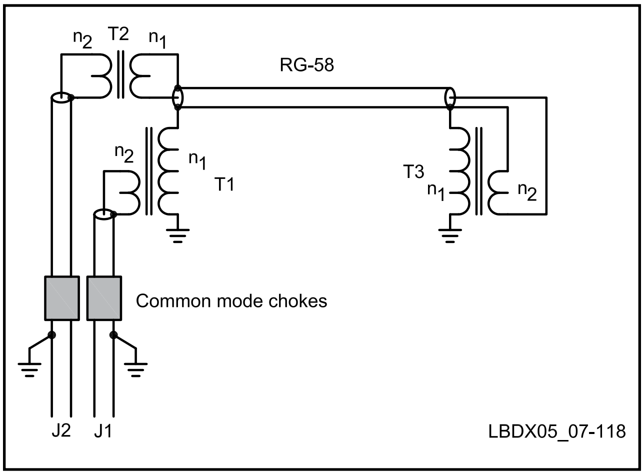


Fig 7-118 — Two direction reversible Beverage using coaxial cable as antenna wire. Note that all ground connections shown in the schematic need to be made with separate ground rods that are spaced at least 5 to 10 meters.

T2 needs to transform 50 Ω (the RG-58 coax impedance) to 75 Ω (the feed line to the shack). Use 4 turns for n2 and 3 turns for n1 (4:3 turns ratio = 1.8:1 impedance ratio).

If we had used 75-Ω coax such as RG-59 for the antenna coax, we still would have required a 1:1 impedance ratio transformer in order to galvanically separate the fat antenna wire from the feed line coax (isolation transformer)! In that case T3 would have required a different turns ratio: 450:75 = 6:1 impedance ratio, which requires a 2.45:1 turns ratio, which means 3 turns for n2 and 7 turns for n1.

2.15.6. Snake Antenna

The snake antenna is the reverse fed version of the BOG (Section 2.12). As with the BOG, the snake works because of common-mode excitation on the outside of a coaxial feed line. The entire shield picks up signal from the outside.

Seen from the outside (by the radio wave we want to capture) the shield of the coax acts as a single fat wire antenna. As far as receiving signals, everything goes on *outside* the coax. The velocity factor of the coax is not a consideration with the coax acting as a fat wire antenna, contrary to what some think. The velocity factor of the coax relates to the coax acting as a transmission line, and then everything happens *inside* the coax.

So, let's stay outside of the coaxial structure for a minute, and look at **Fig 7-119**. The fat wire, lying on the ground has a surge impedance of 200 to 300 Ω, depending on the quality of the ground. The better the quality (conductivity), the lower the impedance, and also the lower the signal output. A wave coming from the left induces EMF on the fat wire, and at the end of the thick wire the EMF voltage is fed into a transformer, which we call the reflection transformer. The reflection transformer (200 to 300 Ω to 50 or 75 Ω) is a classic 4:1 impedance transformer as described in Section 1.25.2 and shown in Fig 7-30.

The principle of operation is exactly the same as explained in Section 2.15.1. for the two-wire "above the ground" reversible Beverage.

At the other end of the fat wire we have the termination resistance (200 to 300 Ω) connected to a good ground system. The required quality of this ground connection is described in Section 2.11. In the case of a snake a good low resistance ground is even more important because of the low signal output and the increased chances of common-mode ingress.

From there on, looking to the left, the system consists of two common-mode chokes and a ground connection between the two; the ground connection must be at least 10 meters from the termination ground rod.

You can make a snake like the two-wire antenna in Fig 7-118, where you have two directions available with the flip of a switch in your shack. The difference between this and the "high" Beverage antenna as described is that the antenna impedance is typically around 200 to 300 Ω instead of 450 to 600 Ω; the velocity factor is much lower (typically 0.55-0.6 instead of approximately 0.95); and its output is down about 10 to 15 dB. The lower output should not be a problem, as you can, if necessary, catch up with a good preamp.

2.16. Arrays of Beverages

Sections 1.6 through 1.12 of this chapter explain the principles of broadside and end-fire arrays. Beverages make ideal elements for an antenna array. In vertical arrays as well as Beverages, broadband-phasing systems can easily be implemented because the feed-point impedance is stabilized through lossy mechanisms such as series resistors in vertical arrays or "natural" loss mechanisms in Beverages. If the losses are made large enough to swamp out or dilute mutual coupling and resonance effects, the antenna feed-point impedance will remain stable and predictable even when elements are end-fire phased with a unidirectional pattern and close spacing.

Section 1.16 explained that in receiving arrays using short

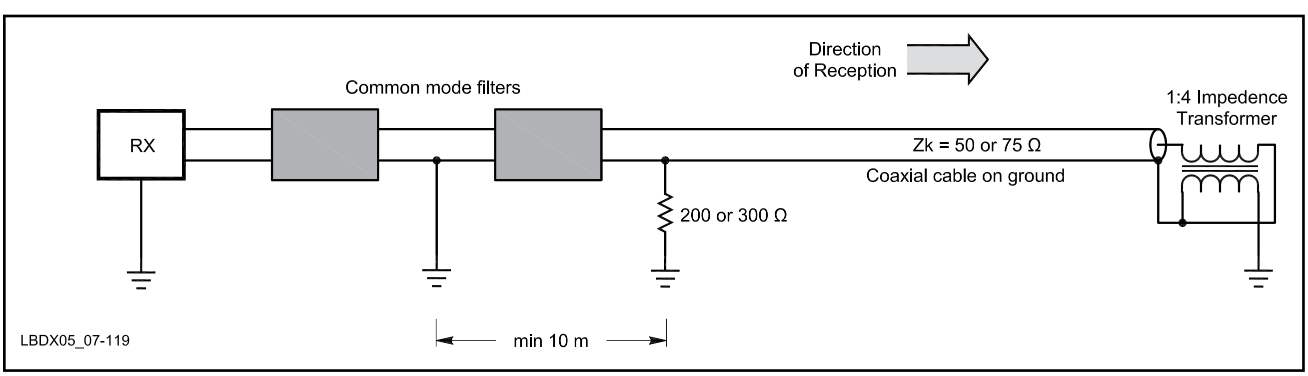


Fig 7-119 — The snake antenna is the back-firing version of the BOG (Beverage On Ground). See text for details.

verticals as elements, you should use heavy resistor loading (75 Ω) in combination with the very low radiation resistance (1 to 2 Ω) to achieve this goal. You can use lossy elements because all you need is to overcome the noise, which on the low bands is noise received by the antenna rather than internal receiver noise.

The radiation resistance of a Beverage is very low (1 to 2 Ω). The remainder of the Beverage surge impedance (300 to 500 Ω) is made up of dissipative losses in the ground under the antenna plus the termination resistance. With a single-rod ground termination these losses may be as high as 25% of the total system loss. This means that the overall loss resistance is typically 200 to 300 times higher than the Beverage's radiation resistance, making for a very stable feed-point impedance and very low mutual coupling in the feed-point impedance even with other closely spaced Beverages nearby.

Beverages also have the ideal characteristic of providing a relatively constant feed-point impedance over wide frequency range. This makes arrays of Beverages ideal candidates for wide-bandwidth phasing systems, eliminating complex switching systems that are required when large verticals are used as array elements, since the verticals must be tuned at each operating frequency.

Can you really improve Beverages by phasing them? First of all you do not need to phase them to get more "gain." Gain or sensitivity is *not* an issue on the low bands. Signal-to-noise ratio, and hence directivity is the only real issue. Theoretically you can achieve a narrow forward lobe (and hence a very high DMF) using a very long Beverage. To achieve the same 55° frontal lobe beamwidth as from an Eight Circle antenna (Section 1.30), you would need a 3-λ long Beverage (about 550 meters on 160 meters), which poses two problems:

- Not many have that much room for such long Beverages.
- You run into space-diversity problems with this long an antenna, and you will never reach the results predicted by modeling.

The only way to achieve a similarly narrow forward beam and hence a very high RDF is to use a broadside array of Beverages, exactly the same as what we did with short verticals in Section 1.11. You can obtain excellent results with shorter Beverages (1 to 2 λ long) but these require a broadside spacing of a little over λ/2. There is no free lunch!

Improving the front-to-rear ratio (remember the F/R ratio is *not* the same as F/B), in other words improving the DMF (see Section 1.8), is also quite simple. Just put up two Beverages in an *end-fire* arrangement, exactly the same way I discussed using short verticals (Section 1.6). End-fire Beverage phasing can also help reduce or eliminate signals picked up from the vertical down leads (Section 2.8).

For calculating Beverage arrays I use a model where each Beverage is set up as an inverted-V shape, with the highest point half-way (at 2 meters high), and sloped down to the ground at both ends. This is because with the T-shaped λ/4 terminations (see Fig 7-66B) in an array the termination wires would be too close together and mess up the modeling results.

2.16.1. Beverages and Mutual Coupling

Beverage antennas are very lossy and tightly coupled into the nearby earth. As a result the mutual coupling between even very closely spaced Beverages is barely visible as impedance changes. The fact that Beverages are nonresonant antennas emphasizes this characteristic. Elements at or near resonance exhibit the highest degree of mutual coupling (that's why Yagi antennas work!).

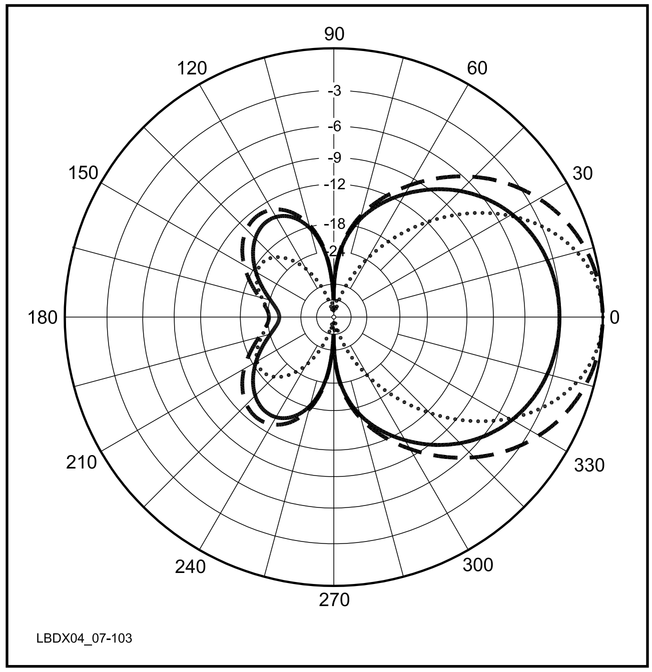


Fig 7-120 — A single 160-meter long Beverage (solid line) and for two such Beverages in phase, side-by-side, spacing 40 meters (dashed line) and 90-meter spacing (slightly over λ/2, dotted line). Actual gain is irrelevant for receiving.

This can experimentally be confirmed this by putting some RF into one Beverage and measuring the signal injected into an adjacent Beverage. Parallel Beverages spaced from each other by the same distance as their height above ground have about 30 dB of isolation.

I confirmed this by modeling also. I modeled two Beverages fed in-phase and monitored the feed impedance while changing the separation between the antennas.

2.16.2. Broadside Array of Beverages (Beverages in Phase)

In Section 1.11 I discussed how the broadside configuration (for example, spacing two antennas λ/2 apart) can give a substantial narrowing of the forward lobe. This configuration does not, however, improve the F/B since signals arrive from the back and from the front in-phase. Narrowing the forward lobe is especially beneficial when you cannot use long Beverages to achieve this result. If you only have room for a couple of short Beverages, but you have the room to space two of them λ/2 apart, you're in business for a super antenna. Unfortunately, if you are limited in space this is usually true in all directions, so broadside configurations are not a cure-all for everyone!

Fig 7-120 shows the radiation patterns at a 20° elevation angle for a single 160-meter long Beverage (1 λ) and broadside pairs spaced 40 and 90 meters at 1.83 MHz. **Table 7-33** lists the directivity data. As expected, the RDF improves considerably since we are narrowing the forward lobe, while the DMF changes less, except for the case of 90-meter spacing. Here

Table 7-33

Two Broadside 160-Meter Long Beverages at 1.83 MHz

	Single Antenna	*Two, Spaced 40 meters*	*Two, Spaced 90 meters*
DMF	19.0	19.5	21.3
RDF	10.2	10.6	11.9
–3-dB Angle	78°	69°	48°
Two Broadside 80-Meter Long Beverages at 1.83 MHz			
DMF (dB)	11.1	–	14.4
RDF (dB)	7.3	–	9.6
–3-dB Angle	90°	–	48°

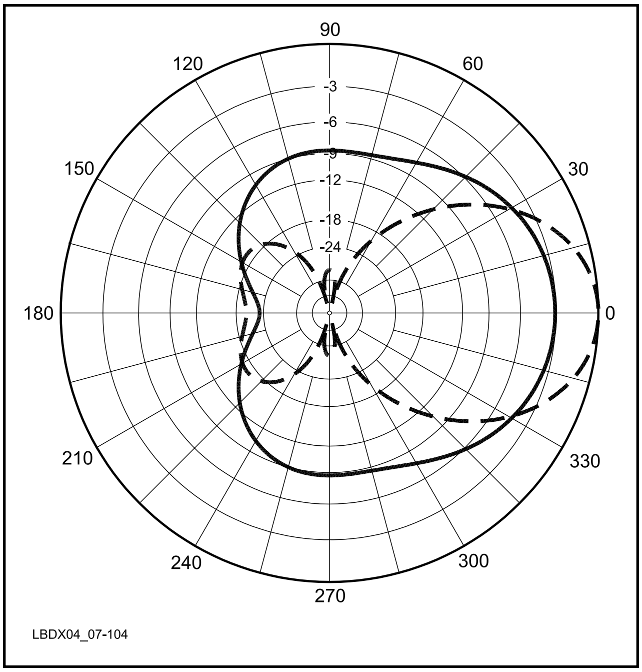


Fig 7-121 — A single 80-meter long Beverage (solid line) and two such Beverages in a broadside configuration spaced 90 meters (dashed line) on 1.83 MHz. Broadside makes the forward lobe narrow but does nothing for the F/B or the elevation angle.

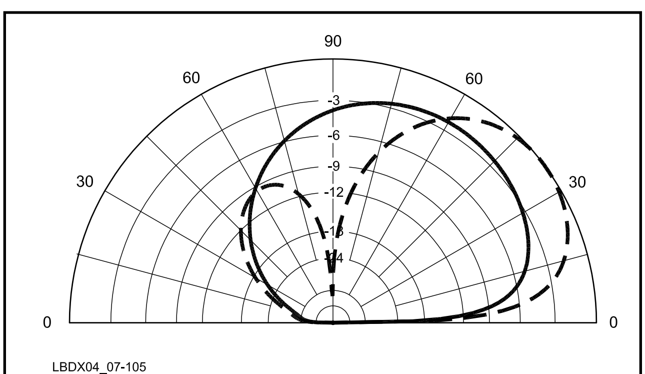


Fig 7-122 — Elevation pattern (in the main direction) for an 80-meter long broadside pair (solid line) and for a 160-meter long broadside pair (dashed line).

the back lobe gets much narrower, although the geometric F/B remains identical.

Note the improvement in directivity in Table 7-36 (later in this chapter) — not in the back but at right angles and in the front, as witnessed by a narrower forward lobe. By definition, two poor λ/2 Beverages make a relatively good pair in a broadside combination. The elevation angle is not lowered, however, and a pair of short Beverages will still exhibit a much higher takeoff angle (48° for 80-meter long Beverages) than a single long Beverage (33° for 1-λ long and 26° for 2-λ long). See **Fig 7-121**.

Arrays of two broadside Beverages are not very common because of the wide spacing (λ/2 or wider) required for good performance. That much space is not often found in the suburban environment where most hams live.

If you've got the space, however, a broadside configuration with a ⅝-λ spacing is a wonderful complement to an end-fire Beverage array (Section 2.16.3). This makes an end-fire/broadside combination a real winner, just as effective as an Eight Circle receiving vertical array (Section 1.30). See **Fig 7-122**.

Large broadside spacings can give a narrow forward lobe. Modeling makes it all seem very easy: Just increase the separation and keep feeding both in-phase. This way you can reduce the forward lobe's –3 dB angle from 48° for a lateral spacing of 90 meters to 35° for a spacing of 135 meters and 27° for a spacing of 180 meters. But that's just modeling. In real life there are two major problems associated with the use of wide-spaced arrays:

- The forward lobe becomes excessively narrow (30° and less), and you either have to be lucky that you're shooting in the right direction or you need many such arrays to cover all azimuths.
- Space-diversity problems: Very often signals arrive at two very wide-spaced elements with different and constantly changing phases.

W8JI's experience is that up to 0.67 or 0.7 λ spacing the system is reliable and predictable, but over 0.8 λ the systems fall apart quickly. He wrote: "Large spacings improve S/N under a wide range of conditions, but are more critical and often require constant re-adjustment of phase. I have to change phase between two Beverages spaced 500 feet as much as 130° during the course of one opening at times! With wide spaced arrays, a phasing control is a *must* if you are going to use them night after night. Some nights phasing them will be useless or actually hurt because phase will be shifting so fast it will be impossible to add the desired signals. On calm stable propagation nights large arrays work exceptionally well, but on nights when we are having geomagnetic storms they are mostly useless!"

2.16.2.1. Feeding an In-Phase (Broadside) Array

While we can merely parallel connect the equal-length coaxes running to the two Beverages and use an impedance transformer to match the parallel impedance to the impedance of your feed line going to the receiver, it is better to use a *zero degree hybrid combiner*, followed by a 2:1 impedance transformer, as shown in Fig 7-26.

Make sure there is no phase inversion in the two Beverage feed transformers. The multiple ground symbols shown

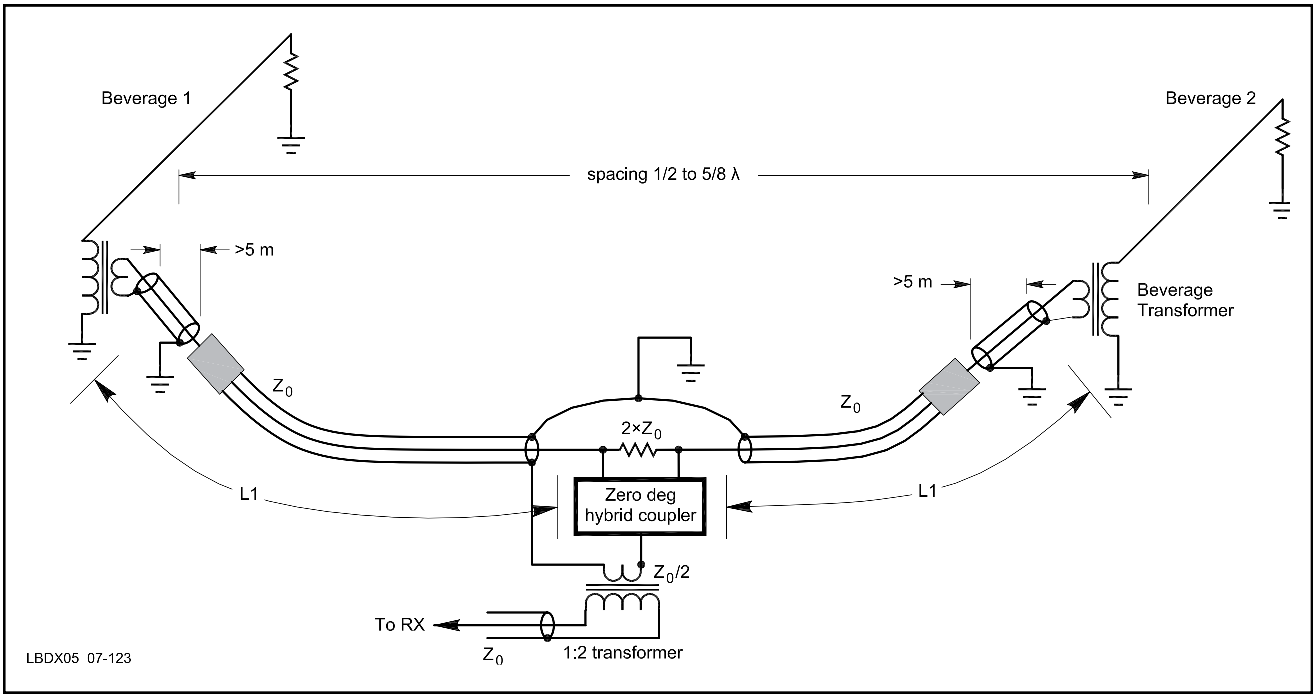


Fig 7-123 — Complete feed system for a broadside Beverage array. The schematic includes all the necessary steps to be taken to prevent common-mode signals from being coupled into the feed line.

in **Fig 7-123** all indicate separate ground rods, located at least 5 meters from one another. It makes no sense to implement such a top notch receiving array without implementing all possible measures to prevent common-mode signal ingress. Common-mode chokes, braid breakers (see Fig 7-99 and Fig 7-100) and all other measures are described in detail in Section 2.7.2.9.

DX Engineering **(www.dxengineering.com)** sells a suitable commercial splitter/combiner, including the impedance-matching transformer at its output (model RSC-2).

2.16.3. The End-Fire Beverage Array

If you are after improved directivity in the rear of the antenna (improving the DMF), you can put up two Beverages fed as an end-fire phased array, just like what I described for short vertical elements in Section 1.6. The Beverages you wish to phase together need to be identical (same height and length).

Modeling shows that as you reduce the spacing (called *stagger* or *offset* for end-fire phased Beverages) you can maintain the same directivity, but your output (gain) drops. Gain is, of course, not a major issue with low-band receiving antennas. And as opposed to what happens with narrow-spaced verticals in an array, you do not have increased mutual coupling when the stagger distance is decreased. See **Fig 7-124**.

The stagger distance and the phasing angle (ψ) are related to the null angle in exactly the same way as explained in Section 1.6. We can use the data in Table 7-1 for Beverages also. Assume you have two Beverages and you can stagger them 30 meters (66°) on 160 meters. You want to optimize the suppression for an offset null angle of, say, 30°. The phasing angle will need to be 125°. See **Table 7-34**.

With end-fire phased Beverages the term *spacing* is normally used to indicate the lateral (side-by-side) spacing between two Beverages in an array. It is obvious that unlike a vertical array, we need to slightly side-position the Beverages — otherwise they would be on top of one another.

Table 7-34
End-Fire Pair of 160-meter Long Beverages (30-m Stagger, 5-m Spacing, 1.83 MHz) Performance Vs Phasing

	1 Wire, Well Terminated	*Two ϕ = 125°*	*Two ϕ = 140°*
Gain (Ave Gnd) in dBi	–10.1	–8.4	–9.2
DMF (dB)	17.4	27.5	30.1
RDF (dB)	10.1	11.3	11.6

2.16.3.1. How Much Spacing?

In Section 2.16.1, I pointed out that Beverages hardly couple to one another. I have used spacings of 5 meters very successfully, and this can even be further reduced to 2 meters without any ill effects. Changing the spacing from 10 meters to 1 meter reduces the output by only 0.5 dB.

An interesting aspect of using two Beverages in an end-fire configuration is that the F/B of the individual elements is not so important to achieve a very good directivity (DMF). After all, we get good directivity with two vertical elements, and they do not have any F/B by themselves! In an end-fire Beverage array it is important that the two elements have a similar F/B, as you are going to subtract the signals from both elements. You only get zero if you subtract two identical numbers. It does not matter what the numbers are though. Two Beverages with a 10-dB F/B can produce 30 to 40-dB nulls in an array, just as much as a pair where the individual elements already show 25 dB F/B.

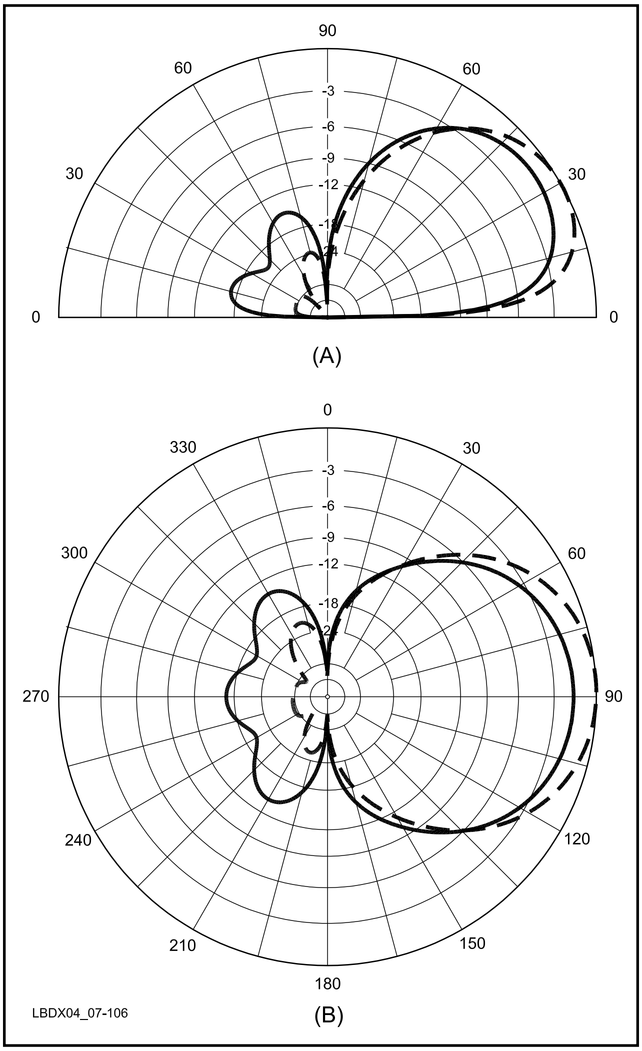


Fig 7-124 — At A, elevation patterns for a "poorly terminated" 160-meter long Beverage (solid line) at 1.83 MHz compared to an identical (poorly terminated) Beverage (dashed line), in an end-fire configuration (spacing: 5 meters, stagger: 30 meters, ψ = 125°, R = 300 Ω). At B, comparison of azimuth patterns for the same antennas. Note how both elevation and azimuth patterns have been cleaned up.

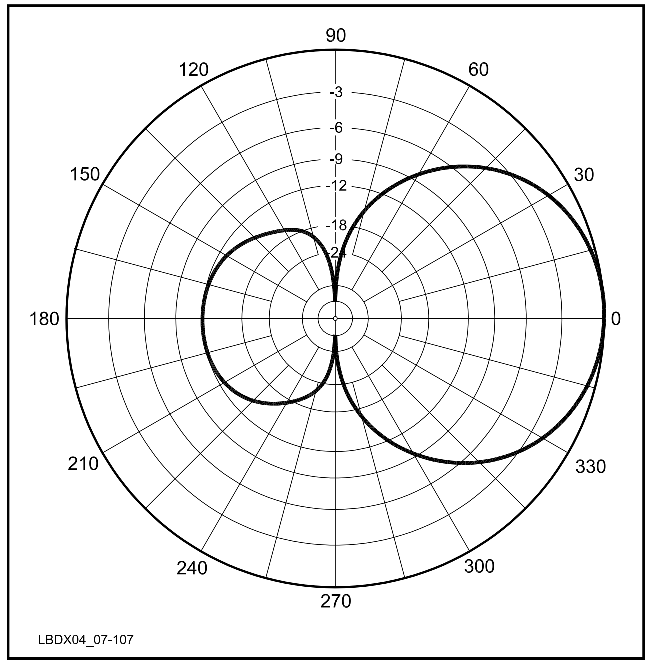


Fig 7-125 — Azimuth pattern for the end-fire Beverage array in Fig 7-124, with one Beverage correctly terminated (425 Ω) and the other one incorrectly terminated with 850 Ω.

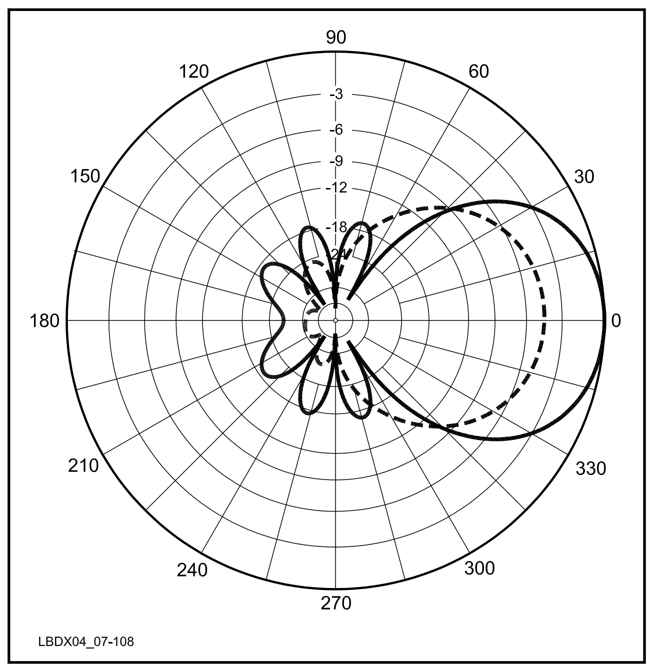


Fig 7-126 — Azimuth pattern on 1.83 MHz for a single 320-meter Beverage (solid line); azimuth pattern for end-fire pair of 160-meter long Beverages, half the length (dashed line).

Fig 7-125 shows the pattern for the same array as Fig 7-124, but where one of the Beverages is purposely terminated improperly with 850 Ω and the other one terminated properly with 425 Ω. The DMF of this array is 16.9 dB versus 30.1 dB for the case with identical termination resistances (425 Ω). Obviously, you should take care to use similar and stable ground systems on both sides and to use exactly the same termination resistances.

2.16.3.2. Two Short Ones or One Long One?

Let us compare two end-fire phased 160-meter long Beverages with a single 320-meter long one. (I wonder how many people can actually put up a 320-meter long antenna!) See **Fig 7-126**. The 320-meter long Beverage has a narrower forward lobe of 60° at the half-power points, vs 70° for the end-fire pair. This results in a better RDF for the long Beverage of 13.2 dB vs 11.6 dB, while the DMF is clearly better on the end-fire pair, at 30.1 dB vs 20.8 dB.

In plain language, the single long Beverage has a narrower forward lobe and will be able to discriminate better against noise arriving from all directions, as well as QRM and noise coming from a direction close to the wanted direction. The pair of shorter Beverages will give a much better discrimination

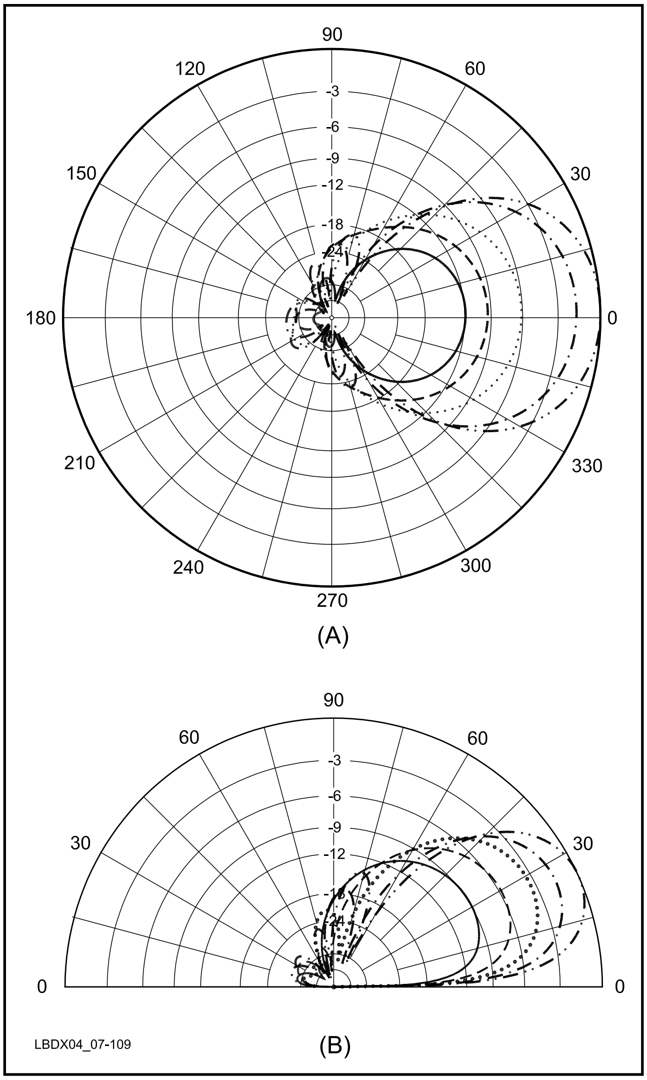


Fig 7-127 — End-fire phased Beverages of various lengths on 1.83 MHz (5 meters spacing, 30 meters stagger, ψ =140°): solid line = 100 meters; dashed line = 150 meters; 200 meters = dotted line; 250 meters = dashed-dotted line; 300 meters = dotted-dotted-dashed.

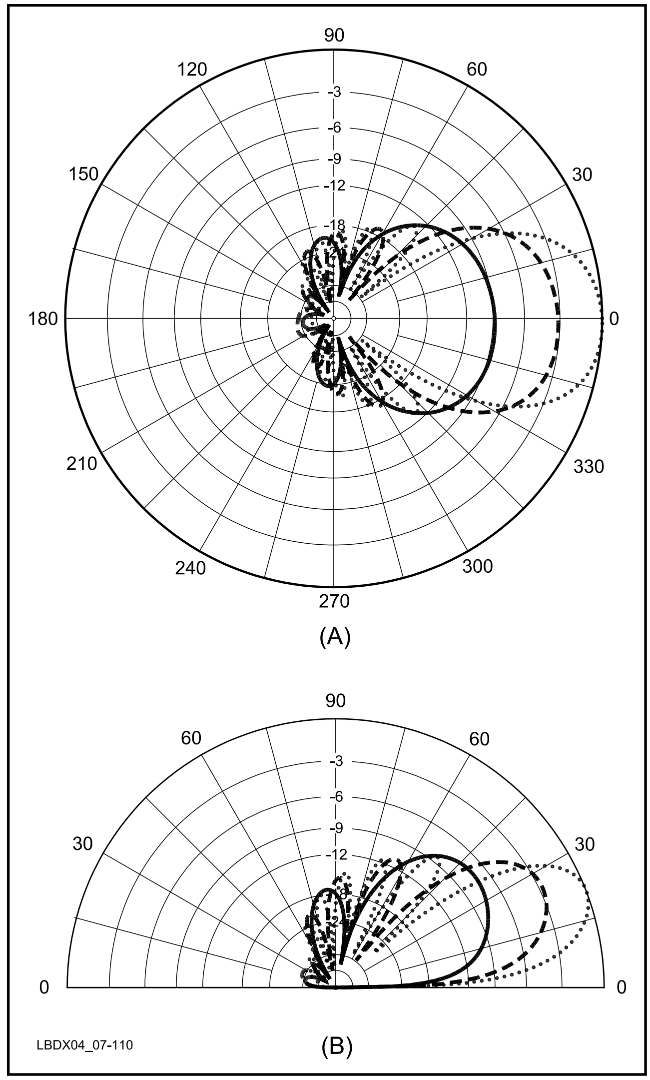


Fig 7-128 — End-fire phased Beverages of various lengths on 80 meters (5 meters spacing, 30 meter stagger, ψ = 70°): solid line = 100 meters; dashed line = 200 meters; dotted line = 300 meters.

against all QRM and noise coming off the back of the antennas. How narrow a forward lobe you can live with depends only on how many Beverages you can put up to cover all wanted directions.

2.16.3.3. How Much Stagger Distance?

If two Beverages (and the same holds true for verticals) are staggered λ/2 ("spaced" is the term for verticals), the array will need no extra phasing line to obtain a zero off the back at a zero wave angle, since signals at the end of equal-length feed lines are already out-of-phase.

To use a Beverage over a wide frequency range, you must examine the *stagger distances* you can use so that the system works correctly on all bands. The rule is simple: The stagger distance should be not larger than λ/2 at the highest frequency on which the antenna will be used. For 7, 3.5 and 1.8 MHz, the stagger distance should not be greater than 20 meters.

If you want to use the Beverages on 80 and 160 meters only, I would advocate using a stagger of 30 meters, which is a good compromise between size and gain. As with the vertical arrays, the gain drops as you reduce the stagger (spacing) distance.

Fig 7-127 shows azimuth and elevation patterns for end-fire Beverages of various lengths for 160 meters. In Section 1.6, I described how you can move the point of maximum rejection in both the horizontal and vertical planes by changing the phase angle. If you consider the *average* rejection in the back (the DMF), changing the phase angle in a limited range will have little influence.

Fig 7-128 shows the 80-meter patterns in the back of two 150-meter long Beverages in an end-fire array with 30 meters stagger and 5 meters spacing for various lengths. From the shack you could change the length of the phasing line using relays or a coaxial switch, but I have found this to be of no practical value, *unless* you would want to put a null directly in the direction where you have a noise source on ground wave.

Fig 7-129 shows the 3D pattern for the longest array in Fig 7-128. See **Table 7-35**.

Table 7-35
Directivity of End-Fire Beverages
(Stagger = 30 meters, ϕ=140°) on 1.83 MHz

	100 m	*150 m*	*200 m*	*250 m*	*300 m*
DMF (dB)	24.6	30.6	29.0	33.0	33.8
RDF (dB)	10.3	11.4	12.4	13.1	13.9

Directivity of End-Fire Beverages
(Stagger = 30 meters, ϕ = 70°) on 3.65 MHz

	100 m	*150 m*	*200 m*	*250 m*	*300 m*
DMF (dB)	22.3	27.4	27.0	29.6	32.1
RDF (dB)	10.5	12.6	13.9	14.0	15.9

Table 7-36
160-m Long Beverages in End-Fire/Broadside Combination (ϕ = 120°)

	Single End-Fire Cell	*Cells Spaced 45 m*	*Cells Spaced 90 m*	*Cells Spaced 135 m*
DMF (dB)	31.8	32	34	34.7
RDF (dB)	11.5	11.9	13.0	14.1
3-dB angle	70°	61°	46°	34°

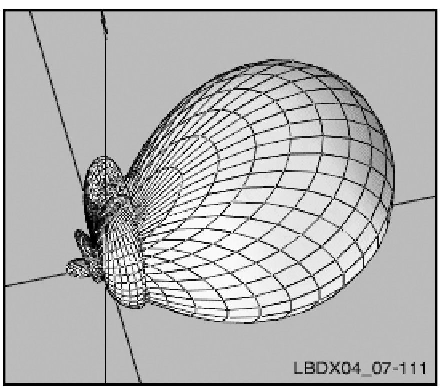


Fig 7-129 — 3D pattern of a 250-meter end-fire phased pair of Beverages on 160 meters. Note the extreme rejection of radiation in the back of the array. The DMF of this array is 34.2 dB (RDF = 12.2 dB). (Pattern by *Antenna Model*.)

John, K9DX, rightly points out that the success of two end-fire Beverages may to a great extent come from phasing out omnidirectional signals picked up by the vertical down leads at both ends of the Beverage, or the vertical component in the sloping down leads, if used. **Fig 7-130** is a picture of some of W8LRL's end-fire Beverage arrays.

2.16.4. End-Fire Plus Broadside Combinations of Beverages

The ultimate array of Beverages is, like arrays of verticals, the end-fire/broadside combination. By putting two end-fire Beverage groups side-by-side (in-phase) you can further narrow the forward lobe and hence improve the RDF.

Fig 7-131 shows the radiation patterns of such an array, where each cell consists of two 160-meter long Beverages, spaced 5 meters, with a stagger distance of 30 meters and

Fig 7-130 — One of the many end-fire phased Beverages at W8LRL's super station.

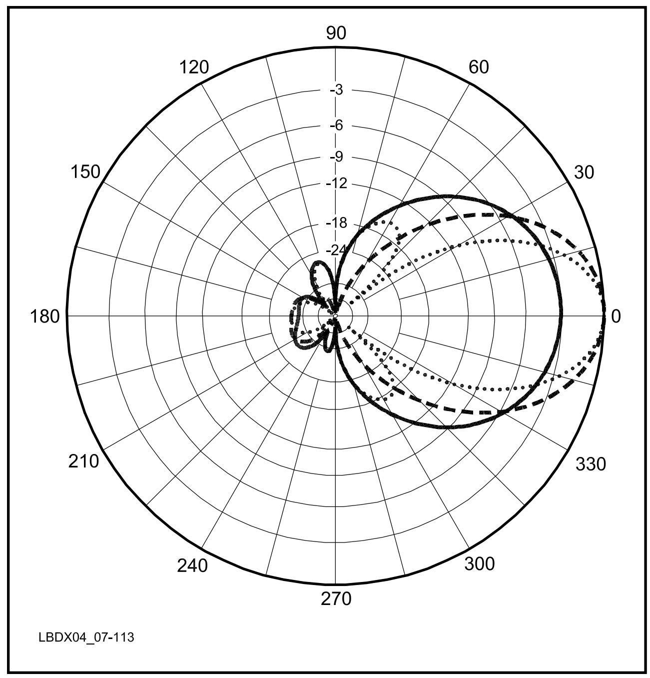


Fig 7-131 — Two end-fire cells in a 1.83-MHz broadside combination: Solid line: a single cell; dashed line: two cells with 90-meter separation; dotted line: two cells with 135-meter separation. See text for details.

phasing angle = 120°. A single cell is also shown in Fig 7-131 as reference.

Table 7-36 gives the numerical data. This is simply awesome! The broadside configuration, obviously, does not do anything to the geometric F/B (at 180°) of the array elements.

2.16.5. Feeding an End-Fire Beverage Array

Thanks to Tom, W8JI, who brought to my attention the crossfire principle for feeding an end-fire array, feeding an array of end-fire Beverages has become very simple. Let us examine the design of an array for use on 160, 80 and

Table 7-37
Design Data for the 3-Band End-Fire Phased Beverage Array

	1.825 MHz	*3.65 MHz*	*7.15 MHz*
20-m Spacing (degrees)	45°	90°	180°
Phasing Angle (degrees)	139°	98°	16°
Crossfire Phasing Cable Length (Phasing Section)	180 – 139 = 41°	180 – 98 = 82°	180 – 16 = 164°
Physical Length (VF = 0.66)	12.34 m	12.34 m	12.6 m

40 meters. First rule: the stagger at the highest frequency to be used should be no longer than λ/2 (see Section 1.20).

For a spacing of 20 meters, and considering we want to lift the zero angle about 30° off the ground, we can calculate the design data using Table 7-1.

In Section 1.20 I discussed the crossfire feeding principle for end-fire arrays. This principle can, of course, be applied to phased Beverages just as well as to phased verticals. From **Table 7-37** the length of the crossfire phasing cable, expressed in meters, is the same for the three bands, because the system, as explained in Section 1.20, tracks frequency. Look at **Fig 7-132** and imagine A and B as feed points for two Beverages instead of two verticals. What we call here the "stagger distance" is what we used the call the "spacing" in the case of two end-fire fed verticals. In case of two verticals, they are positioned in line toward the target. With Beverages, we need to offset the antennas, otherwise the back Beverage would literally lay on top of the front Beverage. We call this offset the "spacing." It is not critical at all. I would recommend a spacing equal to at least the height of the Beverage wire above ground.

With a Beverage you do not even need an inverting transformer (180° phase shifter). All you need to do is flip the connections of either the primary or the secondary at one of the transformers T1 or T2). From the back antenna (Fig 7-129) you run the phasing line length (12.34 meters in the table above) plus, for example, 5 meters (L) to the combiner box. Run the same length L (5 meters) from the front element to the combiner box (**Fig 7-133**). In order to keep the phase delay as accurate as possible under all circumstances, we use a 0° hybrid combiner and not a simple parallel combiner. The combiner box contains the hybrid (T3) and a 2:1 impedance ratio transformer (T4). Both these elements are described in detail in Section 1.22.

Assuming we use 75-Ω feed lines, R (the load resistor on the hybrid) is 150 Ω, and the output impedance of the combiner is 37.5 Ω. Thus we need to transform this impedance back up to 75 Ω using a transformer (T4) as shown in Fig 7-26.

As we do use the coaxial cable to give us the correct phasing angle (line length = phase angle), we must make sure that the SWR on the line is *flat*. This means we have to pay a little more attention to matching the feed line to the phased Beverages compared to a system using just one Beverage. We have seen in Section 2.5.4 and Fig 7-88 that we can obtain a flat SWR curve better than 1.2:1 from 160 through 40 meters. That should do.

It is evident that if we put up such a good performing Beverage array, we should pay the necessary attention to prevent ingress from common-mode signals. Make sure you have *good* antenna earth grounds at the receiving end. If necessary use two or three ground rods equally spaced at a distance at least equal

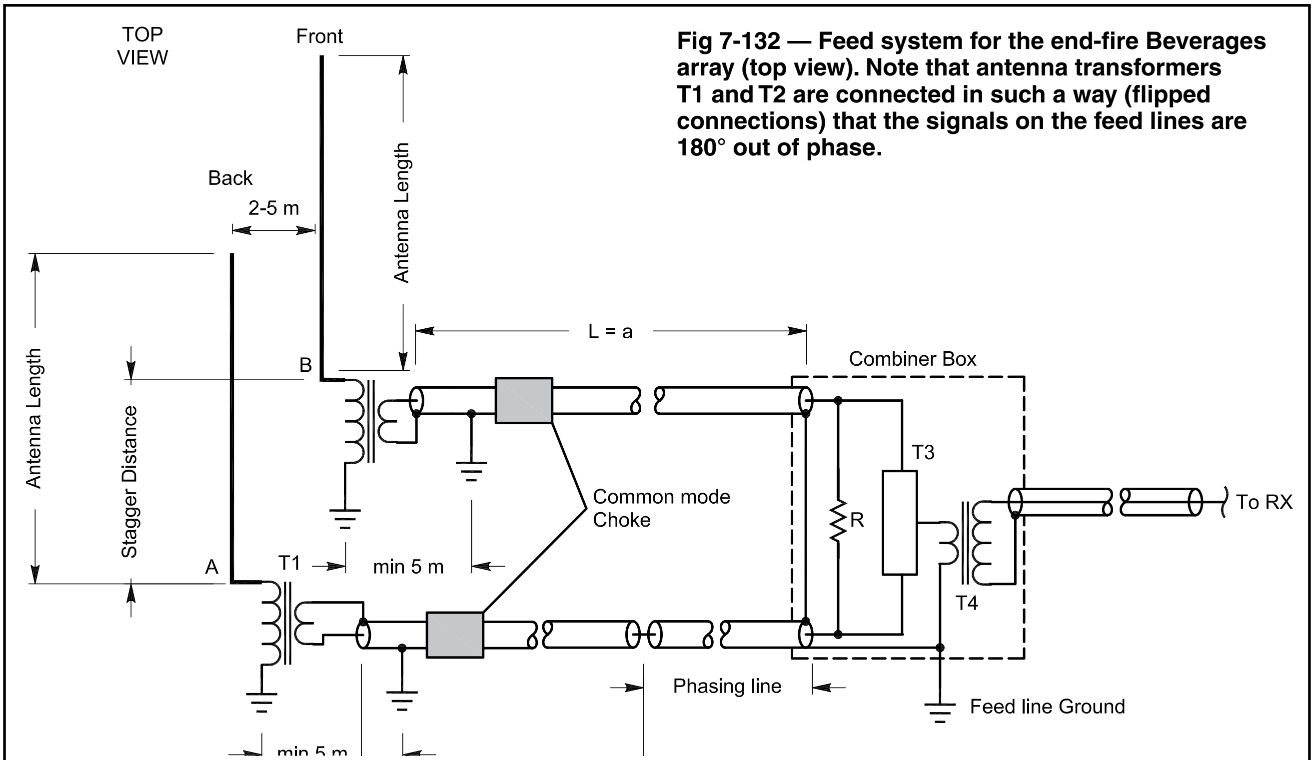


Fig 7-132 — Feed system for the end-fire Beverages array (top view). Note that antenna transformers T1 and T2 are connected in such a way (flipped connections) that the signals on the feed lines are 180° out of phase.

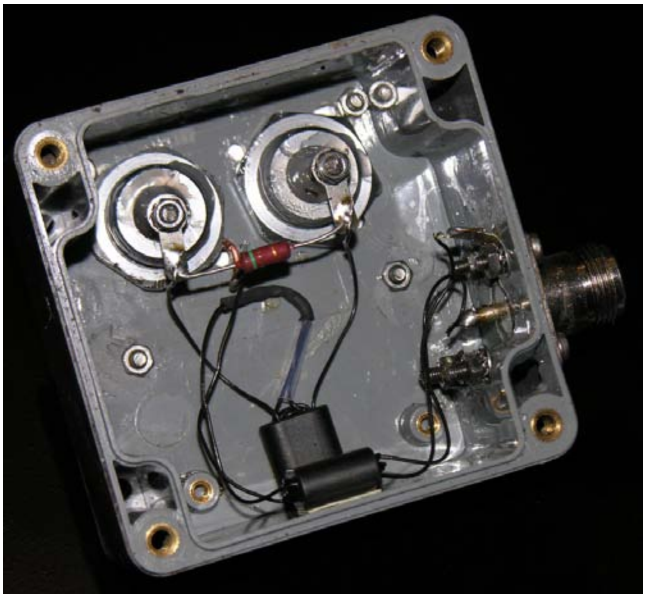


Fig 7-133 — Combiner box for the 2-element end-fire array at ON4UN.

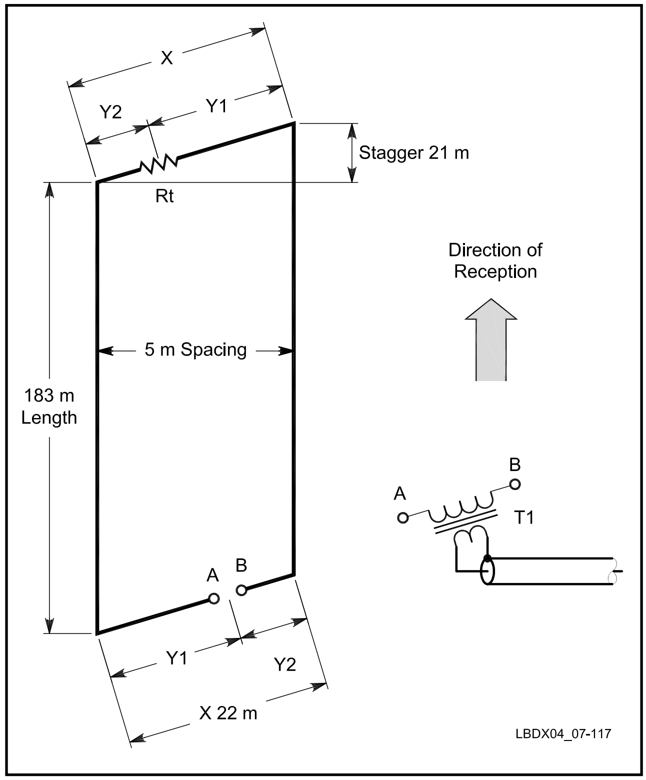


Fig 7-134 — Top view of the layout of the W8JI Parallelogram array, which is basically a 2-element end-fire array of Beverages.

to the length of the ground rods. Use quality common-mode chokes (Section 2.7.2.9) on both lines, near the transformers T1 and T2.

Use good quality coaxial cable, preferably quad shielded water-blocked RG-6 or ½ inch Hardline. No cheap stuff — you are building a top-notch receiving antenna!

Now you are all set for a unique experience in low-band listening delights!

2.16.6. The W8JI Parallelogram

Thanks to Tom, W8JI, for the following description of the antenna.

Although it looks like a large horizontal loop, this antenna actually is an array of two ground-connection-independent end-fire staggered Beverages. The short end wires form two single-wire feed lines at each end. One end wire is series terminated with exactly twice the resistance of a normal Beverage, while the short wire at the opposite end is the feed-point.

Stagger and spacing determines feed and termination location, the offset at the two ends being mirror images. A top view, looking down from straight above the antenna is given in **Fig 7-134**.

Notice the feed point is offset toward the termination end (front) of the antenna, and away from the null direction. This is typical for crossfire phased arrays. Crossfire arrays respond away from the delay line direction, exactly opposite conventional arrays. The termination is offset the same amount, but moves toward the feed-point end of the antenna.

The feed-point terminals, being floated (push-pull), provide 180° phasing between the two elements. The extra line length to the forward (left and front) element provides the "Stagger" delay. Consider the actual wire length of a "short side" called "X" (which is the same as Y1+Y2). This length is the same on both ends of the antenna. The difference between Y1 and Y2 must equal or be slightly less than stagger (S). To determine the offset of feed point and load:

1) Measure the length of the end-wire, length "X."
2) Measure stagger in the end-fire direction, "S"
3) (X – S)/2 = Y2
4) Y1 = X – Y2

You may want to slightly offset the feed by making Y2 longer and Y1 shorter. This will move the null upward, forming a cone. It is best to model the results. Let's review a system.

X = 22 meters
S = 21 meters (stagger)
Y2 = (X – S)/2 = 1 meter
Y1 = 22 – 1 = 21 meters

The difference is 21 meters, and that is the phase delay = (21 × 1.83 × 360)/299.8 = 48.3°. We have a 180° shift at the push-pull feed point (between points A and B), so –48.3° rotates to +131.7°.

We have a 131.7° lead in the forward element, with a 48.3° spatial array delay. In the forward direction (toward termination) phase is (–48.3°) + 131.7° = 83.4° out-of-phase. This results in nearly the voltage of one element alone, when the two element outputs of the long sides are summed. Toward the null (feed point) the phase is +48.3° + 131.7° = 180° for a sin 180° = 0 or zero voltage, a perfect 180° null.

On the second harmonic, forward array feed system phase is –96.6° rotated to +83.4°. Spatial array phase is now 96.6°, or –96.6° toward termination. The result is (–96.6°) + 83.4° = –13.2° in the forward direction. The result is nearly twice the voltage of one element in the forward direction. In the reverse direction, array phase is 96.6° – 83.4 ° or 180° out-of-phase. We once again have zero back-fire response.

The general pattern holds true for any length of S less than 180°, although grating lobes would make the pattern useless. This array, with 18-meter stagger S, is usable from

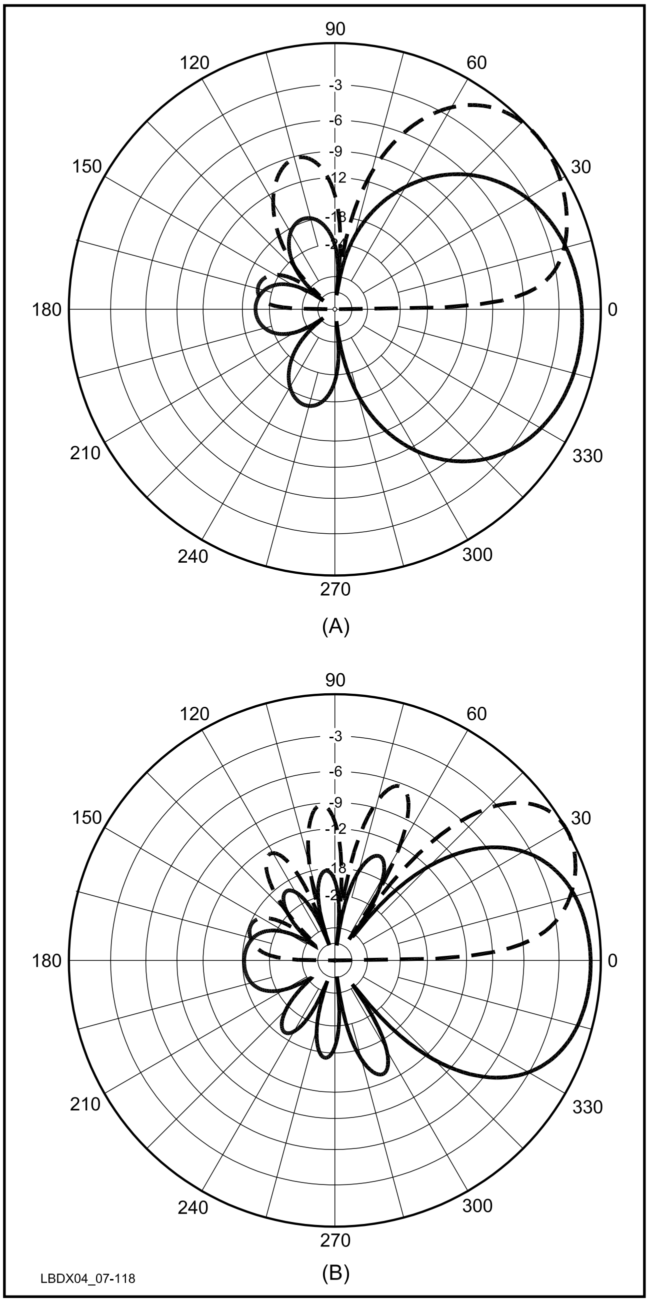


Fig 7-135 — Combined azimuth/elevation patterns at 160 and 80 meters for the 150-meter long Parallelogram array. These are very similar to the patterns obtained by two individually fed end-fire Beverages.

about 5 MHz down to VLF.

[*ON4UN*: Note that exactly the same analysis can be done with individually-fed end-fire arrays. You will also see that the shorter the stagger distance the lower the gain while the response off the back remains a perfect null, not necessarily in-line but at an offset angle (in elevation and azimuth). See **Fig 7-135**.]

W8JI continues: Placing the feed point and termination centered on the end-wires is like using 180° phasing. Placing the feed point and termination at the stagger minus side length distance is like using S – 180° phasing on any frequency with one exception, the antenna fires toward the feed-point offset. This always results in a perfect backfire null, regardless of frequency.

Table 7-38
W8JI Parallelogram

	1.8 MHz	*3.6 MHz*
RDF (dB)	9.2	11.8
DMF (dB)	18.6	19.6
–3-dB Angle	87°	69°

If you wish to remove signals from the rear, the W8JI Parallelogram Array is a good broadband solution. It not only requires one transformer, one termination, and no ground systems at either end, it also is useful on several bands without switching anything!

Remember the above example is a reasonably short antenna, performance improves as side-length increases. There is no reason why this antenna can't be used from a band where it is only λ/2 long to bands where it is several wavelengths long, as long as you properly choose stagger and width.

Table 7-38 shows calculated performance data on 160 and 80 meters.

2.16.6.1 Feeding the W8JI Parallelogram

The feed point impedance is approximately 900 Ω, so a 16:1 transformer with minimal inter-winding capacitance should be used. The transformers described later in Section 3.9 (feed systems for elongated loops) are well-suited for this application. Don't forget to install a common-mode choke and a good earth ground approximately 5 to 10 meters beyond the choke.

2.17. Summing Up, Arrays of Beverage Antennas

The best thing about Beverages is their simplicity. Even if they are not well engineered, and not installed or tested with care, they always seem to work "pretty well." Even the simplest forms of BOG antennas seem to work. But careful engineering, proper feeding and the use of good grounds and careful attention to common-mode problems are certainly areas where many existing Beverages can be improved.

- End-fire Beverages help in canceling out the omnidirectional pick-up caused by the vertical (or sloping) down leads at both ends of the transformers (this is an important advantage of using end-fire Beverages).
- Two 160-meter long Beverages in end-fire configuration allowed me to dig one layer deeper in the 160-meter noise.
- The crossfire feed system is simple. (Thanks Tom, W8JI!) It works on up to three bands without any changes or switching.
- To obtain best results (which are just short of spectacular), make sure you terminate *both* the antennas properly and have a very low SWR on the feed lines.
- If you are limited in length, use two short Beverages in an end-fire array (even two end-fire phased λ/2 Beverages are pretty good!)
- If you can put up a pair of λ/2 spaced Beverages, put up two groups of end-fire arrays rather than a broadside group of individual Beverages.
- You can't imagine how good an end-fire pair is unless you compare it side-by-side with a single one in the same direction. I have seen some spectacular improvements, like signals generally off the back dropping from S7 to S1 or

less! Simply amazing.

- In a QTH where there is no direction where most of the noise comes from, it may pay off more to reduce the –3-dB forward angle by going to a broadside array, than trying to achieve the best possible front-to-rear with an end-fire array.

2.18. Beverages and Beverage Arrays Compared

In **Table 7-39** I compare DMFs and RDFs and the forward lobe at a 20° elevation angle for a variety of Beverages and Beverage arrays. All figures are for 1.83 MHz, over "average" ground.

2.19. Beverage Maintenance

Once you have built or rebuilt your Beverage antenna system, keep full records of all technical data, so that you can check periodically if anything changes.

If you have a feed line system with hubs and remote switch boxes, check the SWR on the feed line system as well as the total loss. To measure the SWR, connect calibrating resistors (75 or 50 Ω precision loads) at the end of the lines (where you would normally connect your Beverage transformer). Use a scanning antenna analyzer (I use the AIM 4170) to check the SWR, and save the plots. Next, short circuit or open circuit the lines, at the points where you just had the load resistors. Measure the return loss and calculate the cable loss (cable loss = ½ of the return loss when the end is open or shorted). The AIM 4170 does that automatically and shows the cable loss on the screen of your computer. Keep a record of all these measurements.

After the feed line system, also check the technical parameters of the antennas. You can do this from your shack. I connect the master feed line, which is normally connected to the receiver (transceiver), to my antenna analyzer. Then I run a check on each of the antennas, logging all the relevant parameters. Save the plot data for future reference.

Now that you have the initial health data for your system (antennas and feed system), don't forget that a periodic checkup, as with you doctor, will keep you out of trouble.

At least twice a year I make new plots and compare with the old ones. Have the results of the original (last) tests, it will be easy for you to find the source of any problem that has arisen.

2.20. Summing Up, Beverage Antennas

I have been using Beverages since 1968. For me, Beverage antennas have undoubtedly been the key to working my last 50 countries on 80 meters. As far as 160 meters is concerned, I would not like to think what it would be like without them.

Since I started using end-fire phased Beverages, a new world again opened up for me. Unfortunately, I cannot run broadside phased Beverages; the land available to me is much too small for that ("only" 10 acres, **Fig 7-136**). I cannot run end-fire phased Beverage arrays in all directions, but I must say that my hearing to the US has been dramatically improved through the use of an end-fire pair. That is because most of the amateur-made noise on the bands (for example, during contests) comes from the southeast, the opposite direction to the USA. Well, all of that has dropped another 20 dB since using the

Fig 7-136 — View to the northeast (toward Japan) across the fields where I have my Beverages in winter. In the front, 10 and 15-meter Yagis; to the right, the 160 meter $\lambda/4$ vertical.

Table 7-39
Performance Data for Beverages on 160 Meters Over Average Ground

Beverages and Arrays of Beverages	*DMF (dB)*	*RDF (dB)*	*3 dB Angle (degrees)*	*Output (dBi)*	*Reference*
89 m long single Beverage	11.1	6.5	122	–15.9	Table 7-23
176 m long single Beverage	10.6	10.1	86	–10.6	Table 7-23
268 m long single Beverage	21.3	12.9	66	–7.8	Table 7-23
Broadside 160 m long Beverages, 90 m spacing	21.3	11.9	48	–7	Sect 2.16.2
Broadside 300 m long Beverages, 90 m spacing	23.1	14.2	44	–2	Sect 2.16.2
80 m long end-fire Beverages Stag = 30 m, ψ = 140°	20.0	9.7	77	–15.5	–
160 m long end-fire Beverages Stag = 30 m, ψ = 140°	30.1	11.6	69	–9	Sect 2.16.3
300 m long end-fire Beverages Stag = 30 m, ψ = 140°	33.8	13.9	57	–4	Sect 2.16.3
160 m long Beverages in end-fire/broadside array*	34.0	13.0	46	–6.4	Sect 2.16.4
160 m long Beverages in end-fire/broadside array**	34.7	14.1	34	–6.4	Sect 2.16.4

*End-fire cell ψ = 140°, stagger: 30 m, broadside spacing: 90 m
**End-fire cell ψ = 140°, stagger: 30 m, broadside spacing: 135 m

end-fire pair. It is just incredible. At first I though the antenna was "dead" as there seemed to be so little noise.

I use the Beverages all the time on 80 and 160 meters. Unfortunately, I do have to take down most of my Beverages during the summer, which for me was "off-time" on the low bands.

3. THE FAMILY OF ELONGATED TERMINATED LOOPS

I would like to thank Gary Breed, K9AY, for reviewing this section.

Loops have been around a long time. W8JI told me he'd been using arrays of loops to work Japan in the 1970s through 100-kW LORAN pulse emissions on 160 meters. These made it possible for Tom to be the first eastern-USA station to work Japan through the LORAN noise. In those days the JAs transmitted just above 1900 kHz, where all the East Coast LORAN pulse transmitters operated. Tom used an array consisting of no less than eight wire diamond elements in a series end-fire configuration, terminated at the far end. Tom told me it was Mr Top Band himself, the much revered Stew Perry, W1BB, who told him that commercial antennas were also available using that principle as early as the 1960s, and that got Tom going! Nowadays, when we talk about small (terminated) loops, everyone thinks of EWEs, Flags, Pennants and such. This section of the chapter explores these antennas.

The EWE, the Pennant, Flag, Delta, Diamond, and the K9AY all operate on the same principle of the terminated loop. The differences are in the way ground is used, how they are fed and their physical shape.

All these antennas basically produce a cardioid pattern. See **Fig 7-137**. The depth of the null and the null angle vary with the shape of the loop and the termination. The optimum termination value can vary with local ground conductivity. The directivity figures (DMF and RDF) are listed in **Table 7-40**. The directivity is very much the same as for a half wave long Beverage that has been terminated for best directivity with a complex termination.

Most of the published designs are based on optimum dimensions, but almost *any* loop size and shape, grounded or ground-independent, can be made to work. First, the feed point and termination must have significant separation — for example, between the two bottom corners of a triangle or square — or be separated by the ground connection (as in the K9AY loop) or by the two ground connections (as in the EWE). Then it needs the right termination to establish the current distribution and phase along the antenna conductors to create the directional pattern. The termination may be resistive, or it may be complex, requiring R+L or R+C to obtain the desired pattern.

While just about "any dimension" will give you a cardioid pattern for a specific termination resistance (impedance) over a more or less narrow frequency band, only the optimized designs combined bandwidth with performance and ease of terminating and feeding (with resistive impedances).

On this issue of the loop's size, K9AY wrote: "Most of the terminated loop designs are 'low pass' in their characteristics. That is, they are quite consistent in feed point Z, termination and pattern below a cutoff frequency. For the K9AY loop, the published dimensions in September 1997 *QST* have that cutoff around 160 meters, which means that it behaves similarly in the AM broadcast band, or lower. A different termination is needed for optimum 80-meter performance, which is just above the "cutoff" frequency. The tradeoff is in the received signal levels. Smaller antennas, though more broadband, need more preamp gain. Larger antennas capture more signal. The voltage across the feed point is proportional to the area enclosed by the loop."

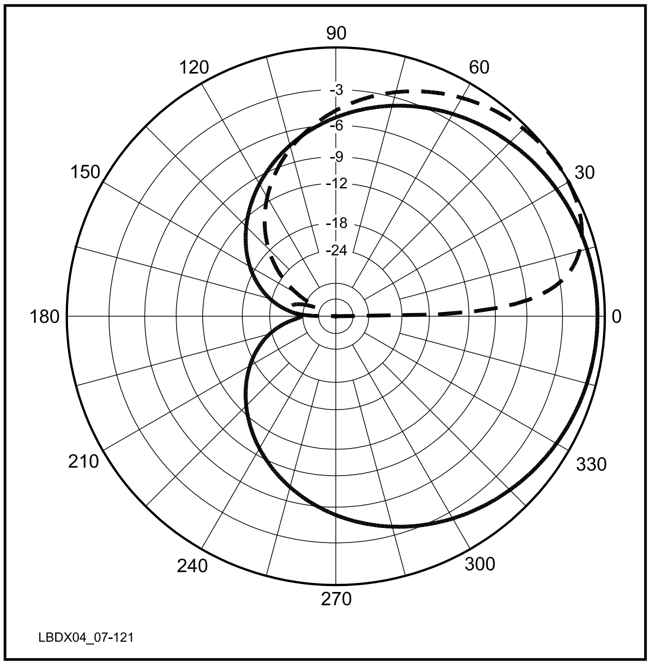


Fig 7-137 — Radiation patterns of a EWE antenna (solid line = azimuth; dashed line = elevation). Whatever the dimension, whatever the shape (inverted U or inverted V), the shape of the pattern is always the same, provided the termination resistance is correct.

Table 7-40
Typical Directivity Figures for Elongated Terminated Loop Antennas

	DMF dB	*RDF dB*
Typical Elongated Loop	–11	–7.4

Fortunately all of these loops have been optimized (thanks to K9AY and the late Earl Cunningham, K6SE) so that their cardioid null is constant over a wide frequency range (160, 80 and 40 meters for most designs) and this can be obtained with a simple resistive termination equal to the feed impedance of the antenna.

These elongated terminated loops are really a simple pair of verticals with a horizontal feed line, one being base-fed, the other being fed via the top wire. That provides crossfire phasing. When the element spacing is less than $\lambda/4$, this always fires toward the feed point end of the array. When properly terminated the loops show nearly constant current at all points on the loop. This constant current is achieved by terminating the far end of the loop in the same impedance as the feed point. Thus, the portions considered to be the vertical "elements" have equal current, with a phase equal to 180° plus the electrical length of the connecting wires, which is slightly more than the element spacing.

It is impossible to make a simple comparison between two phased bottom-fed verticals. Due to the terminating resistor the antenna works much more as a traveling-wave antenna than as a standing-wave antenna. The antenna current is nearly uniform or slowly tapers (due to loss from radiation and resistance) in the antenna system and contains no or greatly suppressed "end reflection." It is the terminating resistor that equalizes the currents. The conductor and radiation losses are compensated by slight changes in the value of the resistor. Therefore the antenna has no standing waves and can indeed be called a traveling-wave system. This is also why the antenna is very broadband, and why it exhibits a good F/B on both 160 and 80 meters.

In other words, the terminated loop is like two constant-current verticals with classic end-fire phasing. The only difference is the horizontal component in the radiation due to the radiating delay line (sloping or horizontal) of the terminated loops.

This is the operating principle for the entire family of elongated terminated loops. Sloping (inclined) wires used in some of the designs are just like a horizontal wire in combination with a vertical wire, and the same reasoning as above can be used.

3.1. Modeling Small Receiving Loops

K9AY, loop specialist par excellence, wrote on the Top Band reflector: "*NEC-2* modeling programs are sufficiently accurate for modeling these loops, even for ground-connected antennas, but do not try the usual technique of increasing the number of segments until convergence is reached. When designing the original K9AY loops, I found that the model closely matched observed behavior with one segment for each 1½ feet of wire (on 160). Models for other loops behave similarly using this segmentation scheme."

There is nothing magic about the dimensions of a receiving antenna such as the K9AY loop or the EWE. You can easily design your own using a *NEC-2* based modeling program such as *EZNEC*. Just enter the dimensions and change the value of the terminating resistance until the best F/B is obtained. If you are looking for a well-balanced design, however, you will probably end up with the dimensions published in the original design.

I'd like to warn you once more that there is a difference between antenna modeling and antenna building. Building a receiving loop (or any other receiving antenna for that matter) is not a mathematical exercise. I have seen umpteen radiation plots with high praises for a 67.37 dB F/B for a model. Great! Great what? Great mathematics, that's all. Look, the proud designer of this antenna needed a 937.735 Ω terminating resistance! Go to the radio shop, or call Allen-Bradley or Ohmite and ask them for such a resistor. To obtain that kind of phenomenal F/B the ground conductivity must be stable "like a rock." It should not vary more than 0.01 Ω. And the length of the antenna should be measured to within 1 mm at worst…

There is a definite difference between a model and reality, as I have said over and over. And I will refrain from publishing more digits of accuracy than make sense. We should think in terms of absolute and relative error all the time.

A little experimentation is normal after building the antenna according to the model. Also, once the antenna is working well enough to obtain 20 dB or more F/B, it takes careful measurements to discern any additional improvement.

3.2. The EWE Antenna

Floyd Koontz, WA2WVL, is the originator of the EWE. His publications (Ref 1263 and 1264) are must-reads on this novel receiving antenna. Incidentally, the curious name "EWE" came about because Floyd noted that his new antenna looked very much like an upside-down letter "U." He jokingly submitted a drawing of a female sheep — a ewe — to headline his February 1995 *QST* article, which he impishly titled with the triple entendre "Is This EWE for You?"

A EWE is small, requires little engineering, and can be designed to cover 80 and 160 meters. Moreover, the EWE is low profile and can be built for little money. In appearance, the EWE resembles a very short Beverage, though in fact it is an array of two short vertical antennas, where the horizontal wire is part of the (radiating) feed system. See **Fig 7-138**.

The horizontal wire is only about 5 to 15 meters long and is about 3 to 6 meters above ground. In other words, a EWE can fit in many tiny yards. The original description by WA2WVL uses 3-meter high "elements" with a spacing going from 4.5 to 18 meters, depending on the band (80 or 160 meters).

A good rule of thumb is to build a EWE where the length is twice the height. This rule can be followed for elements going from 3 to 6 meters in height, and will result in an optimum termination resistance of approximately 1000 Ω for both bands. The smaller the array, the easier it is to obtain a good deep null — but the lower the output level. The larger versions will be near the cut-off frequency where a deep null cannot be obtained.

The output (gain) of the antenna depends on its size. The gain for 160 meters varies typically between –20 dBi and –30 dBi over average ground, and gain is about 9 dB higher on 80 meters. The inverted-V-shaped EWE looks quite attractive and can easily be made into a hand-rotatable receiving antenna: Just walk outside and anchor the sloping wires to a different set of ground rods. It should be useful for DXpeditions as well.

As with all ground-based antennas a good ground will lower the takeoff angle (typically about 35° for average ground). As usual, you have to be cautious with the modeling results as they are calculated for a flat and perfectly homogenous ground.

The value of the terminating resistor is quite critical if you want to obtain a good notch in the back response. Since the ground resistance is part of the terminating resistance, a good low-impedance, stable ground is required. The ground resistance of a single ground rod may vary from as low as 50 Ω to as much as several hundred ohms. Although the terminating resistance is high (typically 1000 Ω), a ground consisting of more than one ground rod or a combination of a ground rod with various short radials is a good idea.

Varying ground conductivity not only has an influence on the effective total termination resistance but also on the velocity factor of the various elements of the antenna. The velocity factor is critical in determining the current distribution necessary to obtain a cardioid pattern.

It has been reported that the best way to tune an array for best F/B is to find a medium-wave BC transmitter in the back of the antenna. Using a medium-wave signal during day time ensures that you will receive it on ground-wave, and as such have a constant and stable signal source. Tune the terminating resistor for a full notch, striving for a minimum of 25 dB. This value will be very close to the best value for 160 meters and

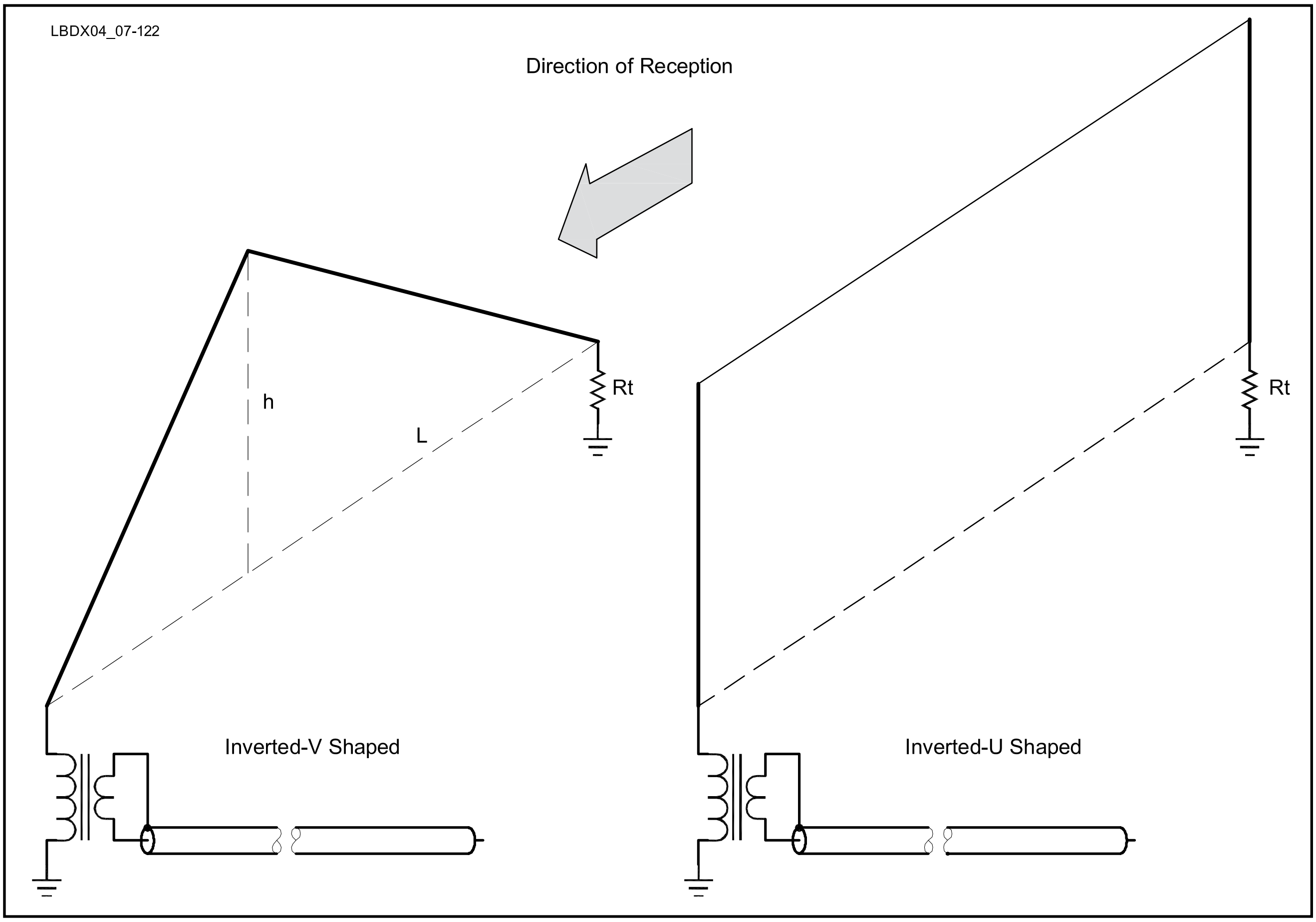


Fig 7-138 — Layout of the EWE antenna. The inverted-V-shaped EWE requires one insulated support; the inverted U-shaped EWE requires two base-insulated metallic supports that are part of the radiator. Unlike Beverages, all these terminated loops that have the termination resistor sitting on one side and receive from the direction opposite to where the resistor is located.

close to what's best on 80 meters.

You should use broadcast stations between 1600 and 1700 kHz to do the null adjustment if you can. It is impossible to reliably adjust the termination resistance using signals arriving by skywave. You could also use a nearby local ham (in-line off the back) or set up a small signal generator off the back, using a small vertically polarized antenna, at a distance of at least 1 λ from the EWE (which means outside its near field).

The feed point impedance of a properly terminated EWE usually has a reactive component, but the real part is usually in the 400 to 600-Ω range. Being a grounded antenna, the EWE has an advantage over an "elevated" elongated loop, in that problems with common-mode ingression at the antenna feed point are a little easier to solve.

The EWE antenna can be fed with a 1:9 transformer such as described in Section 2.7.2. Further, common-mode signals getting on the shield of the feed line can be avoided by applying the techniques outlined in Section 2.7.2.9. The features of the EWE can be summed up as:

- Easy to build, but requires good grounds and radials at both ends.
- Less critical in feeding compared to elevated loops.
- More sensitive to varying ground conditions compared to elevated loops.

3.3. The Rectangular Loop or Flag Antenna

A major drawback to the EWE is its extreme sensitivity to any change in the local soil conductivity. EA3VY was the first to come up with the idea of adding a ground wire between the bottom of the two verticals as an effective way to minimize the effect of different soil conductivities.

As pointed out in Section 3, many dimensions will work but only the optimized ones provide the best directivity over the largest frequency range with a single termination resistor. See **Fig 7-139**. These standard values developed by EA3VY and K6SE are 4.27 meters (14 feet) high by 8.84 meters (29 feet) wide, with a bottom wire height of 2 meters (6 feet) above ground. These dimensions result in an antenna that has excellent characteristics from 7.5 MHz down. K6SE said, "The height by length ratio of the Flag is not a 'happy medium' value. The correct dimensions were arrived at by designing the antenna to be broadband (ie, exhibit zero reactance at its feed point over a very wide frequency range). This required that the termination resistor value be equal to the feed point resistance. To meet these criteria, the height and width dimensions had to be juggled until everything fit into place."

This configuration has a gain of approximately –28 dBi on 160 meters, –18 dBi on 80 meters and –10 dBi on 40 meters. The loop is terminated with 950 Ω and fed in the centers of

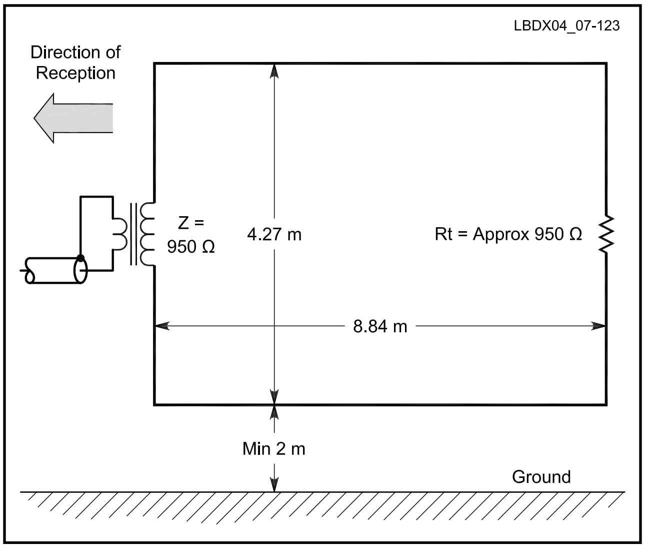


Fig 7-139 — Optimized dimensions for the 160/80 meter rectangular loop.

opposite vertical sections. Its input impedance on 160 meters is 950 Ω with a small reactive component (10 to 50 Ω).

You can scale up the dimensions to get more signal output from the antenna. If you are not interested in using the antenna on 40 meters and if you have a lot of space, a flag twice the size of the K6SE/EA3VY design has a gain of –16 dBi on 160 meters, 12 dB more than the standard Flag. That is quite a difference and can be helpful overcoming problems with common-mode signal ingress (see Section 3.9). This scaling principle holds true for all types of elongated loop antennas.

A properly designed Flag exhibits a high F/B ratio over any type of soil and at virtually any height above ground without need to change the dimensions or termination value. The directivity figures (DMF and RDF) are given in Table 7-42 later in this chapter.

K6SE, who had thoroughly tested all shapes and sizes of elongated terminated loop antennas, reported that the Flag antenna is probably the best of all configurations from the standpoint of being broadbanded and having gain. It has about 6 dB more gain than an equal-sized Pennant (see Section 3.4).

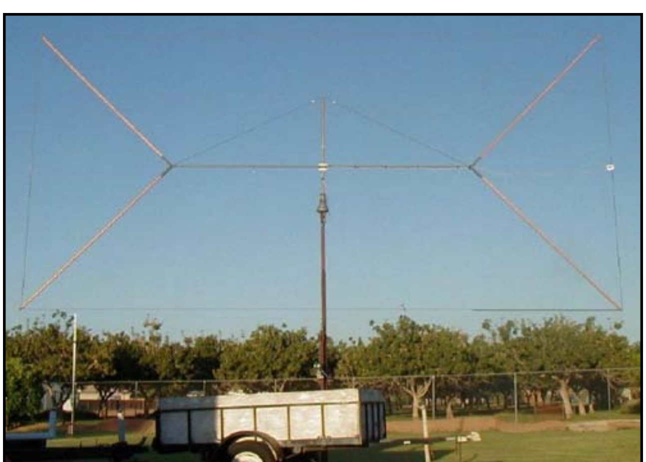


Fig 7-140 — The W7IUV rotatable Flag.

The shape and the size of this loop makes it attractive for a rotatable version. W7IUV describes such a design on his Web site (**w7iuv.com**). **Fig 7-140** shows W7IUV's elegant loop. W7IUV says: "The choice of boom material was easy; a friend had some chain link fence top rail laying in his horse pasture. I volunteered to clean it up for him. I wanted to use bamboo for the spreaders, but none could be found for free or even cheap. Several materials were experimented with before settling on wooden clothes poles. This material is almost as light as bamboo and is readily available in most lumberyards and home improvement stores. The spreaders were attached to the boom with square steel tubing welded to the ends of the boom. The tubing formed nice 'sockets' for the wood poles to slip into. One through bolt holds it in place. If welding is not your thing, consider making a spreader mounting plate from a square piece of aluminum about 12 × 1 × ¼ inch. Drill for U-bolts in appropriate places."

3.4. The Triangular Loop or Pennant

While optimizing the flag, K6SE and EA3VY developed the triangular-shaped loop. It's named the *Pennant* because its triangular shape resembles a flag pennant. The loop is fed in the center of the vertical section, while the load is situated where the sloping wires meet. It looks like the only limit as to shapes and dimensions of these kinds of loops is your own imagination.

Fig 7-141 shows the dimensions of the optimized Pennant developed by K6SE and EA3VY. As explained earlier, other dimensions can be used, but these yield the highest F/B over the widest frequency range for a single value of termination resistance. This design can be used from 160 through 40 meters.

The correct termination resistor for the Pennant is about 900 Ω, which can be placed either at the point of the Pennant or in the center of the vertical section — the results are identical. The feed point is at the opposite end. Gain on 160 meters is about –34 dBi (which is 6 dB less than the Flag), –24 dBi on 80 meters and –16 dBi on 40 meters. The exact values depend on local ground conductivity. Note that the 6-dB lower signal level is proportional to the difference in area enclosed

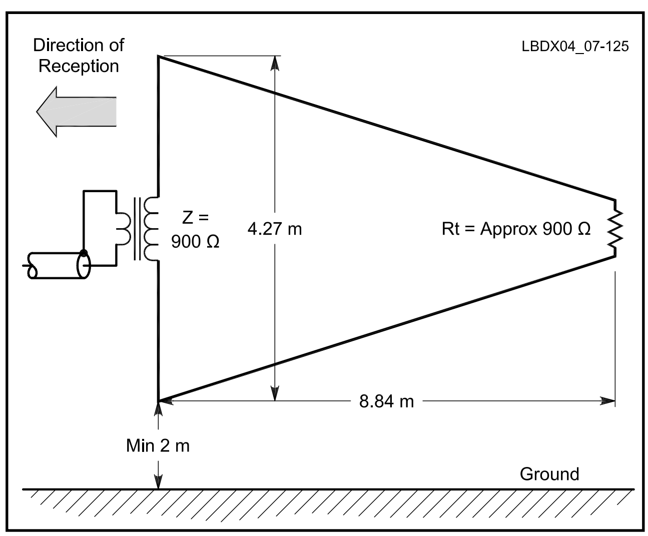


Fig 7-141 — Layout and dimensions of the optimized Pennant antenna.

by the loop, which is half as much as the Flag: ½ voltage = ¼ power = –6 dB.

3.5. The Delta-Shaped Loop

A EWE in the shape of a Delta loop has every bit as clean a pattern as the more classic ones. I think this antenna is really attractive for DXpeditions and for semi-rotatable setups. Just moving the base line around is all you have to do to change directions. During modeling it appeared that the best F/B was obtained with the terminating resistor mounted approximately 20% from the bottom corner of the Delta loop. The antenna is fed in the opposite bottom corner. See **Fig 7-142**.

The FO∅AAA Clipperton Island DXpedition used this antenna in their operation. It is the only design that requires only one support and can be easily rotated manually, a very desirable feature for DXpedition use. The Clipperton team found the antenna to be very successful and many other subsequent DXpeditions have also used the Delta configuration.

The optimum termination resistance for this design is also 950 Ω. The antenna was optimized for 160 meters and its output is about –33 dBi on 160 meters, –22 dBi on 80 and –13 dBi on 40 meters.

3.6. The Diamond-Shaped Loop

The diamond shaped loop is derived from the flag and was developed thinking of a cubical quad element. The idea is to use the construction techniques and hardware for a 20-meter cubical quad element. See **Fig 7-143**.

It is not necessary to use an equilateral loop — it can be elongated if that is easier to do mechanically. The elongated version also seems to provide a wider bandwidth than the equilateral configuration, for a fixed termination resistance. The feed impedance and the termination resistance is approximately 1000 Ω in both cases and for the three lower bands.

Equilateral version:
L = H = 9 meters
Gain: –29 dBi on 160 meters; –18 dBi on 80 meters;
–10 dBi on 40 meters.

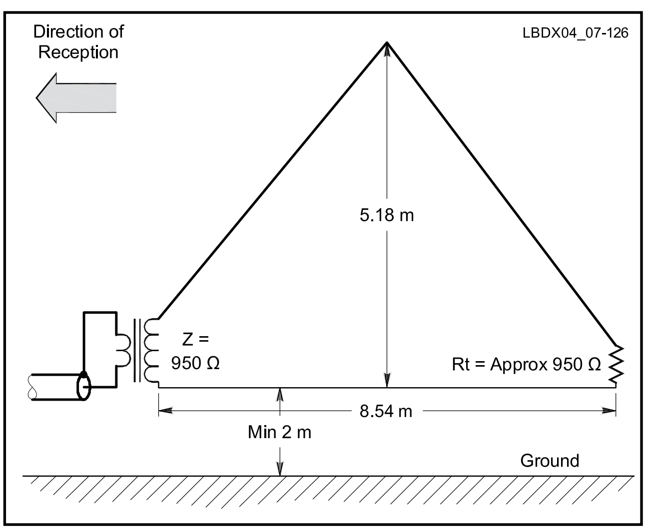


Fig 7-142 — Layout and dimensions of the optimized Delta antenna.

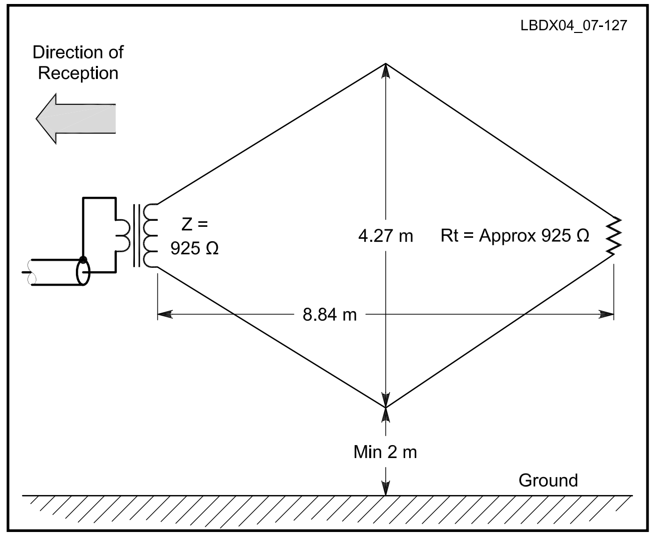


Fig 7-143 — The diamond-shaped loop does not have to be exactly equilateral. The exact shape can be adjusted by the hardware you have available.

Elongated version:
H = 7 meters,
L = 10 meters
Gain: –31 dBi on 160 meters; –21 dBi on 80 meters;
–2 dBi on 40 meters.

3.7. The K9AY Loop

Gary Breed, K9AY, described this loop very well in his September 1997 *QST* article (Ref 1265). This is another variant of the EWE, where the bottom wire of the loop is grounded in the center. See **Fig 7-144**. He pointed out in his article that the loop can really be any shape. The diamond shape K9AY used was dictated by practical construction considerations

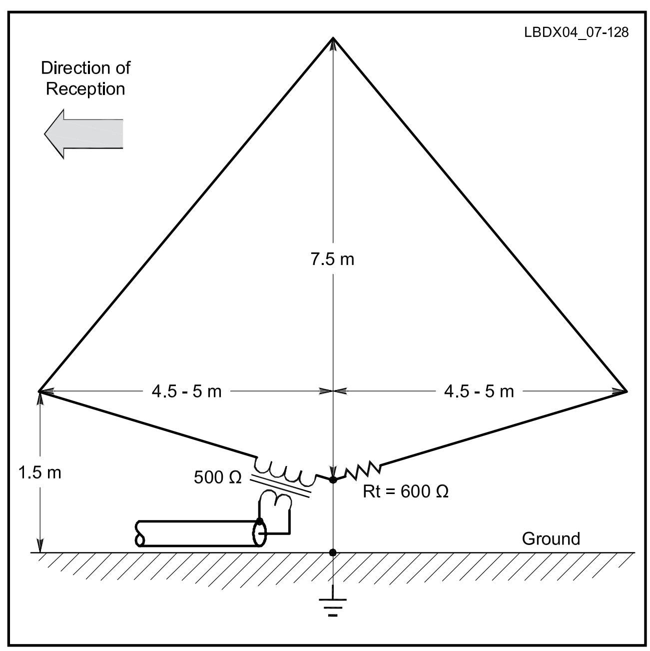


Fig 7-144 — The K9AY loop.

rather than anything else. All of the K9AY loops can be used on both 160 and 80 meters, although optimal performance may require slight adjustment of the terminating resistance, as K9AY pointed out in his article.

While in all the previously described loops the feed point and the termination are separated by distance (they are at opposite ends of the elongated loops), in the K9AY loop "separation" is achieved by grounding the loop between the adjacent feed and termination points. Having the feed and termination so close to one another makes it easy to switch directions from a single box.

Gary points out that the published dimensions in September 1997 *QST* have a cutoff around 160 meters, which means that it behaves similarly in the AM broadcast band, or lower. A different termination is needed for optimum 80-meter performance, which is just above the cutoff frequency.

K9AY says: "The shape of the loop affects the shape of the pattern. The delta/triangle shape for the K9AY loop was chosen for two reasons: 1) it only needs one support, and 2) it results in a null at about 30-40° elevation, which is ideal for in-country QRM reduction. A rectangle will work, but the null will be at a lower angle. This would be good for knocking down neighborhood noises, but it will have less front-to-back on skywave arriving signals. Something close to a square is OK, but a rectangle that wider than its height will have its null at a very low angle."

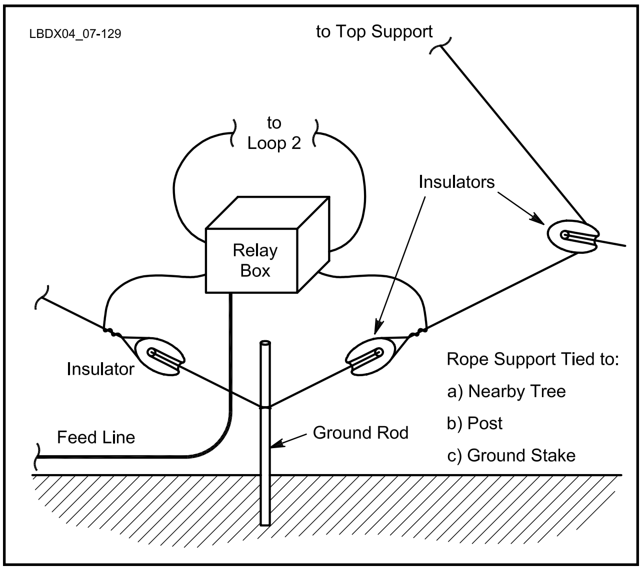


Fig 7-145 — Two loops can be suspended from one mast, which makes it possible to switch four directions. The switch box can be mounted at the base of the mast.

As with the other type of elongated terminated loops, the K9AY loop antenna can also be made smaller or larger by scaling it. The biggest tradeoff with small loops is in the received signal levels. Smaller antennas, though more broadband, will require more preamplifier gain. Larger antennas capture more signal, but are more difficult to build because they require more space, and they may not work on 7 MHz.

It is easiest to use an insulated mast for the loop. Make sure you use a good ground, such as a 1.5-meter long ground rod. The ground rod is the antenna ground and should preferably not be used for grounding the shield of the coax, especially if the coax is not buried or placed on the ground, as I've recommended previously.

You can successfully use a metal mast that is insulated from both the ground and from the antenna wires. The termination resistor varies from 500 to 600 Ω, and the feed-point impedance is around 500 Ω.

The K9AY loop can be adjusted just like any elongated loop, by varying the termination resistance for best F/B ratio. Since the highest rejection is at relatively high elevation angle, it is not possible to find a sharp null when testing on ground wave at an almost 0° elevation angle. On the other hand skywave signals are unstable in nature, and not very suitable for a nulling exercise at higher angles. But even if you do have a null at ground wave angles by pruning the resistor value for maximum F/B (maybe only 10 to 15 dB), this will still result in a 35 dB notch in the same direction at higher elevation angles (40° to 50°).

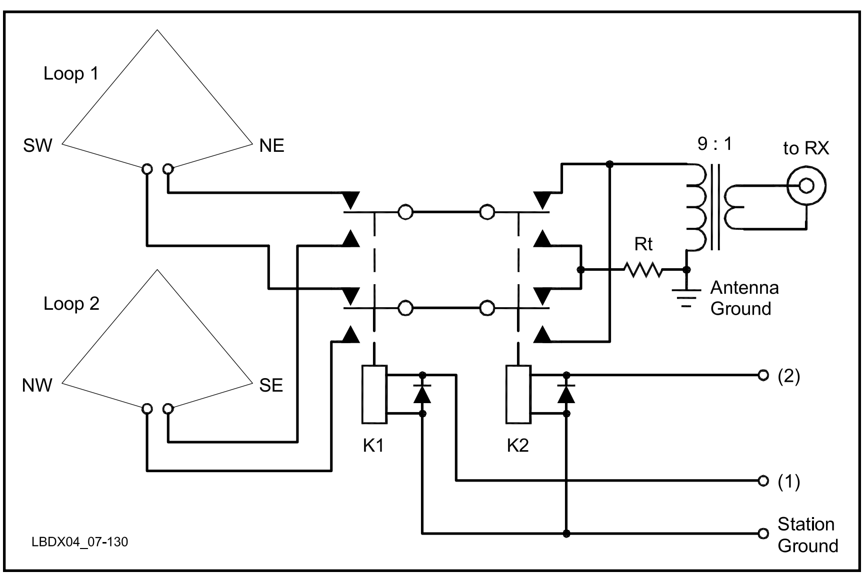


Fig 7-146 — Switch box for the K9AY loop. A split-winding transformer has replaced the 9:1 transmission-line transformer used in the original *QST* article.

The feed-point impedance is around 500 Ω, so a 9:1 transformer is indicated. I recommend a transformer with separate antenna and coax grounds using independent windings. You should install a common-mode filter near the switch box, ground the coax shield at least 5 to 10 meters from that point, and run the coax under ground. (See also Sections 2.6.2 and 2.7.2.9.)

The K9AY has a few advantages over the other loops that are above ground:

- Being grounded at the feed point, the problems with a balanced feed point don't exist.
- The feed line can be buried in the ground from right at the antenna, which

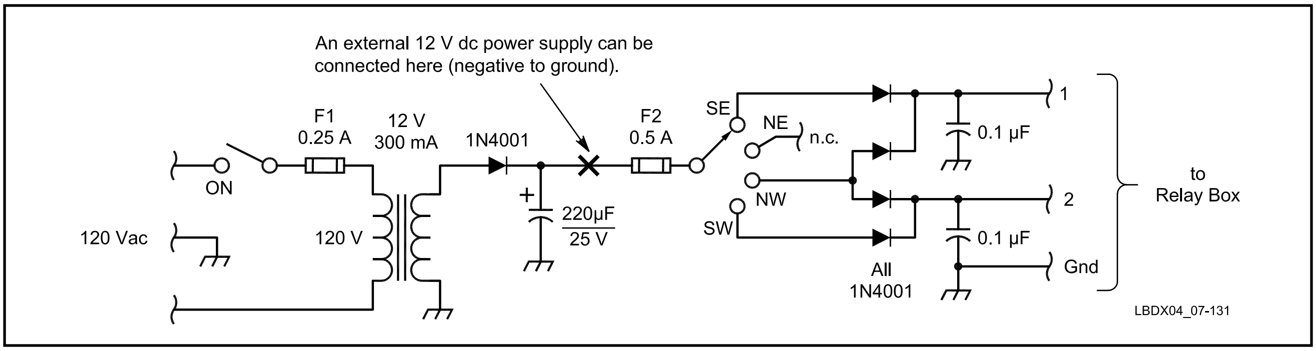


Fig 7-147 — Direction-switching control for the K9AY switch box.

Fig 7-148 — Array Solutions sells a switch box and remote control unit that also allows you to change the value of the terminating resistance from the shack.

effectively helps keep common-mode currents off the shield.

- The antenna is small and can easily be switched in four directions. See **Fig 7-145** through **Fig 7-147**.
- Because the "ground image" is part of the antenna, the signal levels with the K9AY loop are greater than with a Flag, Pennant or other ground-independent type with the same enclosed area.

There are commercially-made versions available. See **Fig 7-148** for the version from Array Solutions, at **www.arraysolutions.com**. Wellbrook Communication in the UK (**www.wellbrook.uk.com**) is another possible source.

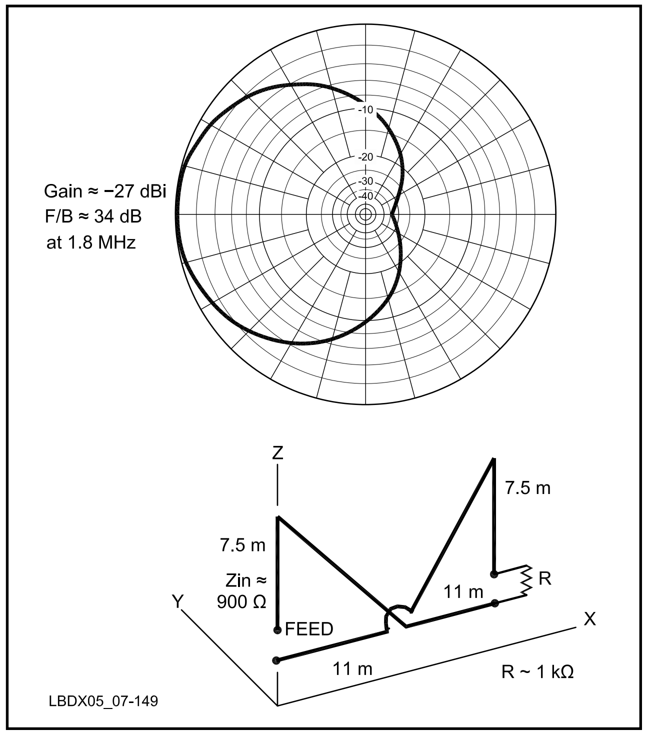


Fig 7-149 — The double half Delta loop is the most recent member of the family of elongated terminated loops. Conceived by Harry, AA7JV, its effectiveness was demonstrated during his TX3A DXpedition to Chesterfield Island.

3.8. Double Half-Delta Loop RX Antenna

(Thanks to George Wallner, AA7JV, for this antenna design.)

I guess all low band enthusiasts who were active on 160 and 80 meters during November 2009 will remember the outstanding DXpedition by AA7JV and HA7RY from Chesterfield Island (TX3A). George and Tomi have undoubtedly proven they knew what they were doing by producing — day in and day out — extremely good signals on a 15,000 km path into Western Europe. Some days I was copying their signals Q5 more than 2 hours before my sunset, and I also worked TX3A more than 1 hour before sunset in Belgium. TX3A also heard very well, and for that purpose George developed a new receiving antenna, belonging to the family of elongated loops.

His "Double Half Delta Loop" antenna requires two support poles, separated by approximately 22 meters, as shown in **Fig 7-149**. The minimum antenna height is with the horizontal wire at approximately 1.5 meters above ground, and the higher this wire, the less noise it will pick up. As the antenna is ground independent, you can just move the whole thing up a few meters without changing anything else.

Modeled over average ground (conductivity 5 ms and dielectric constant = 13), the antenna has a gain of –27 dBi and better than 25 dB F/B (RDF = 9.95 dB, DMF = 20.01) using a load resistor of 1.3 kΩ on 160 meters. On 80 meters, using a load resistor of 1000 Ω, the gain is –11.72 dBi, F/B = 22 dB, RDF = 9.15 dB and DMF = 18.31 dB. The best termination resistor value depends mainly on the ground conductivity. See the modeling file *ch7-TX3A-loop-array-fig7-149.ez* on the CD.

It is important to remark that this "double" loop has a much narrower –3 dB forward lobe angle (~100°) than the earlier described elongated loops (~140°). Hence the significantly better RDF figure as well.

Note that there is nothing special about the dimensions of this antenna. The more area the wires enclose, the larger the signals (higher gain), but the RDF may get slightly worse. Also, the larger antenna will not work on the higher bands. (The upper cut-off is approximately where the total wire length reaches ¼ wavelength.)

While the antenna can be used on 80 meters without preamp, it will definitely require a 10 to 15 dB preamp on 160 meters. The usual precaution to prevent common mode signals traveling on the outside of the feed line to ingress into the antenna (see Section 3.8) are a must.

3.9. Feeding Elongated Receiving Loops

3.9.1. Impedances

Optimized loops have a feed-point impedance that is essentially resistive over the design frequency range. The K9AY and the EWE have a feed-point impedance around 500 Ω. All the other loops (Flag, Pennant, Diamond, Delta) show about 950 Ω.

3.9.2. Symmetric — Asymmetric

Asymmetric Loops

In theory, the requirements for a EWE or K9AY transformer (like that for a Beverage) seem to be less onerous than for a Pennant or Flag because the antenna operates in an unbalanced mode.

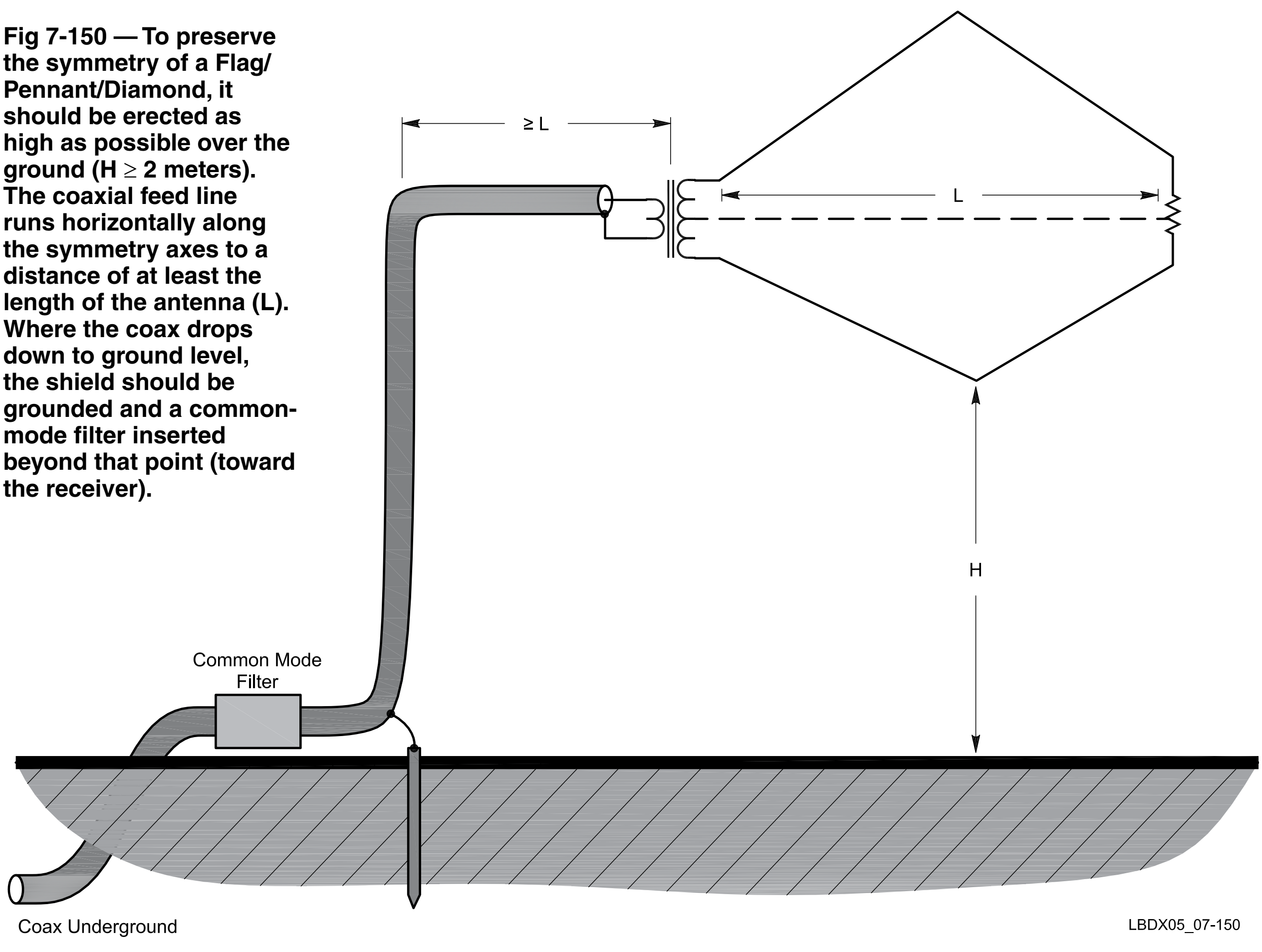
If this is truly the case, there would be no other modes of operation to suppress, where the antenna can receive signals from all directions, filling in desired pattern nulls. That is the ideal world, using an ideal ground. But it's pretty likely that these antennas have less than perfect grounds. A few 1.5-meter long rods at the feed point of the antenna do not present a 0-Ω ground impedance, and hence the antenna is less than perfectly "unbalanced." As such, it is susceptible to common-mode ingress onto the outside of the feed line or onto any other conductor attached directly or indirectly to the ground system. That is why you should always:

- Use a transformer with separate primary and secondary windings, even with antennas that are nominally unbalanced.
- Have a common-mode choke in the feed line near the transformer.
- Ground the feed line shield to a different ground than the antenna ground, at least 5-10 meters from the antenna ground in case of a EWE or K9AY.

Symmetric Loops

With symmetrical loops (Flag/Pennant/Diamond), which are fully balanced systems, preserving that balance is essential. For instance, do not use a metal box to house the transformer since it adds to the capacitive coupling from the loop to the feed line shield. A metal box may create imbalance unless very

Fig 7-150 — To preserve the symmetry of a Flag/ Pennant/Diamond, it should be erected as high as possible over the ground (H ≥ 2 meters). The coaxial feed line runs horizontally along the symmetry axes to a distance of at least the length of the antenna (L). Where the coax drops down to ground level, the shield should be grounded and a common-mode filter inserted beyond that point (toward the receiver).



carefully positioned. The usual operating impedance of such loops is fairly high, and any added unbalanced capacitance can upset the voltage division between the loop terminals. This allows the loop to respond more to common-mode signals, where it acts like a short "longwire" antenna. Plastic boxes are cheaper and better in this case.

Although we consider the loop as fully balanced by itself, this really is not the case because one side of the antenna is physically closer to ground. Even a perfectly balanced feed transformer (in terms of capacitive balance) is not always a guarantee for a perfectly balanced *system*, because the antenna itself is not perfectly balanced.

This is also the reason why such loops should be well in the clear and everything around them should be *as symmetrical as possible*. The feed line should run away from the feed point along one of the axes of symmetry of the loop, preferably for length equal to at least the longest dimension of the loop. This will help preserve the symmetry, but on the other end it does create a long run of "in-the-air" coax, which again is like a nice "longwire" antenna. In this case the solution is to add a few common-mode chokes along the feed line so that the longwire effect is broken up.

We should indeed do everything we can to keep common-mode currents off the outside of the coax. The cable can radiate unwanted signals directly through the air and into the antenna itself, even if you ensure low capacitance between primary and secondary of your feed transformer. In other words, coupling can occur from induction and radiation fields even if perfect transformer isolation is obtained. As W8JI says: "Just because we call it a feed line, does not mean it doesn't act like an antenna, one plate of a capacitor, or an inductor coupling through space to the antenna."

In addition to paying attention to the symmetry aspect, it is of great importance to use a transformer having the lowest possible coupling capacitance between primary and secondary. Further we should ground the coax shield to a good quality (low resistance) ground rod once the feed line has dropped down to the ground level **(Fig 7-150)**. Use a ground rod connected to nothing else.

In stubborn cases you can put a second common-mode choke at that point (beyond the ground, looking toward the receiver). Make sure the coax is buried if all possible. I suggest *not* to put a common-mode choke near the transformer. The inductive reactance of the choke could have the same magnitude as the capacitive reactance of the inter-winding capacitance of the transformer [$j\Omega L = 1/(j\Omega C)$], in which case the coil and the capacitor would be series resonant. That condition represents a short, resulting in *zero* attenuation of the common-mode signals.

3.9.3. The Transformer

We have covered transformers for Beverage antennas in great detail in Section 2.7.2. Are transformers for Receiving loops different?

The K9AY and EWE antennas are fed against ground, which means that the requirements are the same as for Beverage antennas. Their feed impedance is around 500 Ω, similar to the feed impedance of Beverage antennas. Everything that was said in Section 2.7.2. applies for these antennas as well.

The other receiving loop antennas (Flag, Pennant, Diamond) require a few more precautions. The essential characteristics of a good transformer for an elongated receiving loop antenna are:

- Lowest possible capacitive coupling between primary and secondary windings.
- Low loss because signals are very low-level.
- Good SWR is *essential* if you want to phase such antennas (you want the phase angle to be the determined by the cable length!)

Table 7-41
Inter-winding Capacitance for Various Transformers

Fair-Rite	*Turns*			
#2873000202	*1t/3t*	*2t/6t*	*3t/9t*	*4t/11t*
Single core	1.6*	3.4	4.6	8.2
Dual core	2.5*	6.2	8.3	11.8
Triple core	4	9.6	12.4	18.5

*These values yield too low a primary inductance to be used on 160 meters without increasing losses.

As the inter-winding capacitance is important, we tested a range of transformers using the binocular cores for this parameter. The results are shown in **Table 7-41**.

For the Flag, Pennant and Diamond receiving loop antenna, with an impedance of between 900 and 1000 Ω, I would recommend using a dual binocular core with 2 turns as the low-Z primary and 7 turns as the secondary (with a 75-Ω feed line impedance the secondary impedance is 919 Ω), and that should be a very good match. With a 50-Ω feed line use the same dual binocular core with 2 turns as the primary and 8 or 9 turns as the secondary.

Transformers wound on a toroidal core with separate and opposite windings are believed to achieve the lowest capacitive coupling between primary and secondary. Elaborate tests I did cannot confirm this (see also Section 2.7.2.9). With the binocular core you can achieve the same low inter-winding capacitance, and these transformers have significantly lower loss.

3.9.4. Summing Up, Feeding Elongated Receiving Loops

- Make sure your loop is physically as well balanced as possible.
- It's better to have the loop 5 meters high than 2 meters.
- Use a transformer with separate primary and secondary windings.
- Use a transformer with the lowest possible capacitive coupling between primary and secondary windings.
- Do not use a metal box to house the transformer.
- Run the feed line along the symmetry axis of the loop for several meters.
- Drop the coax down to ground, where you ground the shield to a good ground system. Insert a common-mode choke at that point (between the ground rod and the shack).

Finally, Tom, W8JI's, advice is worth seriously considering: "While you won't always see a difference with all these precautions, an ounce of prevention is worth it with low noise antennas."

3.9.5. What Else Can Go Wrong?

I have tried to cover every imaginable aspect of common-mode coupling from the feed line to a loop antenna. This kind of coupling can either make it impossible to find a good deep null off the back, or can inject a lot of trash into the antenna. But there are other ways for your loop to not function as it should.

If you cannot get a decent null, and you are not sure whether or not it's your feed line causing the problem, eliminate your feed line. Use a small battery powered receiver and a step-ladder (wooden!) and connect it to the loop with only a very short piece of coax. If you still cannot get a decent null, look for other conductors — most likely other antennas — coupling directly into your loop.

3.10. Termination Resistance

Some users of receiving loops have found it useful to be able to adjust the termination resistor value from the shack. Some use a PerkinElmer Vactrol VTL5C4 (75 Ω to 1.2 kΩ) or VTL5C2 (200 Ω to 5.5 kΩ), which is an opto-coupled variable resistance, where the resistance value is a function of the applied dc voltage. WA1ION covers this in detail on his Web site: (**www.qsl.net/wa1ion/bev/bev_remote_term.htm**).

A word of caution: Using such a remote termination with a Flag, Triangle or Pennant antenna means more chances for common-mode problems! The routing of the control cable supplying the dc voltage — and its common-mode decoupling — is once more a critical issue (see Section 3.9).

The commercial K9AY systems include such a remote controlled termination resistance (see Fig 7-148).

3.11. Decoupling the Transmit Antenna

Resonant transmit antennas in the vicinity of the loops will make them worthless. You either need to decouple the transmit antenna by inserting a high impedance (resistive or reactive) into the antenna during receive, or move the receive loop at least λ/4 from the transmit antenna. And in any of these cases it's a good idea to erect a receiving loop by aiming it directly away from the vertical.

If you cannot separate your receiving antenna that far from the resonant transmit antenna, you will have to "decouple" it. This is not only true for receiving loops but for all types of receiving antennas (see Section 2.11.1). The lower the output of such antennas, the more they will be sensitive to coupling. It has been my experience that Beverages should also be kept at least λ/4 away from a tall resonant transmit antenna.

The rest of the information on tower detuning presented in this section is reprinted with permission from W8JI's Web site. See **Fig 7-151**. Quoting W8JI:

We can minimize re-radiation by making an area or areas of the structure "electrically vanish." We often call this "detuning," even though it is more correctly electrical trapping or sectionalizing of a structure.

Most structures or towers, when detuned, have a section adjusted to represent a parallel tuned circuit. Section A and B carry out-of-phase currents. Picture the current flowing upward in A. It must then flow downward in B. Since it is a closed loop, these out-of-phase currents are equal and flow in opposite directions at resonance. The result is that radiation from sections A and B cancel each other. When section A and B are exactly resonant, sections D and C are isolated by a high impedance. The high impedance is caused by or related to the high current though the capacitor and the inductance of section A. When current is maximum, voltage drop is maximum.

This results in the electrical structure on the right, with section A and B removed. In effect, we have created a trap much like the trap in a dipole. As in the trap dipole, current is maximum in the trap at the trap's resonant frequency.

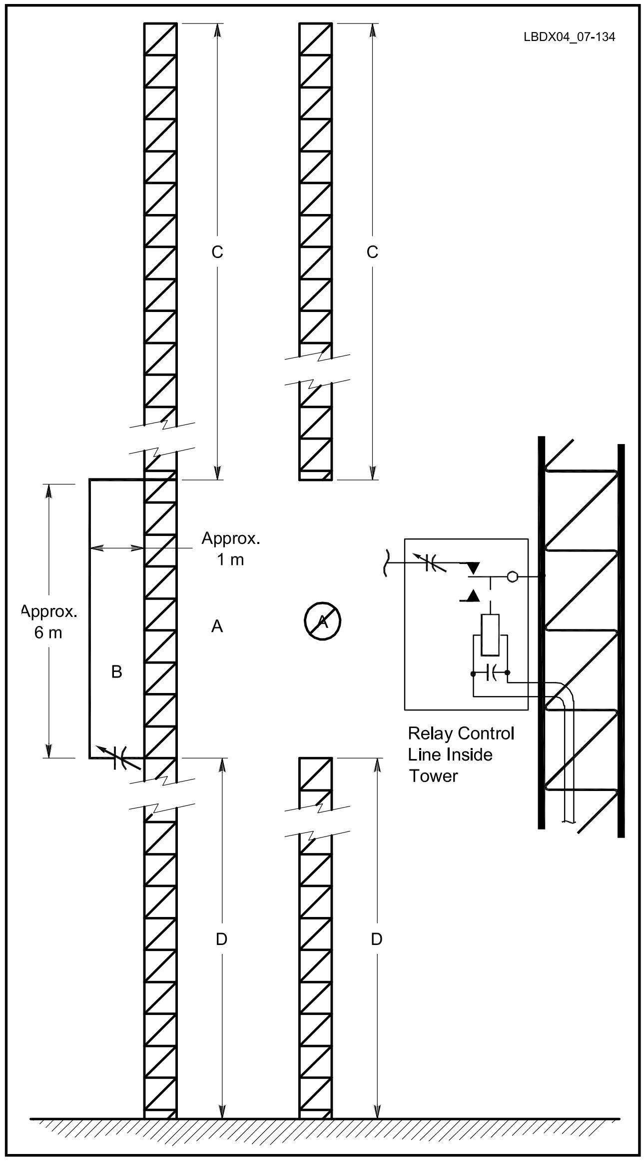


Fig 7-151 — The decoupling effect of inserting a parallel-tuned circuit in a tower. Turning Section B of the tower into a loop tuned to the offending frequency makes that section vanish altogether. See text for details.

The condition of proper tuning occurs with maximum current in section B, *not* minimum current! To electrically sectionalize the tower and isolate C and D (and minimize radiation from A) section B must be tuned for maximum current!

As either section C or D approach resonance by themselves, the tuning condition will change. This would occur when D is grounded and near λ/4 or an odd multiple of λ/4 long, or when C (with whatever is mounted on it) is self-resonant with section A removed!

Under this condition, you would either need to sectionalize and detune C or D with additional detuning, or move the location of sections A and B to a new point that (when isolated) prevents resonance in C and/or D.

A few general rules apply. Pay attention to these guidelines to ensure best results:

- Never parallel-tune a large area. Certainly not an area over 3⁄16 wavelength long.
- The detuning "loop" must have a good solid connection to the structure being detuned. Don't connect the detuning wire out to a separate object or ground stake.
- We want to adjust for *maximum* current in section B, the exception being when that would cause resonance in C or D.
- We cannot have any electrically large structures or wires hanging from the tower in the area being detuned.
- Ideally any cables passing the detuning area should be grounded to the tower at the top and bottom of the detuning area, or pass through that area in the center of the tower or mast. At the very least, cable shields should be bonded to the tower at the top and bottom of the cable run and unshielded cables placed inside the tower.
- Tuning is fairly narrow. ~5% total BW is about all that can be expected in most cases, but this varies greatly with the system including distances to the other affected antennas and the amount of pattern distortion tolerated.
- I'm surprised cables are often not grounded at the top and bottom of tall towers, and that unshielded control cables are not passed through the inside of towers. Cables should always be treated that way for lightning protection if for no other reason!

Capacitor Size

The amount of capacitance and the voltage rating of the capacitor is not easy to predict. The size depends on unwanted power levels that excite the detuned structure, the electrical characteristics of the detuned structure, and the Q of the detuning section. Capacitance values will be fairly high with short sections on lower bands like 160, for example in the range of a few thousand pF for 6-meter-long sections. The exact value would depend heavily on dimensions of the A to B loop.

Voltages across the capacitor are generally not high, although they can be at times. The "loop Q" of A and B affects voltage, as does the amount of excitation and load presented by the impedances of C and D.

MFJ sells a clamp-on calibrated current meter that will not perturb the system. It is a cheap version of a current meter I designed. This is a calibrated meter with internal amplifier that measures current from a few mA to 3 amperes, not the uncalibrated RF-sniffer commonly sold. Some RF sniffers, including those by MFJ, actually change the impedance and resonant frequency of the system because the pick-up transformers are not properly designed and terminated current transformers. Avoid loop-stick type current meters, since they measure *any* external field and can provide misleading results. Use a current meter that is directly inserted in line B, or clamps around line B with the closed core of a terminated current transformer. Use a meter that does *not* perturb the system when removed!

Lacking a current meter, it is possible to tune this system with a grid dip meter, by forming a small one or two turn coupling loop. As an alternative, the loop can be broken at any point near the capacitor and an MFJ-259 or similar antenna analyzer connected in series. Proper adjustment is at the point where minimum impedance occurs. If that impedance is not low, you probably are not effectively detuning the structure.

Multiple Stacked Antennas or Tall Structures

When multiple stacked antennas are used, especially on a fairly tall tower, it may be necessary to sectionalize multiple points. Individual sections between antennas can be resonant, or appear electrically long.

If the tower or structure or any part of the structure or tower becomes resonant when section A is tuned to present a high impedance, then we need to move section A or tune it to some condition other than maximum current (resonance). Adjustments under this condition can only be made two ways:

- A sampling loop can be mounted on the structure 1⁄10 wavelength or more above or below section B and adjusted for minimum terminal voltage
- Field strength of the pattern can be plotted, and the structure tuned for minimum pattern distortion

Never detune an area that contains large Yagis or other electrically large objects, like long conductive guy lines, dipoles, or cables leaving the tower.

(Thanks for the information, Tom.)

3.11.1. Switching the Decoupling Section During Transmit

If you are decoupling your transmitting antenna, you will have to open-circuit the decoupling loop during transmit. If you use high power, a small vacuum relay is a good idea. Do not use a bulky slow relay, as you will need it to switch quickly, in pace with your amplifier keying line. Run the coil voltage through a cable on the inside of the tower and decouple it well at the base of the antenna. I use a ferrite rod with 20 turns of the small control cable on the rod. I added some good quality 0.01 µF decoupling caps to ground too.

The relay should be switched from the line that switches your amplifier — that means that the relay energizing voltage should come on approximately 10 to 15 ms (depending how fast your relay is) before RF appears. The same rules that apply for switching amplifiers apply. Using this system full break QSK is not advised. If you do want to use QSK, you will probably have to switch off the trap system.

3.11.2. Gamma-Matched Tower

If you have a gamma matched tower, you may consider just turning the section where the gamma-match system is installed into a resonant loop, this way decoupling everything that's above the gamma attachment point from ground. This way you only have to make a little switchbox located at the box containing the gamma series capacitor (see **Fig 7-152**).

If the tower is approximately 90° long the gamma-match attachment point will be approximately 8 meters high. Electrically longer and shorter gamma-matched towers will require a longer gamma match (see Chapter 9). Use a small and fast vacuum relay to switch the gamma rod to the feed line or the loop. The return of the loop at ground level should *not* be done through the ground. A heavy conductor or metal (aluminum) pipe is required.

At my location it was necessary to detune my 160-meter quarter-wave vertical to minimize coupling to my small receiving antenna, which is located less than $\lambda/4$ from the transmit antenna. I followed the tuning procedure described by W8JI above using an MFJ antenna analyzer. I tuned the loop with a four-gang BC variable (2000 pF), which I later replaced with paralleled ceramic transmit-type capacitors.

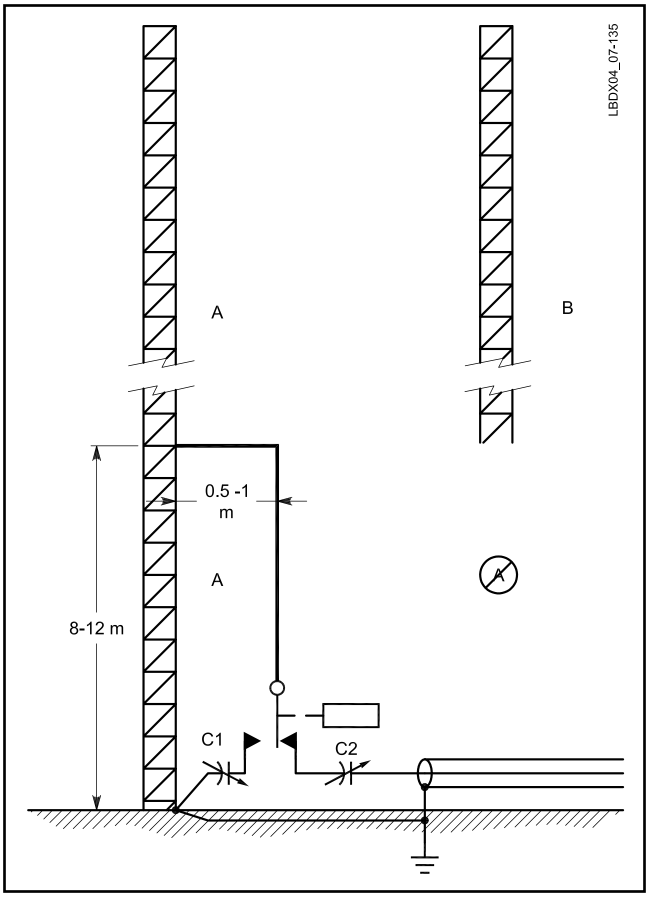


Fig 7-152 — Detuning a grounded gamma-matched tower. C1 resonates the loop; C2 tunes out the inductive reactance of the gamma rod. See text for details.

3.11.3. Detuning Base-Insulated Towers

I described methods for detuning insulated towers in Section 2.11.1.

3.12. Arrays of Loops

Elongated terminated nonresonant loops make excellent candidates for wideband arrays, with elements either fed in-phase or in an end-fire configuration, using the crossfire phasing method. These are broadband, high-loss antennas. But the effects of mutual coupling are hardly visible on their feed-point impedances, similar to arrays of Beverages. Receiving loops can be used as elements in any of the array configurations described for verticals or Beverages.

End-fire

Putting two Flags behind one another and feeding them

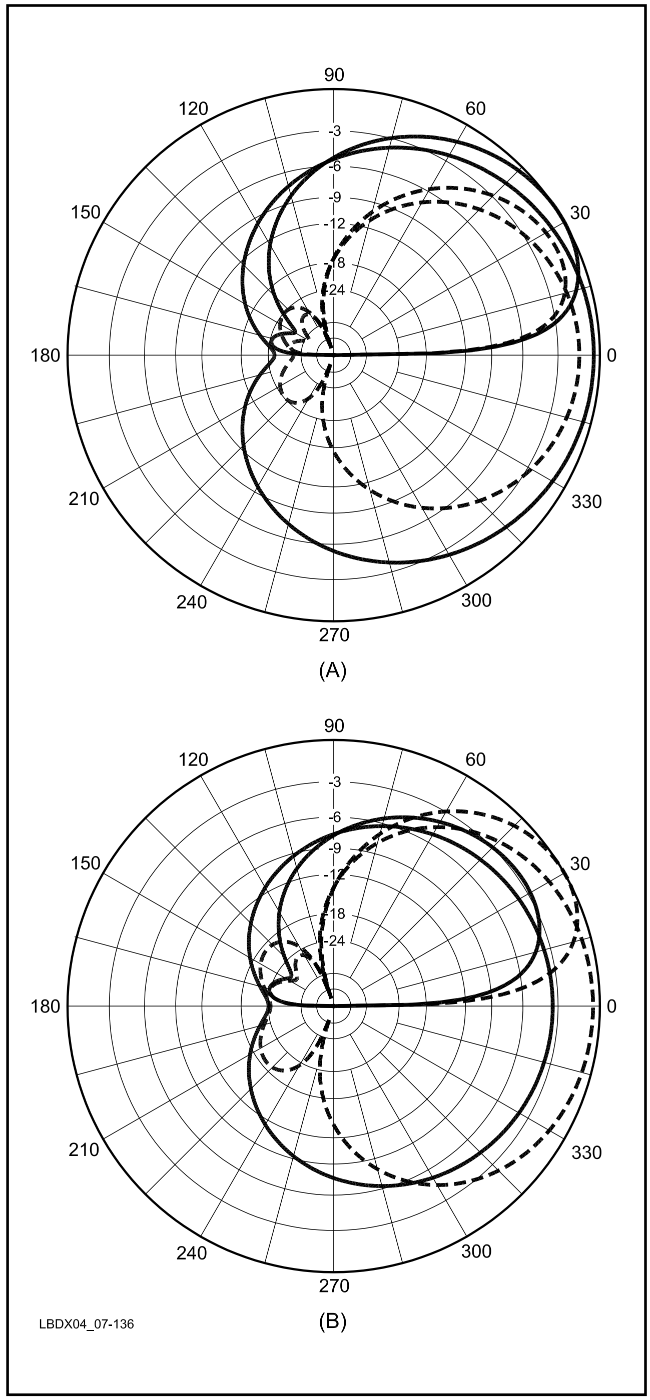


Fig 7-153 — End-fire phased Flags with 20 meter center-to-center spacing. At A, a single Flag on 160 meters (solid line); phased Flags (ψ = 160°, dashed line). At B, a single Flag on 80 meters (solid line); phased Flags (ψ = 140°, dashed line).

Table 7-42
Directivity Figures of Single and End-Fire Phased Flags

	Single Loop 160	*Phased Loops 160*	*Single Loop 80*	*Phased Loops 80*
Gain (dBi)	–29.0	–30.4	–18.5	–16.0
DMF (dB)	11.4	21.6	10.7	19.3
RDF (dB)	7.4	10.0	7.0	9.1
–3-dB Angle	147.0°	98.0°	156.0°	113.0°

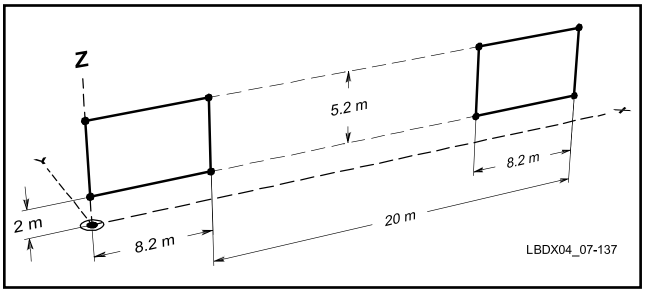


Fig7-154 — Layout of the 2-element end-fire Flag array.

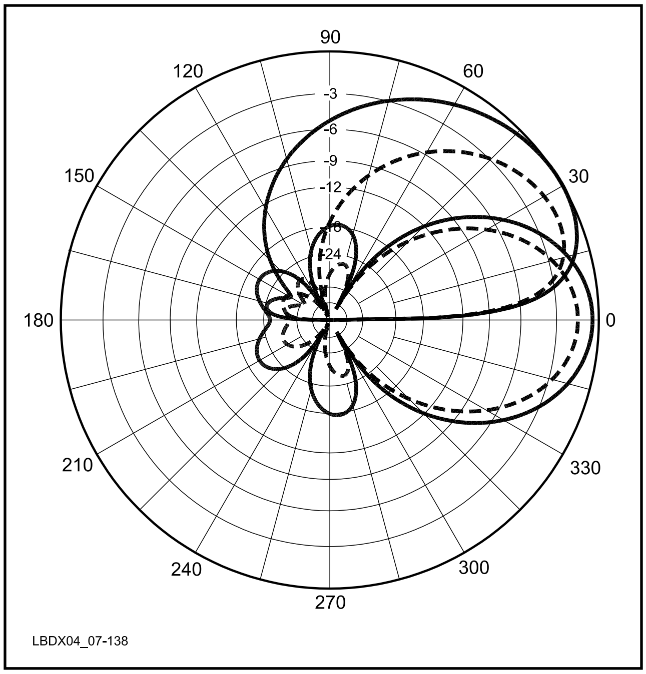


Fig 7-155 — Broadside array of two Flags, separated 0.61 λ (solid lines) compared to broadside array (0.61 λ lateral spacing) of two Flags cells, each cell containing two Flags (dashed lines).

end-fire with the appropriate phasing angle (ψ, see Table 7-1) can do wonders for directivity patterns (see **Figs 7-153** and **7-154**). These Flags, spaced 20 meters (43.4° on 160 meters), were fed with ψ = 150° on 160 meters and ψ = 130° on 80 meters. **Table 7-42** shows directivity figures of phased loop arrays compared to single loops. Note that the DMF jumps approximately 10 dB by going from a single loop to two end-fire phased loops, which is quite a spectacular change. The RDF increases by 2 to 2.5 dB, a respectable figure as well, since most of the improvements are in the back of the array. Note also the substantial reduction in 3-dB beamwidth and also in high-angle because of the close spacing and the high phasing angle. This high phasing angle, of course, results in a relative loss of the array vs a single element, instead of what you might expect in terms of gain for an array. If you phase the system for gain, you would use a much smaller phasing angle, but would reap very little directivity improvement. Again: there is no free lunch!

This array can be fed using the crossfire principle as shown in Fig 7-132 for two Beverages. Here too you should take care that the SWR on the phasing lines is kept to less than 1.1:1. This can easily be achieved by adjusting the turns-ratio of the matching transformer.

Broadside: Floyd Koontz, WA2WVL, described arrays of EWEs in one of his *QST* articles (Ref 1264). Two EWEs in broadside, spaced approximately λ/2, give the lobe-narrowing effect also seen with arrays using verticals and Beverages. There is no improvement in either back-lobe or high-angle behavior. We know from Section 1.11 that ~0.67 λ spacing results in much better directivity (RDF), since it lifts the elevation pattern off the ground. The pattern produced by two broadside Flags, spaced 0.61 λ is shown in **Fig 7-155.**

Broadside/End-fire: The ultimate configuration is the four-element array, being a combination of two side-by-side (in-phase) end-fire cells in an in-phase broadside array (similar to the array with vertical elements in Section 1.12 and the array of Beverages in Section 2.16.4). In this case the improvement from the end-fire cell is combined with the narrowing effect of the broadside combination, yielding very good directivity figures (see Fig 7-155 and **Table 7-43**).

Of course, once you start envisioning broadside arrays, you need a lot of room (100 meters spacing or more). We already know the disadvantages of a receiving antenna with a very narrow –3 dB forward beam angle: Where should you aim it? We should not forget that on the low bands signals often deviate 30° to 45° (and even more sometimes) from the theoretical great-circle direction. Of course, you can put up 12 such arrays. Did I start out saying that loops are receiving antennas suitable for use on small properties? I think these giant arrays are more for modeling fun than anything else!

4. RANKING RECEIVING ANTENNAS

The performance data of receiving arrays using small vertical antennas and the results for Beverage antennas were listed in Tables 7-21 and 7-39.

If you compare the family of elongated terminated loops such as the EWE, Flag, K9AY, etc (**Table 7-44**) with the

Table 7-43
End-Fire and Broadside Combinations of Receiving Loops

	2 End-Fire Loops	*2 Broadside Loops*	*4 End-Fire/ Broadside Loops*	*4 End-Fire/ Broadside Verticals Sec. 1.12*	*4 End-Fed/ Broadside Beverages Sec. 2.16.4*
DMF (dB)	21.5	16.4	26.6	19.5	34.7
RDF (dB)	9.9	10.4	12.3	12.7	14.1
–3-dB Angle	98.0°	55.0°	50.0°	46.0°	34.0°

Table 7-44
Performance Data For Receiving Loops (1.8 MHz, AVG Ground)

Loops and Arrays of Loops	*DMF (dB)*	*RDF (dB)*	*3 dB Angle (degrees)*	*Output (dBi)*	*Reference*
Elongated terminated loop (EWE, Flag, K9AY etc)	~11	7.5	~140	–29	Sect 3
Double half-delta loop (AA7JV)	—	9.3	~100	–29	Sect 3.8
2-element end-fire array of loops	21.6	10.0	89	–30	Sect 3.11

abovementioned families of receiving antennas, this family is right at the bottom of the performance list, showing the least degree of directivity.

But they are also so much better than having to listen on your vertical transmitting antenna. Ask anyone who has such a loop and who's never had a "big" receiving antenna. The merit of such receiving loops is a small footprint and their relative simplicity.

A remarkable performer in this family is AA7JV's double half-delta loop (see Section 3.8), which does significantly better than the "standard" elongated terminated loops.

5. SUMMING IT UP ON SPECIAL RECEIVING ANTENNAS

When I analyzed the results of the poll I ran among over 400 enthusiastic low-band DXers, I was amazed to see some of the top scorers did *not* use Beverages or other special receiving arrays. After studying the data further, I found that those were the guys using directive transmit antennas (Yagis or phased arrays) on 80 meters and — yes — now even on 160 meters.

Happy are the very few who categorically say: "I don't need a special receiving antenna. My transmit antenna is better than Beverages, even phased ones." I only know a few stations that can say that, for example K9DX and N7JW/K7CA on 160 meters.

I have received many requests to write about small receiving antennas that would out-hear W8JI's Beverage arrays and Eight Circle or K9DX's Nine Circle, just to name a few top stations with top acreage. We should all know there is no free lunch in this cruel radio world! You now know the directivity figures (DMF and RDF) of the elongated loop antennas (Flags and such). That's it. You know you can do better with a Four Square receive array, but it takes a *lot* more engineering and building effort than to put up a Flag. Again, there is no free lunch. Even a 2-element end-fire array with vertical elements is substantially better than the elongated loops, because they have no high-angle radiation, while the loops have high-angle radiation from the horizontal or sloping wires. Clearly this is an incentive to build something better than such receiving loops. A 2-element end-fire pair does not have to be big. Six-meter tall elements (top loaded) and a spacing of 10-15 meters can set you on the road.

And, of course, you never can have too many good receiving antennas. Tom, W8JI, and Wally, W8LRL, are the proverbial proofs of the pudding. K4ISV's statement "Can you imagine a fisherman going out with only one bait?" makes a lot of sense. But I would immediately like to add that if you choose to have a bunch of antennas, make sure they do not couple with one another or you will have a totally uncontrolled condition where anything can happen.

In Chapter 1 on propagation I mentioned a fairly typical phenomenon of high-angle propagation before sunset or after sunrise. Under such circumstances a low dipole will outperform a long Beverage because of its angle of radiation. A low dipole is definitely also useful in your gallery of receiving weapons. As Frank Donovan, W3LPL, has said: "You can never have too many antennas." I would like to add: "on condition they are well spaced and don't influence one another in an uncontrollable way…"

6. PREAMPLIFIERS

All special receiving antennas described in this book are low-gain antennas. Antennas with a nominal gain of as low as –15 dBi will normally not require a preamplifier, unless you live in a very quiet rural area and have very long, lossy feed lines. That means that Beverages, even with feed lines of many hundreds of meters long, will — as a rule — not require a preamplifier. Unless of course you like to see your S-meter dancing up and down like a yoyo. But signal readability has nothing to do with dancing S-meter needles. It is only a question of signal-to-noise ratio.

But we have also seen receiving antennas such as loops with gains of –30 dBi, and that is pretty low and does require some signal boosting. How do you know if you need a preamplifier? The rule is simple. During the quietest moment of the day (usually at noon on 160 meters), with your receiver set at the narrowest bandwidth you normally use, can you hear the noise go up significantly when you switch from a dummy load to your receiving antenna? If you can, then you have enough gain in your system.

If you hear the receiver's internal noise and not band noise, then you need a preamplifier. You have a problem because:

- The antenna output might be extremely low (less than –15 dBi, for example).
- The feed line might be very long and/or lossy.
- You need to compensate for filter losses, splitter losses etc.

As a rule you should keep the signal level as low as possible to prevent chances of intermodulation and overload. This is also why our receivers have a front-end-attenuator, as well as a switchable preamplifier.

In most cases you can put the preamplifier in the shack. The signal loss in the feed line is a loss that affects both the signal and external noise. That means that the loss in the feed line does not affect the S/N ratio. In most circumstances moving the preamplifier to the antenna will not affect the system unless the feed line loss is so high that the noise floor is indeed established by the preamplifier and not the antenna.

All my Beverages are fed with ⅝-inch or ½ inch Hardline so the line losses are negligible, even though some of my feed lines are more than 300 meters long. My small receiving

Four Square is very close to the house, so the coaxial losses using RG-6 are very low. I have used both a K9AY duo-band 80 and 160-meter preamplifier (model PRE1, no longer available) and the DX Engineering RPA-1 push-pull preamplifier, both of which I can highly recommend. The small receiving Four Square requires 10 dB of preamplification, but my Beverages never require preamplification. Maybe that only proves that I do not have a super quiet QTH, and unfortunately that is true…

Untuned preamplifiers using MMICs are not the right choice. MMICs have poor IMD performance and are also very poor with even-order harmonic distortion. They have a fixed gain, which is usually way too high anyhow. Last but not least, they withstand little abuse.

Larry, W7IUV, has published a 160-meter preamplifier design (**Fig 7-156**) with excellent characteristics. Go to his Web site (**w7iuv.com**) for further details on this very popular amplifier.

Sergio, IK4AUY, described an excellent preamplifier as part of his article "A High Level Accessory Front End for the HF Amateur Bands" in Apr/March 2003 *QEX* (Ref 1267). The schematic of his preamplifier is shown in **Fig 7-157**. The performance data are just short of spectacular:

- 3rd order intercept (IP3): +43 dBm
- 1 dB compression: +24.5 dBm
- Gain: 12 dB (1 to 30 MHz)
- Noise figure: 3.9 dB at 30 MHz

T1 and T2 are each 12 trifilar (twisted) turns of 0.35 mm (#27 AWG) enameled wire on a FT50-43 ferrite toroidal core. T3 and T4 are nine turns primary and two turns secondary (same wire), wound on a smaller FT37-43 core. If a higher gain than 12 dB is required, you can easily connect two such preamplifiers in cascade.

Each transistor is biased for around 60 mA, so the total current drain is 120 mA at 13.5 V dc. All details on this preamplifier were graciously provided by Sergio Cartoceti, IK4AUY. See also **Fig 7-158** and **www.qsl.net/ik4auy/**.

Many commercial-grade preamplifiers are available, some better than others. Some popular models behave very badly in the presence of strong signals. The ICE single-band preamplifiers (**www.iceradioproducts.com**) work well. I also use the DX Engineering model RPA-1. This broadband amplifier, designed by W8JI and shown in **Fig 7-159**, has outstanding characteristics:

- Gain: 17 dB
- 1 dB compression: at +26 dBm (400 milliwatts output)

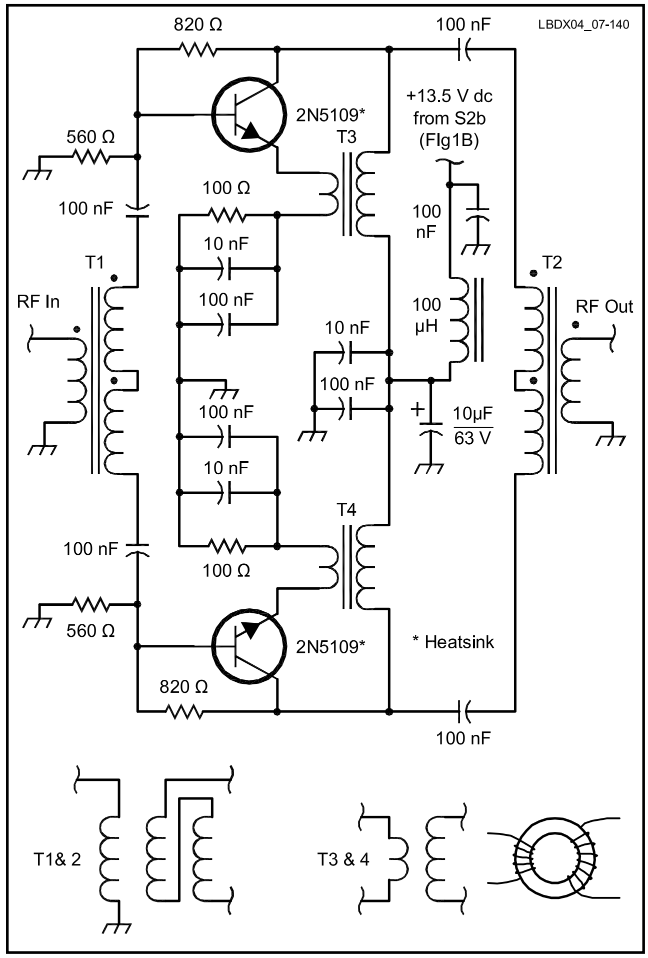


Fig 7-157 — High-performance 12-dB gain amplifier with excellent strong-signal handling characteristics, originally developed by Ulrich Rohde, N1UL, and modified by Sergio Cartoceti, IK4AUY. T1 and T2 are each 12 trifilar (twisted) turns of 0.35-mm (AWG #27) enameled wire on an FT-50-43 ferrite toroid core. T3 and T4 are each a 9-turn primary and a 2-turn secondary (0.35-mm AWG #27 enameled wire) on an FT37-43 core.

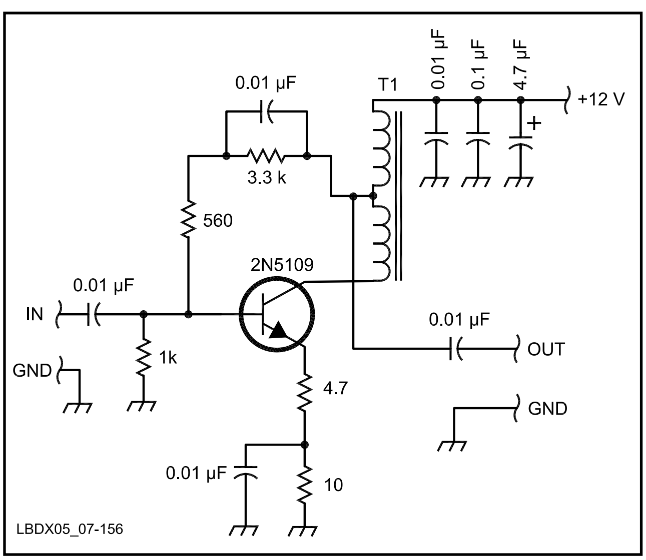


Fig 7-156 — The March 2009 version of the popular W7IUV preamp.

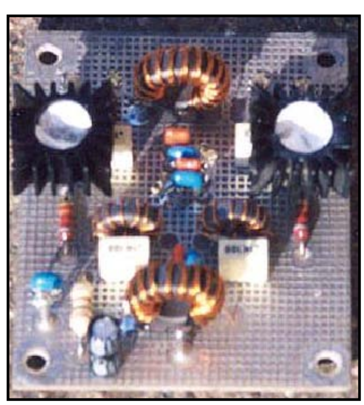


Fig 7-158 — The IK4AUY preamp, modified from the design by Ulrich Rohde, N1UL.

Table 7-45
Preamplifier Performance Characteristics

Preamplifier	*Output IP3 (dBm)*	*Gain (dB)*	*Input IP3 (dBm)*	*NF (dB)*	*RF* (dB)*
W7IUV	39	18	21	5**	16**
IK4AUY	43	12	31	3.9	27.1
Advanced Receiver Research	30	20	10	5.4	4.6
DX Engineering (RPA-1)	43	17	26	3.4	22.6
ICE 124A	14	16	–2	5**	–7**

*W7ZOI Receiver Factor = Input IP3 – NF
**estimated value, not measured

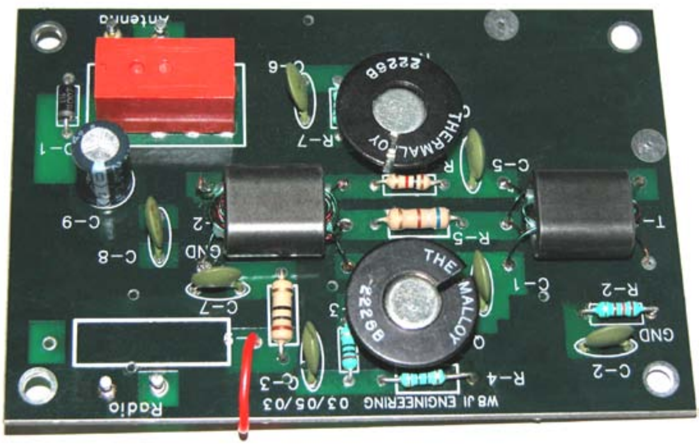


Fig 7-159 — The DX Engineering push-pull amplifier designed by Tom Rauch, W8JI.

- 3rd order intercept point: +43 dB
- Noise figure: 3.4 dB

For more details: **www.dxengineering.com**. Tom, W8JI, has published measurement data on his Web site: **www.w8ji.com/pre-amplifiers.htm**.

6.1. How to Assess the Relative Quality of Preamplifiers

Wes Hayward, W7ZOI, introduced a figure of merit for receiver systems, which he called the Receiver Factor (RF) with bandwidth invariant parameters:

RF = IIP3 – NF

where
IIP3 = 3rd order input intercept (dBm)
NF = noise figure in dB

These are cumulative values for a system. The first is a measure of strong-signal performance, while the other defines weak signal behavior.

If the input intercept of an amplifier is known the intermodulation distortion is well defined for all input levels. We measure IMD at the output and if not otherwise specified IP3 for amplifiers is referenced to the output (OIP3).The difference is the simple stage gain:

IIP3 = OIP3 – G

where
IIP3 = 3rd order input intercept (dBm)
OIP3 = 3rd order output intercept (dBm)
G = gain in dB.

Many years ago, Anzac engineers called this Amplifier Factor (IIP3 – NF) for a single-amplifier stage (the higher the number, the better, of course). If we apply this to the above-mentioned preamplifiers we get **Table 7-45**.

6.2. Protecting the Preamp Input During Transmit

When the transmit antenna and the receive antenna are close spaced, detuning the transmit antenna will help isolate the transmit antenna from causing pattern changes to the receive antenna, but it will not help to eliminate the high signal levels the preamp will see during transmitting. The decoupling factor for close spaced transmitter and receive antennas can be quite low (30-35 dB) which places nearly a watt or more of power into the input of the receive preamp when transmitting 1500 W of power. That number is often 20 dB or more than the preamp's 1 dB compression point when referenced to the input of the amplifier. The simple reed relay configuration of Fig 3-17 in Chapter 3 will ensure the preamp is not damaged.

Contents

CHAPTER 8

The Dipole Antenna

Klaus Owenier, DJ4AX, is an all-around ham. Klaus recently retired as a teacher in electronics and electromagnetics at the Ruhr-University Bochum. He was one of the first members of the world-famous Rhein-Ruhr DX Association (RRDXA), worldwide winner of the CQ World Wide DX Contest club competition for many years in a row. Klaus is an antenna expert and an excellent contester and CW operator. He has been a valuable and consistent presence during CQ Worldwide contests at OT*T for many years. Also for this 5th edition, my friend Klaus was found ready and willing to be my guide, counselor and helping hand. His critical analyses on dipole antennas have been very instrumental in the reworking of this chapter. Thank you, Klaus.

The first antenna most amateurs encounter is a dipole. I remember how, as a young boy, I put up my first 20-meter dipole between a second-floor window of our house and a nearby structure. It was fed with 75-Ω TV coax, and it worked — whatever that meant. For a while (almost 50 years ago!) my whole antenna world was limited to a dipole. But there is more to dipoles.

Although we often think of dipoles as ½ λ long, center-fed antennas, this is not always the case. The definition used in this chapter is that of a center-fed radiator with a symmetrical sinusoidal standing-wave current distribution.

1. HORIZONTAL HALF-WAVE DIPOLE

1.1. Radiation Patterns of the Half-Wave Dipole in Free Space

The radiation pattern in the plane of the wire has the shape of a figure 8. The pattern in the plane perpendicular to the wire is a circle (see **Fig 8-1**). The three-dimensional representation of the radiation pattern is shown in the same figure and is a ring (torus). In free space the gain of this dipole over an isotropic radiator is 2.14 dB. This means that the dipole, at the tip of the ring where the radiation is maximum, has a gain of 2.14 dB compared to the theoretical isotropic antenna, which radiates equally well in all directions (its radiation pattern is a sphere).

1.2. The Half-Wave Dipole Over Ground

In any antenna system, the ground acts like an imperfect or lossy mirror that reflects energy. Assuming a perfect ground to simplify matters, we can apply the Fresnel reflection law, where the angles of incident and reflected rays are identical.

1.2.1. Vertical Radiation Pattern of the Horizontal Dipole

The vertical radiation pattern determines the *wave angle* of the antenna; the wave angle is the angle at which the radiation

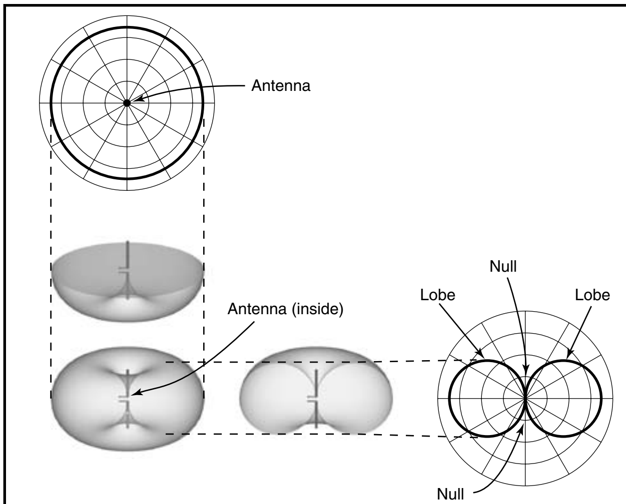


Fig 8-1 — Radiation pattern as developed from the three-dimensional pattern of a half-wave dipole in free space. Upper left, vertical-plane pattern, and right, horizontal pattern.

is maximum. Since obtaining a low angle of radiation is one of the main considerations when building low-band antennas, we will usually consider only the lowest lobe in case the antenna produces more than one vertical lobe. In free space, the radiation pattern of the isotropic antenna is a sphere.

As a consequence, any plane pattern of the isotropic antenna in free space is a circle. In free space, the pattern of a dipole in a plane perpendicular to the antenna wire is also a circle. Therefore, if we analyze the vertical radiation pattern of the horizontal dipole over ground, its behavior is similar to an isotropic radiator over ground.

1.2.1.1. Ray Analysis

Refer to **Fig 8-2**. In the vertical plane (perpendicular to the ground), an isotropic radiator radiates equal energy in all directions (by definition). Let us now examine a few typical rays. A and A' radiate in opposite directions. A' is reflected by the ground (A) in the same direction as A. B", the reflected ray of B', is reflected in the same direction as B.

The important issue is the phase difference between A and A", B and B", etc. Phase difference is created by path-length difference (length is directly proportional to time, since the speed of propagation is constant), plus any phase shift at the reflection point itself. Horizontally polarized rays undergo a 180° phase shift when reflected from perfect ground. This can be simulated by feeding an image antenna with I' = – I (see Fig 8-2).

If at a very distant point (in terms of wavelengths) the rays at points A and A" are in phase, then their combined field strength will be at a maximum and will be equal to the sum of the magnitudes of the two rays. If they are out-of-phase, the resulting field strength will be less than the sum of the individual rays. If A and A" are identical in magnitude and 180° out-of-phase, total cancellation will occur.

If the dipole antenna is at a very low height (less than ¼ λ), A and A" will reinforce each other. Low-angle rays will be almost completely out-of-phase, resulting in cancellation, and thus there will be very little radiation at low angles. At increased heights, A and A" may be 180° out-of-phase (no radiation at zenith angle), and lower angles may reinforce each other. In other words, the vertical radiation pattern of a dipole depends on the height of the antenna above the ground.

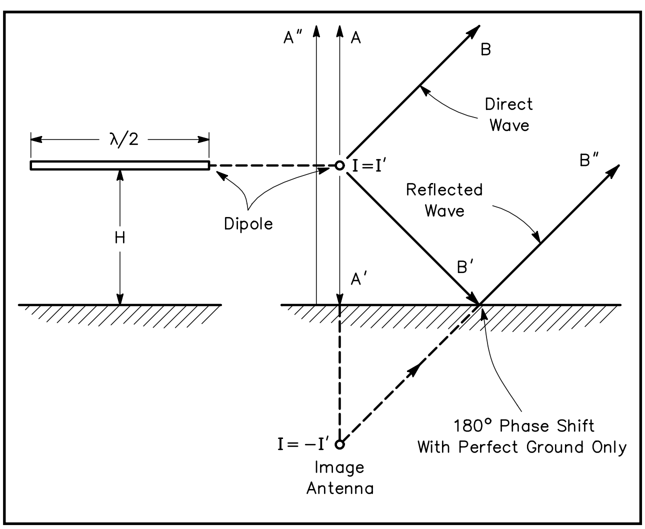


Fig 8-2 — Reflection of RF energy by the electrical "ground mirror." The eventual phase relationship between the direct and the reflected horizontally polarized wave will depend primarily on the height of the dipole over the reflection ground, and to a small degree on the quality of the reflecting ground (as an electrical conductor).

1.2.1.2. Vertical Radiation Pattern Equations

The radiation pattern can be calculated with the following equation.

$$F_{\alpha} = \sin (h \sin \alpha) \qquad \text{(Eq 8-1)}$$

where

F_{α} = normalized field intensity at vertical angle α

h = height of antenna in degrees

α = vertical angle of radiation

One wavelength equals 360°. Eq 8-1 is valid only for perfectly reflecting grounds. For real ground, the reflected wave must be multiplied by the complex reflection coefficient. This is shown in **Fig 8-3;** its total phase difference is then >180°, its magnitude <1.

In special cases Eq 8-1 has simple solutions:

1) For certain lobes we have F_{α} = 1, which occurs when (h sin α) = 90°, 270°, 450°, etc. These represent the first, second, third lobe etc. For the first lobe Eq 8-1 can be rewritten as:

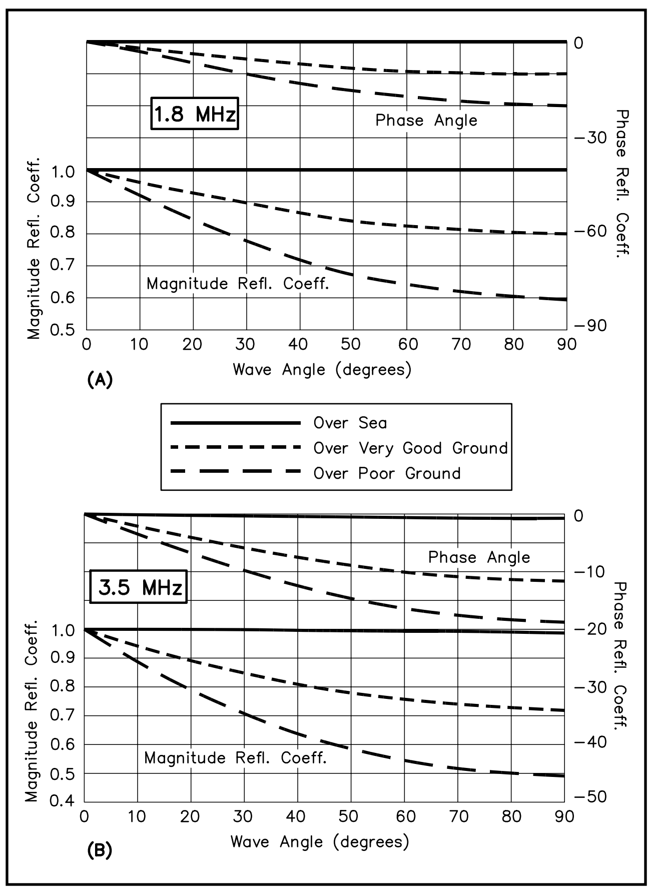


Fig 8-3 — Reflection coefficient (magnitude and phase angle) of horizontally polarized waves over three types of ground: saltwater, average and very poor. See text for details.

$$H_1 = \frac{74.95}{f \sin \alpha} \qquad \text{(Eq 8-2)}$$

where

H_1 = height of antenna in meters

f = frequency, MHz

α = vertical angle for which the antenna height is sought.

2) For *nulls* we have $F_\alpha = 0$ when (f sin α) = 180°, 360° etc. These represent the first null, second null…, respectively. For the first null Eq 8-1 can be rewritten as:

$$H_1 = \frac{149.9}{f \sin \alpha}$$

Table 8-1 gives the major-lobe angles as well as reflection-point distances for heights ranging from 18 meters (60 feet) to 60 meters (200 feet) for 40, 80 and 160 meters.

1.2.1.3. Sloping Ground Locations

In many cases, an antenna cannot be erected above perfectly flat ground. A ground slope (Ref 630) can greatly influence the wave angle of the antenna. *HFTA* (*High Frequency Terrain Assessment*) from N6BV calculates the radiation pattern of dipoles (or Yagis) as a function of the terrain slope. (HFTA is included on the CD with recent editions of the *ARRL Antenna Book*.)

Table 8-2 shows the influence of the slope angle on the required antenna height for a given wave angle on 80 meters. The table lists the required antenna height and the distance to the reflection point for a horizontally polarized antenna. A positive slope angle is an uphill slope. The results from this table can easily be extrapolated to 40 or 160 meters.

1.2.1.4. Antennas Over Real Ground

Up to this point, a perfect ground has been assumed for most of the results presented. Perfect ground does not exist in practical installations, however. Perfect ground conditions are approached only when an antenna is erected over salt water.

Radiation efficiency and reflection efficiency

Contrary to the case with vertical antennas, a horizontal antenna does not rely on the ground to provide a return path for antenna currents. The physical "other half" takes care of that. This means that the ground will practically not play an important role in the radiation efficiency of the antenna. The radiation efficiency is related mainly to the losses in the antenna itself (conductor, insulator, loading coils, etc), although of course some of the total radiated energy can be dissipated in ground losses.

Table 8–1
Major Lobe Angles and Reflection Point Distances for Various Dipole Antenna Heights

Antenna Height		*40 Meters*			*80 Meters*			*160 Meters*		
		Angle	*Distance*		*Angle*	*Distance*		*Angle*	*Distance*	
(ft)	*(m)*	*(deg)*	*(ft)*	*(m)*	*(deg)*	*(ft)*	*(m)*	*(deg)*	*(ft)*	*(m)*
60	18	36	83	25	90	0	0	90	0	0
80	24	26	163	50	54	58	18	90	0	0
100	30	20	266	81	40	118	36	90	0	0
120	36	17	391	119	33	187	57	90	0	0
140	42	15	540	148	28	268	82	77	31	9
160	48	13	710	217	24	362	110	59	97	30
180	54	–	–	–	21	467	142	49	154	47
200	60	–	–	–	18	584	178	43	213	66

Table 8-2
Slope Angle Versus Antenna Height at 3.5 MHz

Slope Angle	*20° Wave Angle*		*30° Wave Angle*		*40° Wave Angle*	
	Height	*Distance*	*Height*	*Distance*	*Height*	*Distance*
(deg)	*(ft)*	*(ft)*	*(ft)*	*(ft)*	*(ft)*	*(ft)*
35	–	–	–	–	906	10,364
30	–	–	–	–	430	2,441
25	–	–	819	9,367	275	1,029
20	–	–	396	2,249	201	553
15	768	8,789	258	966	158	340
10	378	2,146	192	528	131	227
5	251	937	153	329	113	161
0	189	520	129	224	100	120
–5	153	329	113	161	91	91
–10	131	227	102	121	85	72
–15	116	166	94	94	81	57
–20	107	127	89	75	79	45
–25	101	101	87	61	78	36
–30	97	91	86	49	78	28

Horizontally and vertically polarized antennas both rely on the ground for reflection of the RF in the so-called Fresnel zone to build up the radiation pattern in combination with the direct wave, as shown in Fig 8-2. The efficiency of the reflection depends on the quality of the ground, and is called the *reflection efficiency*.

Reflection coefficient

The reflection from real ground is not like on a perfect mirror. The reflection coefficient is a complex number that describes the reflection from real ground:

- With a perfect mirror, all energy is reflected. There are no losses; the reflection coefficient magnitude is 1.
- With a perfect mirror, the phase of the reflected horizontal wave is shifted exactly 180° compared to the incoming wave.
- With real ground, part of the RF is absorbed, and the reflection coefficient magnitude is less than 1.
- With real ground, the phase angle of the reflection coefficient is greater than 180°. Except when the antenna wave angle is quite high, the deviation from 180° is very small. This deviation typically varies between 0° and 25° for reflection angles (equal to wave angles) between 0° and 90°.
- The magnitude of the reflection coefficient, which becomes smaller as the ground quality becomes poorer, is the reason that the dipole over real ground shows less gain than over perfect ground.

The reflection coefficient is a function of the wave angle. The smaller the wave angle, the closer the reflection coefficient magnitude will approach 1. This explains why the loss with a dipole (poor ground vs perfect ground) is higher at high angles (for example, at the zenith) than at low angles. See **Fig 8-4**.

The fact that the dipole over poor ground seems to have a lower radiation angle than over perfect ground is because at lower angles there is less loss. In other words, over poor ground it just has less loss at low angles than at high angles.

The filling in of the deep notch at a 90° wave angle for the dipole at ½ λ (and 1 λ) (Fig 8-4B and D) is because the reflected wave is considerably attenuated and phase shifted and can no longer cancel the direct wave. Note that changing the height of the antenna could compensate for the effect of the additional phase shift.

Again refer to Fig 8-3 showing the reflection coefficient (magnitude and phase) for a horizontally polarized wave. The information is for a horizontally polarized antenna over saltwater, average ground and very poor ground, for both 160 and 80 meters.

Radiation patterns

Fig 8-4 shows vertical patterns of a horizontal half-wave dipole over both near-perfect ground (salt water) and desert, the two extremes. **Table 8-3** lists the wave angle and the relative loss for a half-wave dipole over five different types of ground and for two antenna heights. Note that for a dipole at ½ λ, the peak wave angle drops from 30° over seawater to 26° over desert. At the same time there is a radiation loss of 1.21 dB.

For an antenna at ¼ λ height (Fig 8-4A), maximum radiation occurs at 90° over a perfect conductor. Over very poor ground (desert), the maximum radiation is at 59°. This is not because more RF is concentrated at this lower angle, but only because more RF is being dissipated in the poor ground at the 90° angle than at 59° (the reflection coefficient is much lower at 90° than at 59°). The difference, however, between the radia-

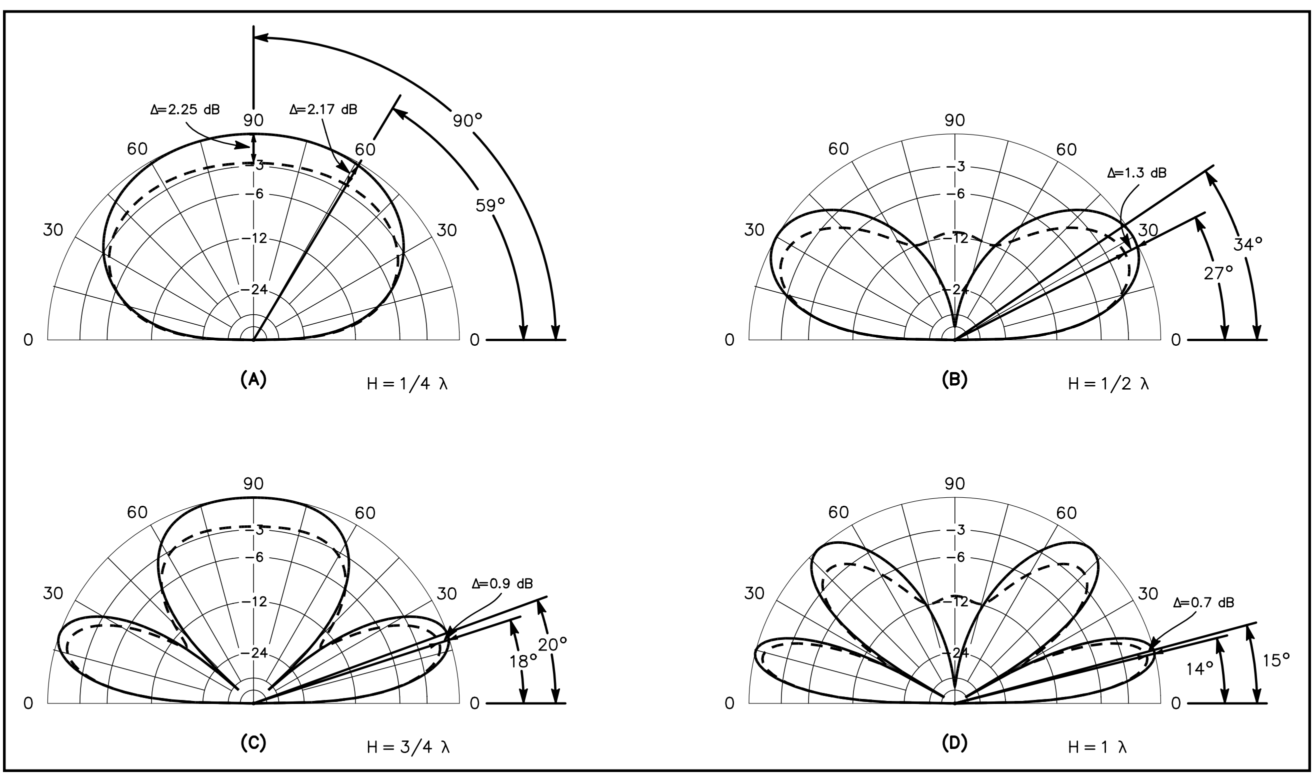


Fig 8-4 — Vertical radiation patterns over two types of earth: saltwater (solid line in each set of plots) and very poor ground (broken line in each sets of plots). The wave angles as well as the gain difference between saltwater and poor ground are given for four antenna heights.

Table 8-3

Relative gain (vs dipole over perfect ground) and wave angle (max vertical radiation angle) for ½-λ dipoles at heights of ¼ λ and ½ λ.

	Height= ¼ λ		*Height = ½ λ*	
	Rel. Loss (dB)	*Wave Angle (deg)*	*Rel. Loss (dB)*	*Wave Angle (deg)*
Perfect Ground	0	90	0	30
Saltwater	–0.05	90	–0.01	30
Very Good Ground	–0.57	71	–0.16	29
Average Ground	–1.23	62	–0.52	28
Very Poor Ground	–2.17	53	–1.21	26

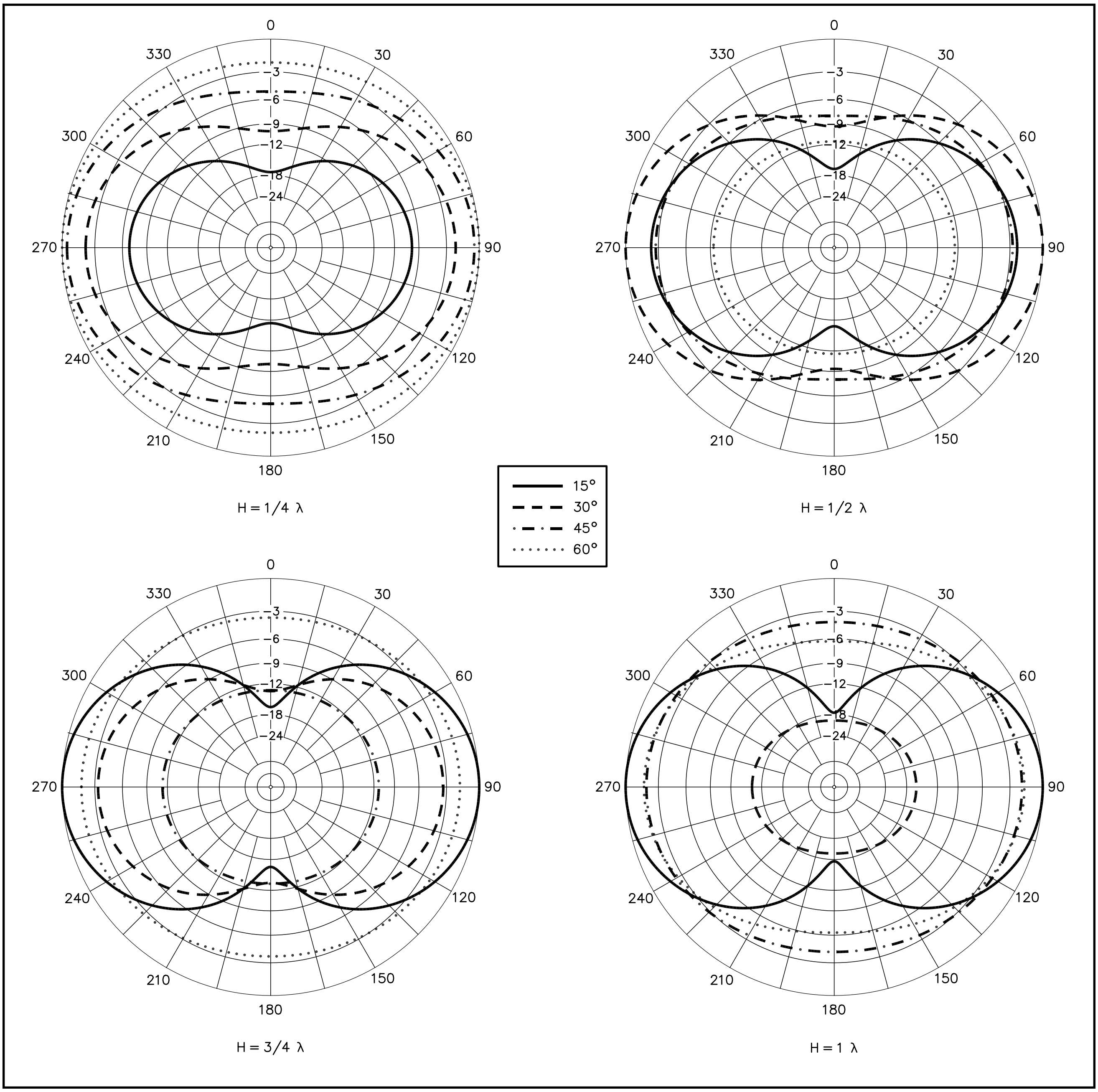


Fig 8-5 — Horizontal radiation pattern for ½-wave horizontal dipole at various heights above ground for wave angles of 15°, 30°, 45° and 60° (modeled over good ground).

tion at 90° and at 59° is very small (0.08 dB). The difference in radiated power at 90° between salt water and a desert type of reflecting ground is 2.25 dB. Since 90° is a radiation angle of little practical use, the relatively high loss at the zenith angle does not really bother us.

With a vertical antenna, poor ground results in loss at low angles, but with horizontal antennas the loss due to poor ground occurs at high angles. Notice that for a height of ½ λ (Fig 8-4B), the sharp null at a 90°-elevation angle for perfect ground has been degraded to a mere 12-dB attenuation over desert-type ground.

Conclusion

We can conclude that the effects of absorption over poor ground are pronounced with low horizontally polarized antennas and become less pronounced as the antenna height is increased. Artificial improvement of the ground conditions by the installation of ground wires is only practical if one wants maximum gain at a 90° wave angle (zenith) from a low dipole (⅛ to ¼ λ). This can be done by burying a number of wires (½ to 1 λ long) underneath the dipole, spaced about 60 cm apart, or by installing a parasitic reflector wire (½ λ long plus 5%) just above ground (2 meters high) under the dipole.

Improving the efficiency of the reflecting ground for low-angle signals produced by high horizontal dipoles is impractical and yields very little benefit. The active reflection area can be as far as 10 or more wavelengths away from the antenna!

Horizontal dipoles, unlike verticals, do not suffer to a great extent from poor ground conditions. The reason is that for horizontally polarized signals, when reflected by the ground, the phase shift remains almost constant at 180° (within 25°), whatever the incident angle of reflection (equal to the wave angle) may be. For verticals, the phase angle varies between 0° and 180°. For vertical antennas, the *pseudo-Brewster angle* is defined as the angle at which the phase shift at reflection is 90°. This means that there is no pseudo-Brewster angle with horizontally polarized antennas such as a dipole, because there never will be a 90° phase shift at the reflection point.

The effects are proved daily by the fact that on the low bands big signals from areas with poor ground conditions (mountainous, desert, etc) are always generated by horizontal antennas, while from areas with fertile, good RF ground, we often hear big signals from verticals and arrays made of verticals.

1.2.2. Horizontal Pattern of Horizontal Half-Wave Dipole

The horizontal radiation pattern of a dipole in free space has the shape of a figure 8. The horizontal directivity of a dipole over real ground depends on two factors:

- Antenna height
- The wave angle at which we measure the directivity

Fig 8-5 shows the horizontal directivity of half-wave horizontal dipoles at heights of ¼, ½, ¾ and 1 λ over average ground. Directivity patterns are included for wave angles of 15° through 60° in increments of 15°. At high angles a low dipole shows practically no horizontal directivity. At low angles, where it has more directivity, the low dipole hardly radiates at all. Therefore, it is quite useless to put two dipoles at right angles for better overall coverage if those dipoles are at low heights. At heights of ½ λ and more, there is discernible directivity, especially at low angles.

Fig 8-6 shows the three-dimensional radiation pattern of a half-wave dipole at ½ λ above average ground.

1.3. Half-Wave Dipole Efficiency

The radiation efficiency of an antenna is given by the equation

$$\text{Eff} = \frac{R_{rad}}{R_{rad} + R_{loss}} \qquad \text{(Eq 8-3)}$$

where

R_{rad} = radiation resistance, ohms
R_{loss} = loss resistance, ohms

1.3.1. Radiation Resistance

As defined in Section 2.4 in Chapter 5, *radiation resistance* (referred to a certain point in an antenna system) is the resistance, which if inserted at that point, would dissipate the same energy as is actually radiated from the antenna. *Radiation resistance* is a fictional resistance. For a half-wave dipole at or near resonance, the radiation resistance is equal to the real (resistive) part of the feed-point impedance, assuming a perfectly lossless antenna system.

The relationship of the radiation resistance and reactance of a half-wave dipole to its height above flat ground is shown in **Fig 8-7.** The radiation resistance varies between 60 and

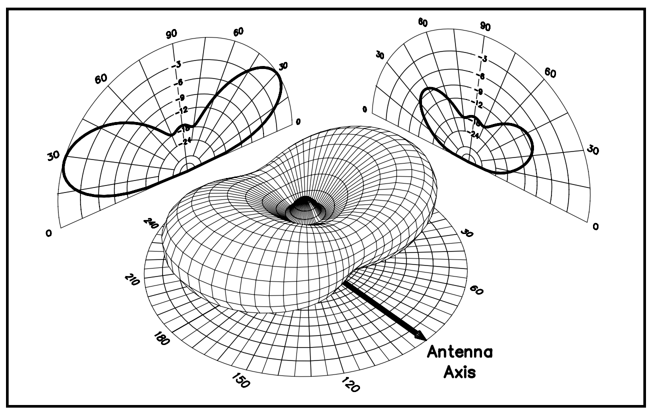


Fig 8-6 — Three-dimensional representation of the radiation patterns of a half-wave dipole, ½ λ above ground.

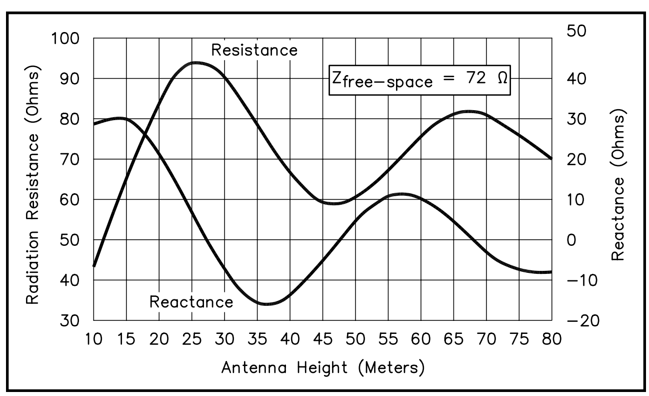


Fig 8-7 — Radiation resistance and feed-point reactance of a half-wave dipole at various heights. Calculations were done at 3.65 MHz using a 2 mm OD conductor (AWG #12 wire) over good ground.

Table 8-4
Resistance of Various Types of Wire Commonly Used for Constructing Antennas

Wire Diameter	Copper dc (Ω/km)	Copper 3.8 MHz (Ω/km)	Copper-Clad dc (Ω/km)	Copper-Clad 3.8 MHz (Ω/km)	Bronze dc (Ω/m)	Bronze 3.8 MHz (Ω/km)
2.5 mm (#10 AWG)	3.4	61	8.7	61	4.5	81
2.0 mm (#12 AWG)	5.4	97	13.8	97	7.2	130
1.6 mm (#14 AWG)	8.6	154	22.0	154	11.4	206
1.3 mm (#16 AWG)	13.6	246	35.0	246	18.2	326
1.0 mm (#18 AWG)	21.7	391	55.6	391	29.0	521

90 Ω for all practical heights on the low bands. For determining the reactance, the dipole was dimensioned to be resonant in free space (72 Ω). The resonant frequency changes with half-wave-dipole height above ground. Where the reactance is positive, the dipole appears to be too long, and too short where the reactance is negative.

1.3.2. Losses

The losses in a half-wave dipole are caused by:

- RF resistance of antenna conductor (wire)
- Dielectric losses of insulators
- Ground losses

Table 8-4 gives the effective RF resistance for common conductor materials, taking skin effect into account. The resistances are given in ohms per kilometer. The RF resistance values in the table are valid at 3.8 MHz. For 1.8 MHz the values must be divided by 1.4, while for 7.1 MHz the values must be multiplied by the same factor. The RF resistance of copper-clad steel is the same as for solid copper, since the steel core does not conduct any RF at HF. The dc resistance is higher by 3 to 4 times, depending on the copper/steel diameter ratio. The RF resistance at 3.8 MHz is 18 times higher than for dc (25 times for 7 MHz, and 13 times for 1.8 MHz). Steel wire is not shown in the table; it has a much higher RF resistance. Never use steel wire if you want good antenna performance.

1.3.2.1. Dielectric Losses in Insulators

Dielectric losses are difficult to assess quantitatively. Care should be taken to use good quality insulators, especially at the high-impedance ends of the dipole. Several insulators can be connected in series to improve the quality.

1.3.2.2. Ground Losses

Reflection of RF at ground level coincides with absorption in the case of non-ideal ground. With a perfect reflector, the gain of a dipole above ground is 6 dB over a dipole in free space. The field intensity doubles (as compared to the intensity of the antenna in a theoretical free space environment), since the same power is now radiated in a half sphere instead of a full sphere; double field intensity means 4 times power, which equals 6 dB gain.

Real ground is never a perfect reflector. Therefore some RF will be dissipated in the ground. The effects of power absorption in the real ground have been covered in Section 1.2.1.4. and illustrated in Fig 8-4 and Table 8-3.

Attempting to improve ground conductivity for improved performance is a common practice for vertical antennas. You can also improve ground conductivity under horizontal dipoles, although it is not quite as easy, especially if your are interested in low-angle radiation and if the antenna is physically high. From Table 8-1 you can find the distance from the antenna to the ground-reflection point. For the major low-angle lobe this is 36 meters away from an 80-meter dipole 30 meters high. Consequently, this is the place where the ground conductivity must be improved. Because of the horizontal polarization of the dipole, any wires that are laid on the ground (or buried in the ground) should be laid out parallel to the dipole. They should preferably be at least 1 λ long. If you look at improving ground conductivity for low angle radiation (eg 20°) you

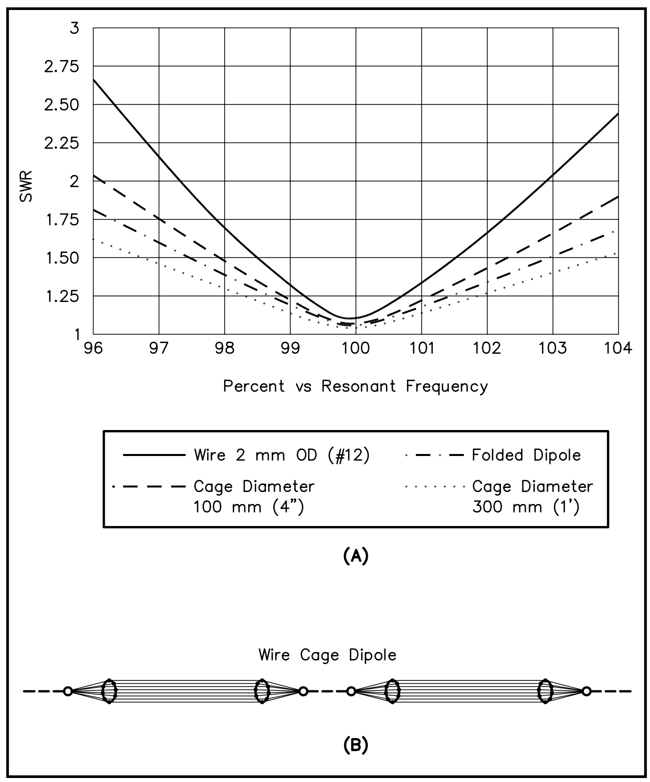


Fig 8-8 — At A, SWR plots for 3.75-MHz half-wave dipoles in free space of various conductor diameters. The total bandwidth of the 80-meter band (3.5 to 3.8 MHz) is 8%. The 100-mm (4-inch) and 300-mm (12-inch) diameter conductors can be made as a cage of wires, as shown at B. Note that the SWR bandwidth of a folded dipole is substantially better than for a straight dipole. The spacing between the wires of the folded dipole does not influence the bandwidth to a large extent.

are looking at distances of approximately 150 meters from the antenna, which is quite impractical.

In view of the small gain that can be realized, especially with high antennas and for low wave angles, it is very doubtful that such improvement of the ground is worth all the effort! The only really worthwhile improvement will be obtained by moving to the seacoast or to a very small island surrounded by saltwater. Don't forget that the quality of the reflecting ground with horizontal antennas is of far less importance than with vertical antennas.

The efficiency of low dipoles (¼ λ high and less), which essentially radiate at the zenith angle (90°), can be improved by placing wires under the antenna running in the same direction as the antenna.

1.4. Bandwidth of a Half-Wave Dipole

The SWR bandwidth of a full-size half-wave dipole is determined by the diameter of the conductor. **Fig 8-8A** shows the SWR curves for dipoles of different diameters. Large conductor diameters can be obtained by making a so-called *wire-cage* (Fig 8-8B). I used a wire-cage approach on my 80-meter vertical, with 6 wires forming a 30-cm (12-inch) diameter cage. **Fig 8-9** shows the effective equivalent diameter of such a cage conductor as a function of the number of wires making up the cage. Instead of using a wire cage you can also use a configuration consisting of a number of identical wires in a plane.

Fig 8-10 shows the effective equivalent diameters of such a flat multi-wire configuration. Example: Three parallel wires, each measuring 2 mm OD and equally spaced 5 cm, have an effective equivalent diameter of a solid conductor of 50 × 0.65 = 32.5 mm.

A folded dipole shows a substantially higher SWR bandwidth than a single-wire dipole. A folded dipole for 80 meters, made of AWG #12 wire, with a 15-cm (6-inch) spacing between the wires, will cover the entire 80-meter band (3.5 to 3.8 MHz) with an SWR of approximately 1.75:1, as compared to 2.5:1 or more for a straight dipole.

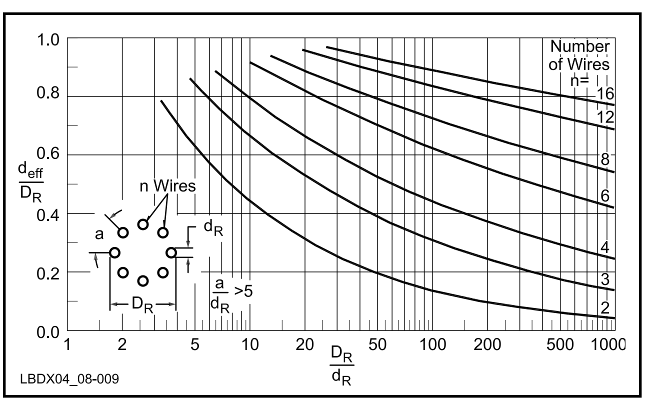


Fig 8-9 — Normalized effective diameter d_{eff}/D_R for a wire-cage conductor, made out of n conductors (diameter D_R). Example: a wire cage made out of 6 wires of 2-mm diameter, spaced equally on a circle measuring 20 cm in diameter, had an equivalent diameter of a single solid conductor of 0.62 × 20 = 125-mm diameter. (Source: *Kurze Antennen* by Gerd Janzen, ISBN 3-440-05469-1).

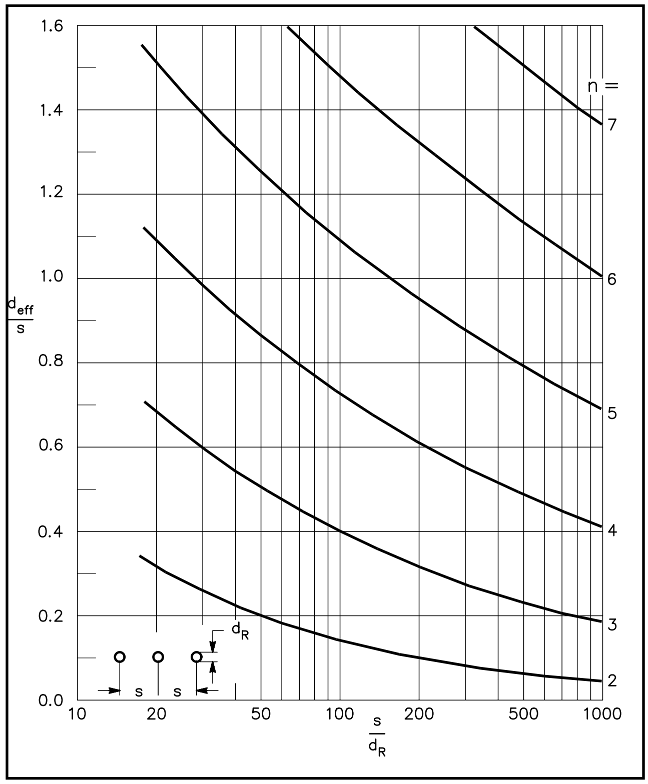


Fig 8-10 — Normalized effective diameter d_{eff}/D_R for a flat multi-wire conductor, made out of n conductors (diameter d_{eff}/D_R) spaced uniformly with spacing S. (Source: *Kurze Antennen* by Gerd Janzen, ISBN 3-440-05469-1).

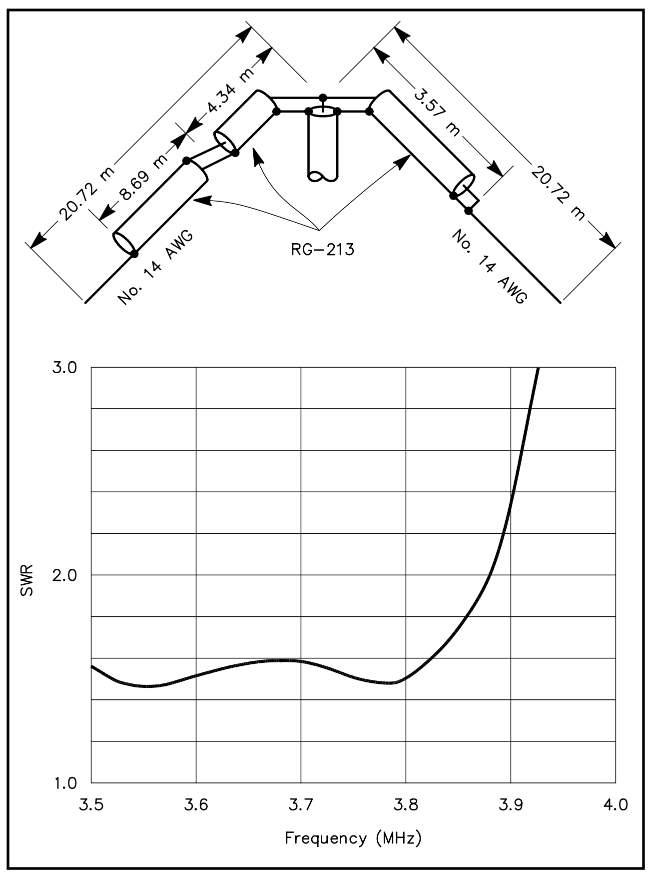


Fig 8-11 — Dimensions and SWR curve of the "80-meter DX Special," a design by F. Witt, AI1H.

1.4.1. Broadband Dipoles

Instead of decreasing the Q factor of the antenna, you can also devise a system to compensate the inductive part of the impedance as you move away from the resonant frequency of the antenna. The "double Bazooka dipole" is probably the best-known example of such an antenna, although it is rather controversial. In this antenna, part of the radiator is made of coaxial cable, connected in such a way as to present shunt impedances across the dipole feed point when moving away from the resonant frequency.

F. Witt, AI1H, designed a better broadband dipole antenna (Ref 1012). **Fig 8-11** shows the dimensions of Witt's *80-Meter DX-Special* antenna, which has been dimensioned for minimum SWR at both the CW and SSB ends of the 3.5 to 3.8-MHz band. Another innovative broadbanding technique was described by M. C. Hatley, GM3HAT (Ref 682).

R. Severns, N6LF, described an 80-meter folded broadband dipole using the principle of the open-sleeve antenna (Ref 1014). **Fig 8-12A** shows the layout of this folded-dipole, where Severns inserted another nearly half-wave long wire between the legs of the folded dipole. The resulting SWR curve is shown in Fig 8-12B.

Of course, there is no reason why you couldn't apply switched inductive or capacitive loading devices, such as those described in detail in the chapters on verticals and large loop antennas, although this is seldom done in practice.

1.4.2. Does a Resonant Dipole Radiate Better Than a Dipole Off-Resonance?

No, an infinitely short dipole would radiate as well as a full-size ½ λ dipole, provided you can get the same power into the short dipole, and provided the (normalized) losses are the same. Such an infinitely short dipole is called a *Hertzian* dipole. It has a constant current distribution and therefore a slightly different radiation pattern than a half-wave dipole. But that should not concern us, since it is a theoretical antenna anyhow.

If you have a really short dipole, the radiation resistance will be very low, maybe a few ohms, and the feed-point impedance at the center will be extremely capacitive (several thousand ohms). This makes it quite difficult to feed this very short antenna with a good efficiency (see Section 2). However, if the antenna is only slightly shorter (or longer) than a resonant half-wave dipole, its feed-point impedance will vary only a few percentage points from what it is at resonance. The reactive components will still be manageable so far as feeding this off-resonance dipole.

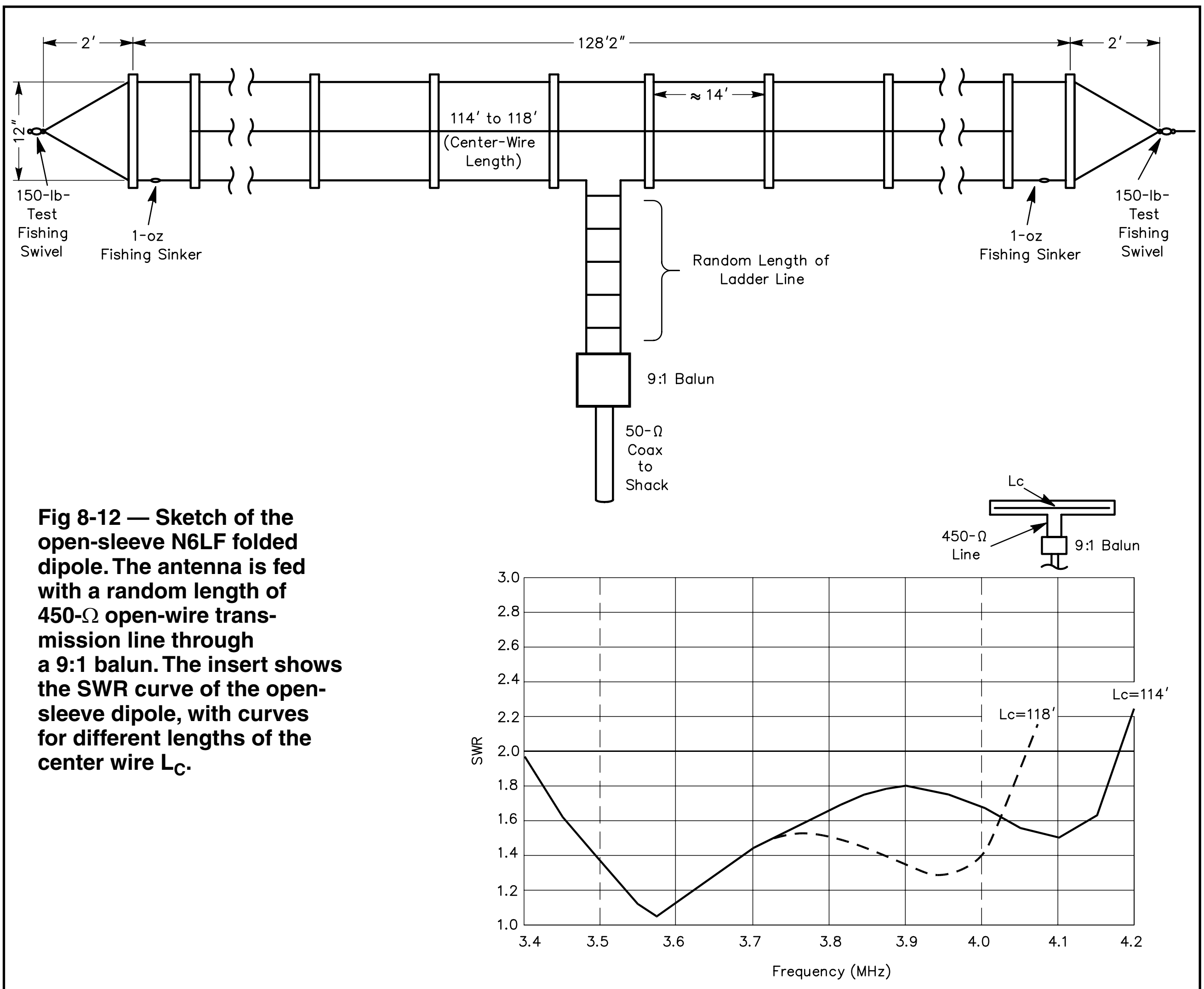


Fig 8-12 — Sketch of the open-sleeve N6LF folded dipole. The antenna is fed with a random length of 450-Ω open-wire transmission line through a 9:1 balun. The insert shows the SWR curve of the open-sleeve dipole, with curves for different lengths of the center wire L_C.

For example, let's take a single-wire center-fed dipole tuned for 3.65 MHz. Let's assume it is at a height where its impedance at resonance is exactly 50 Ω (see Fig 8-7). The antenna impedance at both 3.5 and at 3.8 MHz will be such that the SWR will be approximately 2:1 (still referred to our 50 Ω system impedance), and this is mainly caused by the reactive component. Whether or not this non-resonant antenna will radiate as much power as its resonant counterpart depends exclusively on how much loss there is in the feed system, now required to match a complex impedance: $40 + j\,70\ \Omega$ at 3.5 MHz and $60 + j\,70\ \Omega$ on 3.8 MHz. On the low bands, you can safely say that feed systems will show negligible losses when operated with SWRs below 2:1 or even 3:1.

Summarizing, the off-resonance dipole will radiate just as well as the resonant dipole. The only issue is the ease of feeding this dipole when it is far away from resonance. Under such conditions the feed line will exhibit a higher SWR than at resonance.

This does not mean that "reflected power" (reflected at the load, the antenna) will not be radiated. In other words, if the SWR meter indicates SWR = 3:1 (equals 25% reflected power), it does not mean that 25% of the power is *wasted*. In a lossless feed line system, all power will eventually be radiated, whatever the SWR. Our only concern should be a small increase in additional attenuation in a real-world feed line due to SWR.

In a properly tuned transmitter the effective power (P_{eff}) is always constant. For a higher SWR the reflected power (P_r) increases and the forward power increases correspondingly, in other words: $P_{eff} = P_f - P_r$ = constant. (See also Chapter 6, Section 2.)

1.4.3. Does Frequency of Lowest SWR Equal Resonant Frequency?

Is it true that the resonant frequency of a dipole is the frequency where the SWR is lowest? No, it is not. But is it important to know where is the exact resonant frequency of a dipole? No, it is not. What is generally important is to know the frequency where the dipole will cause the lowest SWR on the feed line.

Let me explain it with an example: A dipole has an impedance of 70 Ω at resonance and thus shows an SWR of 1.4:1 with 50 Ω coax. Somewhat lower in frequency, a combination of a lower resistive part with some capacitive reactance could result in a lower SWR than on the antenna's resonant frequency. It really depends on how fast the reactance changes compared to the resistance. But all of this should not bother us; we should cut our dipole for lowest SWR in the center of the (portion of the) band we want to cover. Whether or not this is the dipole's resonant frequency is irrelevant.

1.5. Feeding the Dipole

1.5.1. Impedance

The impedance of a dipole is usually close to 50 Ω, which means that there often is a perfect match.

1.5.2. Symmetrical Antenna and Asymmetrical Feed Line (Coax)

The half-wave dipole antenna, fed in the center, has a symmetrical layout. This means that the feed point is symmetrical (the current in one conductor is 180° out of phase with the current in the other conductor). If you feed this antenna with a symmetrical feed line (open-wire line, ribbon line), this feed line will not radiate (except at very close distances) because of the opposite feed current in the two close-spaced conductors. The antenna should be fed with an open-wire transmission line if it is to be used on different frequencies (eg, as two half-waves in phase on the first harmonic frequency), because the open-wire line is a low loss line even when operated at high SWR.

If we feed the dipole at its balanced feed point with an asymmetrical feed line, such as a coaxial feed line, things are different.

1.5.3. The Inside World and the Outside World of a Coaxial Feed Line

Viewed from *inside* a coaxial feed line, we see two conductors: the center conductor and the shield (the inside

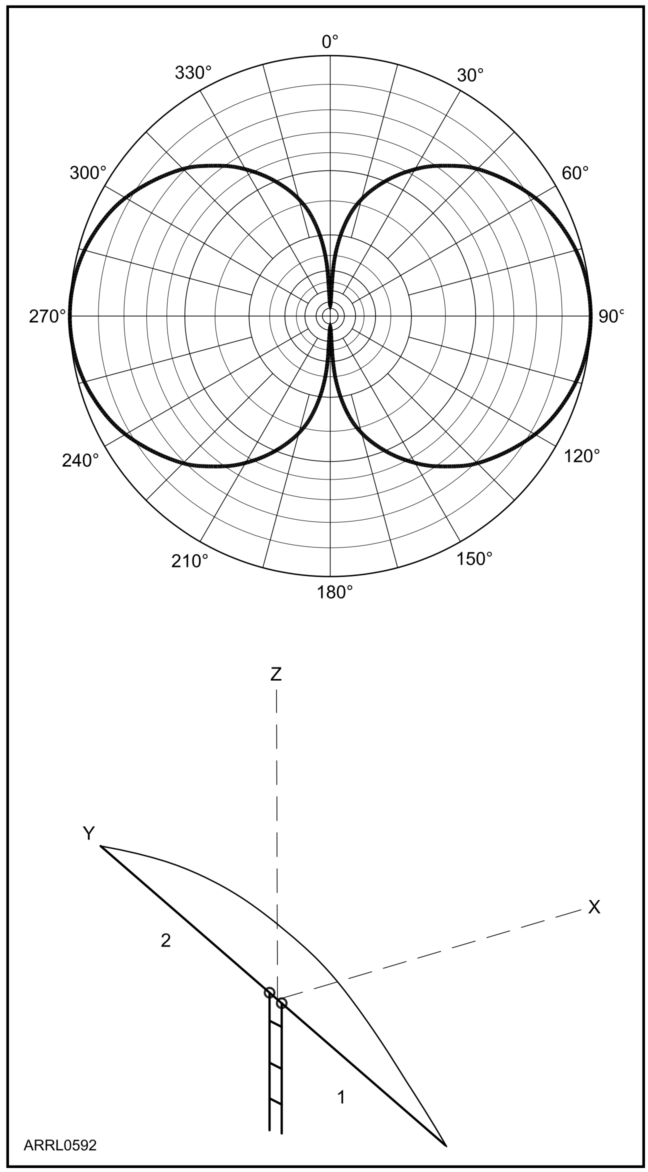


Fig 8-13 — The half-wave dipole in free space, fed with a symmetrical feed line. As the currents in the two conductors of the feed line are out of phase, the line does not radiate. Hence the radiation pattern only radiates in the X-Y plane.

of the outer conductor "tube"). For transporting RF energy, current flows in the center conductor (or more precisely, near the surface of this conductor because of the skin effect) and on the inner side of the outer conductor (the tube, shield or braid). The currents in the center conductor and in the inner side of the shield are equal but opposite in phase. If you use a current probe and measure the current on the *outer side* of the shield, you will see that there is no current at all on the outer side of the shield, provided the feed line is terminated in an unbalanced load (such as a dummy load).

If we move ourselves *outside* of the coaxial cable, and look at the coax from a distance, all we see is a "thick" (heavy gauge) conductor. For the outside world the coax is just a fat conductor.

That means that if for some reason RF current flows on the outer side of the coax, it will behave just like any other conductor: it will radiate. The current will travel on it with the speed it does on any conductor in the air (maybe a little slower because of an insulated jacket on top of the shield), which means that the velocity factor (VF) will be close to maybe 98%, while the speed of travel of the RF signal *inside* the coax will be much slower (66% in case of solid PE dielectric). So, if you use a coax as a radiating wire, apply approximately 0.98 as VF and not 0.66 or similar!

1.5.4. Current on the Inside *and* on the Outside

Fig 8-13 shows the current distribution on a half wave dipole, when fed in its center with a symmetrical feed line. The currents in the two conductors are in phase opposition. The radiation pattern (in free space) is the perfect figure 8, the impedance is 73 Ω and the gain is 2.14 dBi. Just for the heck of it, the SWR in a 50 Ω system is 1.4:1.

Let's now connect a 50 Ω coaxial feed line to the dipole, and let it hang at a right angle from the center of the dipole. The λ/4 length is the electrical length as seen from the "outside" world (using VF = 98%). **Fig 8-14** shows the new situation. The coax cable is connected directly to the dipole and no balun is used. This means that at the feed point wires are simply connected together, and both are driven by the same voltage. If we connect the inner conductor of the coax to wire 1, we need to connect the outer conductor to wire 2. Wire 2 provides the "point to push against."

Notice however that now the "return" current from the current flowing in (on) wire 1 (the right half of the dipole) is now divided into two currents, one returning in (on) the left half of the dipole (2) and one returning on the outer shield of the coaxial feed line (3). This part of the return current is called the "common mode" current. As the feed line (seen from the outside world) is λ/4 long, no current flows at the end, and this wire shows a perfect sinusoidal current distribution that is 90° long. The vertically hanging feed line and the current flowing on its outer shield now also contribute to the radiation of the antenna, and the antenna now radiates not only a horizontally polarized signal but also a vertically polarized component.

The gain has dropped from 2.14 to 1.87 dBi, the impedance has changed from 72 Ω to 51 Ω, and the SWR (in a 50 Ω system) is now 1:1. Going by these measurements one might think this is the better situation of the two, as the impedance match is much better.

We should not forget that we do not only have current on the outside of the coax, but also voltage! If you tape the coax against a metal structure, chances are that the voltage difference between the metal structure and the shield of the coax will cause arcing and burn holes in the plastic coax jacket, which will rapidly turn into a water gutter, due to the capillary effect of the braided copper shield.

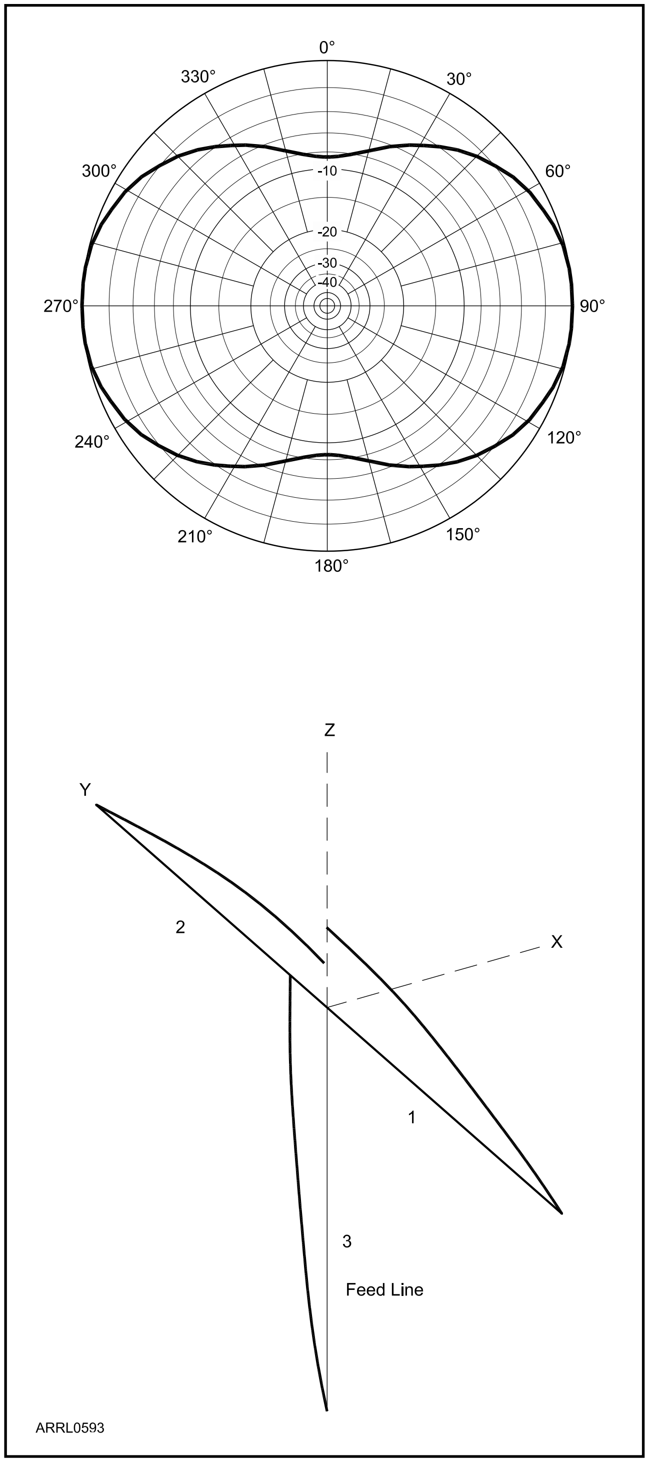


Fig 8-14 — The same half-wave dipole shown in Fig 8-13, but now fed with a coaxial feed line without balun (common-mode choke). In this case the outside of the shield is just another "half-dipole" that collects the return currents. The current on the outside of the coax creates radiation in the Z-Y plane, resulting in less total gain and less directivity.

In addition we are now radiating energy in directions and in a polarization where we should not, reducing the useful gain of the antenna. A third inconvenience is that the feed line may run along buildings where we have installed TVs, audio equipment and other electronic systems; radiation may cause havoc because of the very close vicinity. Finally we may also bring high levels of RF in the shack and cause problems with RF detection and the like.

If the feed line length is not an odd multiple of ¼ λ, the current on the outer shield will be less. In other words, the amount of common mode current will depend on various factors, the main one being the feed line length and whether or not the feed line outer shield is grounded at one or several points.

The cure for common-mode currents on the feed line is to cut the feed line at the feed point, but only as far as the outside world is concerned. If we insert an RF choke (in this case commonly called a common mode choke) that has an impedance high enough to prevent any significant current flow on the outside of the feed line, we have in fact cut the feed line for these unwanted currents. In stubborn cases one can add a second common mode choke at distances of uneven quarter wavelengths (¼, ¾ λ etc) from the antenna feed point. For more details and how to make a current balun, see Chapter 6, Section 7).

Conclusion: always use a 1:1 balun or common mode choke on a dipole!

1.6. Getting the Full-Size Dipole in Your Backyard

The ends of the half-wave dipole can be bent (vertically or horizontally) without noticeable effect on the radiation pattern or efficiency. The tips of the dipoles carry little current; hence, they contribute very little to the radiation of the antenna.

Bending the tips of a dipole is the same as "end loading" the dipole (equal to top-loading with verticals). The folded tips can be considered as capacitive loading devices. For more details see Section 3 in Chapter 9 on Vertical Antennas.

N. Mullani, KØNM, calculated the gain and the impedance of a half-wave dipole with its end hanging down vertically (Ref 691). He concluded that with horizontal lengths as short as 40% of full-size, the trade-offs are rather insignificant, being only about 0.6 dB in gain, and some reduction of SWR bandwidth. Bending the end may actually somewhat improve the match to a 50 Ω feed line, depending on antenna height. KØNM concludes: "Don't be afraid to bend your dipole antennas if you are cramped for space." (See also Chapter 14 for more limited space antenna ideas.)

2. THE SHORTENED HALF-WAVE DIPOLE

On the low bands, it is sometimes impossible to use full-size radiators. This section describes the characteristics of short dipoles, and how they can be successfully deployed. Short dipoles are often used as elements in reduced-size Yagis (see Chapter 13 on Yagis) or to achieve manageable dimensions whereby the antenna can be fit into a city lot.

Short antennas are the subjects of an excellent book (in German language) by Gerd Janzen, DF6SJ/VK2BJZ (Ref 7818). This book is highly recommended for anyone who does not fear a formula and a graph, and who really wants to dig a little deeper into the subject.

2.1. The Principles

You can always look at a dipole as two back-to-back connected verticals, except that the "vertical" elements are no longer vertical. Instead of having the ground make the mirror image of the antenna (this is always the case with quarter-wave monopole verticals), we supply the mirror half ourselves in a dipole. All principles about radiation resistance and loading of short verticals, as explained in Chapter 9 on vertical antennas, can be directly applied to dipoles as well.

2.2. Radiation Resistance

The radiation resistance of a dipole (made of an infinitely thin conductor) in free space will be twice the value of the equivalent vertical monopole. For instance, the R_{rad} for the half-wave dipole made of an infinitely thin conductor is approximately 73.2 Ω, which is twice the value of the quarter-wave vertical (36.6 Ω). Over ground, the infinitely thin horizontal dipole's radiation resistance will vary in a similar way as the full-size half-wave dipole (see Fig 8-7).

2.3. Tuning or Loading the Short Dipole

Loading a short dipole consists of bringing the antenna to resonance. This means eliminating the capacitive reactance component in the feed-point impedance. Different loading methods yield different values of radiation resistance.

It is not necessary however, to load a shortened antenna to resonance in order to operate it. You could connect a feed line to it, directly or via a matching network, without tuning out the capacitive reactance. Therefore you can consider a dipole together with its feed line as a *dipole system*, and analyze the system of a short dipole to see what the alternatives are. Sometimes this situation is referred to as a dipole with "tuned feeders."

There are different ways to operate the short dipole system:

- Tuned feeders
- Matching at the dipole feed point
- Coil loading to tune out the capacitive reactance
- Linear loading
- Capacitive end loading
- Combined loading methods

2.3.1. Tuned Feeders

Tuned feeders were common in the days before the arrival of coaxial feed lines. Very low-loss open-wire feeders can be made. The *Levy* antenna is an example of a short dipole fed with open-wire line. In a more or less typical configuration its overall length is ¼ λ on the lowest band it is supposed to operate on. Using *EZNEC* (*NEC-2*) we calculated the feed impedance of a quarter-wave long "dipole" (Levy) as $Z = 13.6 - j\,1060\ \Omega$.

In principle the antenna can be fed with open-wire feeders (450-600 Ω) of any length from the shack, where we can match it to 50 Ω with an antenna tuner. At first sight, an outstanding feature of this approach is that the system can be "tuned" from the shack via the antenna tuner. It is not narrow banded, as is the case with loaded elements.

But, let us have a look at some numbers. A 450 Ω window ladder line has a matched line loss of 0.5 dB/100 meters on 80 meters. We can analyze the situation with *TLW* (by N6BV), which is available on the *ARRL Antenna Book* CD. At the end of a 30 meter (100 foot) long feed line the impedance becomes

$(44.2 + j\,1221)\ \Omega$. At the load (antenna) the SWR is 200:1, at the end of the line (at the tuner) it is 98:1, both rather mind-boggling values. The total line loss went from 0.15 dB (flat line) to 4.1 dB, which means that we are dissipating somehow more than 50% of the transmitter output power in the 30 meter long feed line. Imagine you run 1000 W. We now have only 389 W left at symmetrical load terminals of the tuner. Forgot to mention one thing: with these impedances we will have approximately 10 kV peak voltage on the transmission line. The ladder line will certainly not take it — it will arc, burn, disintegrate. And we have not yet talked about a tuner that will handle these impedances.

In a typical tuner, using good quality components (Q Coil = 200, Q Cap ≥1000), the coil will dissipate another 47 W and the capacitor a "mere" 18 W. We now have approximately 324 W out of 1000 W to radiate, which means we are looking at a total loss of almost 5 dB! And we have not yet mentioned the balun. Both L and C components will have to withstand voltages of nearly 10 kV! Maybe we could live with a 1.2:1 SWR and just tune out the inductive reactance (+1221 Ω) with a suitable series capacitor. This suitable capacitor would have a value of 36 pF but should be able to withstand a voltage of not less than 7 kV! The reason for these "numbers" is the extreme transformation ratio that is required. This the only reason why, in actual practice, we cannot make a very short antenna radiate as well as an antenna of "reasonable" size. For more details on this subject visit: **www.w8ji.com/short_dipoles_and_problems.htm**.

All of this is to explain that direct feeding of "very" short antennas via ladder or open wire line to a tuner is not a good idea, unless you run really low power and are happy to settle for a lot of losses.

2.3.2. Matching at the Dipole Feed Point

You could, of course, install a matching network at the dipole feed point, although this will be highly impractical in most cases. In the case of a vertical antenna this solution is practical, since the feed point is at ground level.

In principle, having the tuner at the antenna feed point would eliminate the problems with the feed line losses and the unwanted impedance transformation, but you would still be confronted with an extreme impedance transformation ratio. A tuner handling such impedance would still get a lot of beating, and require, for example, a coil in the tuner that can handle well over 10 kV peak voltage. Not an easy task.

2.3.3. Coil Loading

A loading coil simply inserts a series inductive reactance that cancels capacitive antenna reactance. Loading coils can be installed anywhere in the short dipole halves, from the center to way out near the end. Loading near the end will result in a higher radiation resistance, but will also require a much larger coil, and hence introduce more coil losses.

2.3.3.1. Center Loading

The inductive reactance required to resonate the Levy dipole from the previous example is approximately +1060 Ω (data from modeling in free space). To achieve this, two 530 Ω (reactance) coils must be installed in series at the feed point. We should be able to realize a coil Q (quality factor) of 300. With good care 500 to 600 can be achieved as well (Ref 694).

Let's assume a Q factor of 400:

$R_{loss} = 1060/400 = 2.65\ \Omega$

The total equivalent loss resistance of the two coils is 2.65 Ω. The antenna efficiency will be:

$\text{Eff} = 13.6/(13.6 + 2.65) = 83\ \%$

The equivalent power loss is $-10 \log (0.83) = 0.86$ dB.

The feed-point resistance of the antenna is 13.6 + 2.65 = 16.25 Ω at resonance. This assumes negligible losses from the antenna conductor (heavy copper wire). If the use of coaxial feed lines is desired, an additional matching system will be needed to adapt the 16.25 Ω balanced feed-point impedance to the 50 Ω or 75 Ω unbalanced coaxial cable impedance. This example was calculated assuming free-space impedances. Over real ground the impedances can be different, and will vary as a function of the antenna height.

Calculation over real ground

Let's work out the following example using *EZNEC* (*NEC-2*):

Input data:
f = 3.65 MHz
h = 25 meters
$L_{ant} = \lambda/4$ (half size), L = 20 meters

EZNEC tells us that the impedance is $13.59 - j\,1060\ \Omega$ (modeled with 11 pulses).

The required center loading coil has a reactance of 1060 Ω. Assuming a loading-coil Q of 400, the total equivalent loss resistance is 2.65 Ω. The feed-point resistance becomes 13.59 + 2.65 = 16.21 Ω.

Note that the figures obtained by this method are in line with the numbers discussed earlier (free space).

Matching to the feed-line impedance

One way of matching this impedance to a 50 Ω feed line is to use a quarter-wave transformer. The required impedance of the transformer is

$$Z_0 = \sqrt{15.24 \times 50} = 27.6\ \Omega$$

We can construct a feed line of 25 Ω (that's close) by paralleling two 50 Ω feed lines. The SWR at resonance in a 50 Ω system will be ~ 1.1:1. Don't forget you need a 1:1 balun between the antenna terminals and the feed line.

Another attractive matching scheme used by a number of commercial manufacturers of short 40-meter Yagis is to use a single central loading coil, on which we install a link at the center. The link turns are adjusted to give a perfect match to the feed line.

Comparing losses with the open-wire case

The loading coil (Q factor = 400) gives a loss resistance of 2.65 Ω, and the efficiency is 13.9/(13.9 + 2.64) = 84% or a loss of 0.8 dB. Add 30 meters (100 feet) of RG-213 (with 0.23 dB loss), and the total system loss can be estimated at approximately 1.03 dB. Note that this is more than 1 dB better (less than half the loss) than the result we calculated for open-wire feeders.

There are certain advantages and disadvantages to this concept, however. An advantage is that coaxial cable is easier to handle than open-wire line, especially when dealing with rotatable antenna systems. The high Q of the coils will make the antenna narrow-banded as far as the SWR is concerned. In the case of the open-wire feeders, retuning the tuner will solve the problem. With coaxial feed line you may still nee d a tuner at the input end if you want to cover a large bandwidth, in which case the extra losses due to SWR in the coaxial feed line may be objectionable.

Another disadvantage is that the loading-coil solution requires two more elements in the system: the coils. Each element in itself is an extra reliability risk, and even the best loading coils will age and require maintenance.

Calculating the coil value

The required total loading coil reactance value was 1060 Ω. The required inductance is:

$$L = \frac{X_L}{2\,\pi\,f}$$

where L is in μH and f is in MHz. For 3.65 MHz:

$$L = \frac{1060}{2\,\pi\,3.65} = 46.2\ \mu H$$

There are two ways of loading and feeding the shortened dipole with a centrally located loading coil:

- Use a single 46.2 μH loading coil and link couple the feed line to the coil. This method is used by Cushcraft for their shortened 40-meter antennas.
- A 46.2-μH loading coil can be opened in the center where it can be fed by a 1:1 balun.

A real high-Q coil is wound using 3 mm wire, with a coil diameter of 10 cm, a coil length of 210 cm and 35 turns (spacing between turns = 3 mm). The calculated unloaded Q ~ 750.

2.3.3.2. Loading Coils Away from the Center of the Dipole

The location of the loading devices has a distinct influence on the radiation resistance of the antenna. This phenomenon is explained in detail in the chapter on short verticals.

Clearly, it is advantageous to move loading coils away from the center, provided the benefit of higher radiation resistance is not counteracted by higher losses in the loading device. It appears, however, that in practice there is very little difference.

The required coil inductance increases when loading coils are placed farther out on the elements. With increasing values of inductance, the Q factor is likely to decrease, and the equivalent series losses will increase.

I have calculated a case where the 22.5-meter long dipole (for 3.8 MHz) from Section 2.3.3.1 was loaded with coils at different (symmetrical) positions along the half-dipole elements. In all cases I assumed a Q factor of 300.

The results of the case are shown in **Fig 8-15** for a dipole made with 2.5-mm diameter wire and **Fig 8-16** for a dipole with an average conductor diameter of 25 mm. The charts include the reactance value of the required loading coils, the radiation resistance (R_{rad}), and the feed-point impedance at resonance (Z). The radiation efficiency is given by R_{rad}/Z. Over the entire experiment range the efficiency remains practically constant at 88% for 2.5 mm conductor diameter and 91% for 25 mm conductor diameter. You can also use a wire-cage type dipole (see Fig 8-8) to achieve an effective conductor diameter of 250 mm, yielding an efficiency of 95%. **Table 8-5** shows the influence of the coil Q and the effective wire diameter on the radiation efficiency of a shortened dipole.

The fact that the efficiency does not change much by moving the coils out on the elements means that the advantage we gain from an increased radiation resistance by moving the coils out on the dipole halves is balanced out by the increased ohmic losses of the higher coil values. In the experiment I assumed a constant Q of 300, which may not be realistic, as it is likely that the Q of lower-inductance coils will be higher than higher-inductance ones.

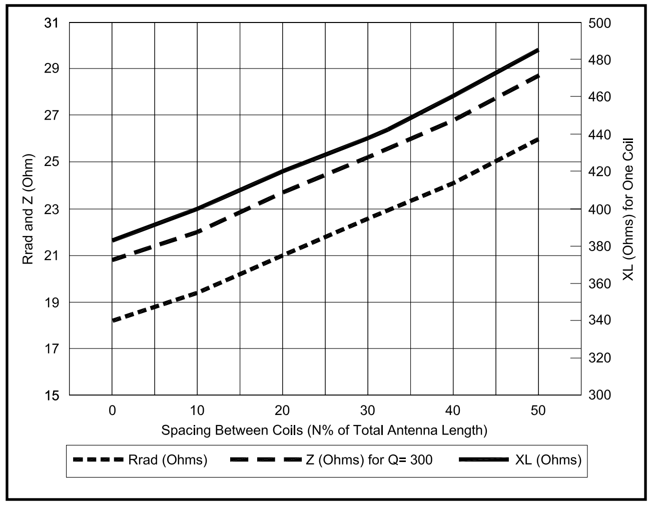


Fig 8-15 — Design data for a 22.5-meter shortened dipole (F = 3.8 MHz, wire diameter = 2.5 mm), showing R_{rad}, Z_{feed} and required loading coil reactance X_L as a function of the spacing between the loading coils. Where the spacing between the coils is zero (center loading) the coil reactance is twice the value shown (2 × 386 Ω).

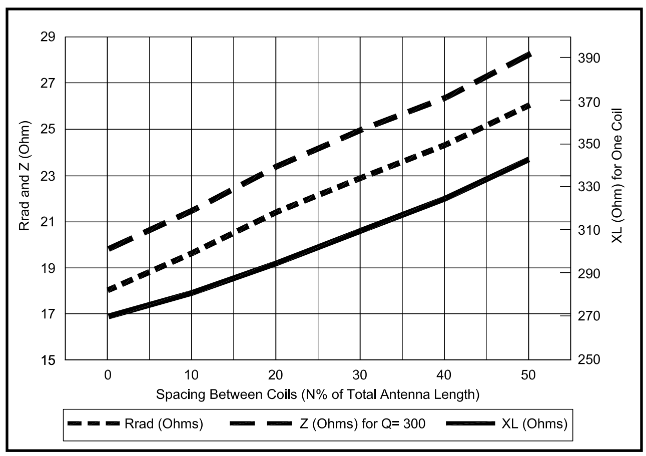


Fig 8-16 — Design data for a 22.5-meter shortened dipole (F= 3.8 MHz, wire diameter = 25 mm), showing R_{rad}, Z_{feed} and required loading coil reactance X_L as a function of the spacing between the loading coils. Where the spacing between the coils is zero (center loading) the coil reactance is twice the value shown (2 × 269 Ω).

Table 8-6
Coil Characteristics for a Wire Dipole

Coil(s) at	*Required Reactance (Ω)*	*Inductance Coil (μH)*	*Rrad (Ω)*	*Gain (dBi)*
Center	$+j$ 645 Ω total	One coil: 27.1	24.4 Ω	1.71
3.25 meters from center	2 × $+j$ 440 Ω	2 coils: 18.5	34.4 Ω	1.75
6.25 meters from center	2 × $+j$ 605 Ω	2 coils: 25.4	40.3 Ω	1.75
7 meters from center	2 × $+j$ 820 Ω	2 coils: 34.4	44.0 Ω	1.75

Table 8-7
Coil Characteristics for a Cage Dipole

Coil(s) at	*Required reactance (Ω)*	*Inductance Coil (μH)*	*Rrad (Ω)*	*Gain (dBi)*
Center	$+j$ 255 Ω total	One coil: 10.7	23.9 Ω	1.83
3.25 meters from center	2 × $+j$ 173 Ω	2 coils: 7.25	36.0 Ω	1.86
6.25 meters from center	2 × $+j$ 230 Ω	2 coils: 9.2	42.0 Ω	1.86
7 meters from center	2 × $+j$ 300 Ω	2 coils: 12.6	46.0 Ω	1.87

Table 8-5
Antenna radiation efficiency as a function of loading coil Q and conductor diameter

Conductor	*Q=100*	*Q=300*	*Q=600*
Diam 2.5 mm	72 %	88 %	94 %
Diam 25 mm	78 %	91 %	96 %
Diam 250 mm	84 %	95 %	98 %

In Table 8-5 we see the influence of dropping the Q to 100 (pretty lousy) and raising it to 600 (excellent). The spread between the minimum wire diameter combined with the worst coil (Q = 100) and the maximum wire diameter with the best coil (Q = 600) is from 72% to 98%, which means a difference of 1.4 dB in signal strength.

Another marked advantage of using the large-diameter conductor is a substantially increased SWR bandwidth. The loaded 22.5-meter long dipole made of 2.5-mm diameter wire has a 2:1 SWR bandwidth of 50 kHz on 75 meters. The same dipole made out of a wire cage (6 wires in a circle with a 300-mm diameter, yielding an effective diameter of 250 mm) has a 2:1 SWR bandwidth of 100 kHz.

I did the calculation above in free space. Over real ground the radiation resistance (and Z) will vary to a rather large extent as a function of the height (see Fig 8-7). With the large diameter dipole (effective 25 mm diameter) we need only a reactance of 310 Ω to center load the 22.5-meter long dipole for 3.8 MHz. This is compared to 765 Ω for a wire dipole of the same length with 2.5 mm diameter.

Calculating the loading coil value

Using an antenna modeling program, such as *EZNEC*, it is fairly straightforward to calculate the loading coil value.

Physical length: 25 meters
Frequency: 3.8 MHz
Conductor diameter: 2 mm

Assumption: we can make coils with Q = 600. Where should we put the coils for best efficiency?

I modeled the antenna in free space (*EZNEC*, *NEC-2*) and put the coils in the middle of the antenna, 2.5 meters from the center, 5 meters from the center and 6 meters from the center and 8 meters from the center. See **Table 8-6**.

The SWR bandwidth remains almost constant (approximately 100 kHz for 2:1 SWR). As noted earlier, it is not critical where you put the coils. The gain obtained through the higher R_{rad} is compensated by the extra loss from coil Q.

If more bandwidth is required, one could make a cage dipole with an effective diameter of 250 mm (see Fig 8-9). **Table 8-7** shows the details. The bandwidth has doubled (200 kHz for 2:1 SWR limits), the loading coils are much smaller (less than half the inductance) and the gain is approximately 0.1 dB higher.

Conclusion

Use loading coils with the highest possible Q (400 to 600) and reduce the Q-factor of the antenna (the rate at which the reactive component of the impedance changes with frequency) by using a large diameter using a cage-type construction. The exact position of the coil does not significantly influence the antenna efficiency and is far from critical.

A high-Q coil for use as an antenna loading coil can usually be made when using a turn to turn space that equals the conductor diameter, and when the L/D (length to diameter) ratio of the coil is between 2 and 4. This is quite different from more common applications of coils such as in amplifier networks, where optimum Q can usually be obtained with L/D ratios close to 1. Always use air-wound coils and stay away from coil forms or dielectric coatings. For more details visit **www.w8ji.com/mobile_and_loaded_antenna.htm** and **www.w8ji.com/loading_inductors.htm**. Commercial high-quality inductors are available from Barker & Williamson (**www.bwantennas.com/coils/coilcat.htm**). The Airdux TL stock coils can be used to make high quality loading coils.

Calculating coils

The CD that comes with this book carries a coil calculator (*coil.exe*) as part of the *Low Band Software* package. This program calculates air and toroidal coils using a rather simple formula. Another coil calculator (available from **hamwaves.com/antennas/inductances.html**) is from the hand of Serge Stroobbantf, ON4AA. The ON4AA Coil Calculator is based on the work by the Corum brothers (one is Dr James F. Corum, K1AON) and the additional work by David Knight, G3YNH.

As an example, let us design a loading coil for the loaded dipole from Table 8-6 (coil at 6.25 meters distance from center). The required inductance is 25.4 μH. For a coil diameter of 7.5 cm, pitch of 6 mm (3 mm wire and 3 mm inter-windings spacing) the *coil.exe* program calculates 32.7 turns with a coil length of 20 cm (L/D ratio = 2.7). ON4AA's program calculates 32.8 turns, which for all practical reasons is the same. The program also calculates an unloaded Q of 875.

2.3.4. Linear Loading

In the commercial world, we have seen *linear loading* used on shortened dipoles and Yagis for 40 and 80 meters. Linear-loading devices are usually installed at or near the center of the dipole. The required length of the loading device (in each dipole half) will be somewhat longer than the difference between the quarter-wave length and the physical length of the half-dipole. The farther away from the center that the loading device will be inserted, the longer the "stub" will have to be. The stub must run in parallel with the antenna wire if we want to take advantage of any radiation from the stub itself (see the chapter on vertical antennas).

In the last several years the linear loading technique has largely been replaced with high-Q coil loading which appears to yield a better efficiency. W8JI calls linear loading devices nothing other than a poor form-factor inductor.

When constructing an antenna with linear-loading devices, make sure the separation between the element and the folded linear-loading device is large enough, and that you use high-quality insulators to prevent arc-over and insulator damage. If directivity is not an issue you can hang the linear loading "stubs" vertically from the dipole.

Modeling the linear loaded dipole

Modeling antennas that use very close-spaced conductors (such as a linear-loading device that looks like a stub made of open-wire transmission line) is very tricky. You need a *NEC-4* based program to obtain proper results (see Chapter 4, Section 1.4).

If linearly loaded dipoles are used as elements of an array, it is very important that the linear loading devices run horizontally (in-line with the element) and not at an angle. If it is at an angle there will be vertically polarized radiation from the linear loading conductors, and this will affect the directivity of the antenna.

2.3.5. Capacitive (End) Loading

Capacitive loading has the advantage of physically shortening the element length at the end of the dipole where the current is lowest (least radiation), and without introducing noticeable losses (as inductors do). End-loaded short dipoles have the highest radiation resistance, and the intrinsic losses of the loading device are negligible. Thus, end loading (equivalent to top loading in case of a vertical) is highly recommended.

In case of a short horizontal dipole, the easiest way to apply capacitive end loading is to let the ends of the dipole drop down vertically. These wires carry little current and hence contribute only marginally to radiation, but their capacity effect lowers the resonance of the shortened dipole. Watch out, the end of these wires carry very high voltage and must be kept out of reach of humans and animals.

Calculating the top load

The easiest way is to use a modeling program based on *NEC-2* or *NEC-4*. Example:

F_{design} = 3.80 MHz
Length shortened dipole = 25 meters
Diameter conductor = 2 mm

Without loading, the dipole resonates at approximately 5.6 MHz. *EZNEC* (*NEC-2*) tells us we need to add approximately 7.35 meters of wire at each end to reach resonance on 3.8 MHz (see **Fig 8-17**). Note that this is nearly the same length (5% longer) than the "missing" length (compared with a full-size dipole).

R_{rad} = 56 Ω, and gain (free space) = 1.78 dBi. This is a mere 0.35 dB less than for the full-size dipole! The SWR (2:1) bandwidth of the full-size straight dipole is ~290 kHz, for this top loaded version it is ~190 kHz. Compared to the coil loaded dipole from Table 8-6 we notice that R_{rad} is a little higher (56 Ω vs approximately 42 Ω) while the calculated gain is about the same (1.78 dBi vs 1.75 dBi). The most important difference is that the 2:1 SWR bandwidth is 190 kHz compared to only 100 kHz for the coil loaded dipole.

A practical way to design such an end-loaded dipole is to simply add plenty of wire at the tips, and cut these ends back (symmetrically) until you obtain the lowest SWR where you want it!

SteppIR element

A SteppIR element (**Fig 8-18**) looks like a trombone

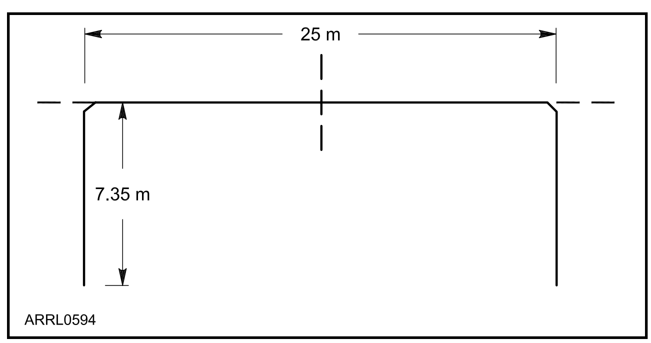


Fig 8-17 — Horizontal dipole with the tips hanging down.

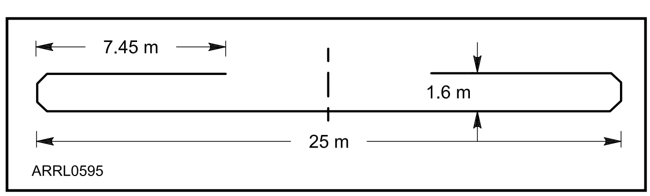


Fig 8-18 — In an 80 meter SteppIR element, the capacitive top loading wires are folded back along the element in the form of a trombone.

(folded dipole) element, but where the element measures approximately 60% of the length of a full size dipole. We modeled such an element for 3.8 MHz. Using a conductor with a diameter of 2 mm we found the element to have an impedance of 38 Ω, a gain of 1.72 dBi and a 2:1 SWR bandwidth of 120 kHz. With this attractive approach you do not need a heavy coil, but the spacing of the two wires of 1.6 meters requires some other mechanical engineering skills to make a stable construction.

2.3.6. Inductive vs Capacitive Loading

If you can get by with simple capacitive end loading (drooping ends, bent wires near the ends of the dipole halves and so on), this is always the most efficient way of "loading" a short dipole. The only small drawback is that top loading structures that are not symmetrical will radiate a vertical component in the far field, but this component is usually so much suppressed that it will not be a problem.

2.3.7. Combined Methods

Any of the loading methods already discussed can be employed in combination. It is essential to develop a system that gives you the highest possible radiation resistance and that employs a loading technique with the lowest possible inherent losses.

2.4. Bandwidth

The bandwidth of a dipole is determined by the Q factor of the antenna. The antenna Q factor is defined by:

$$Q = \frac{Z_S}{R_{rad} + R_{loss}}$$

where
Z_S = surge impedance of the antenna
R_{rad} = radiation resistance
R_{loss} = total loss resistance.

The 3-dB bandwidth can be calculated from:

$$BW = \frac{f_{MHz}}{Q}$$

The Q factor (and consequently the bandwidth) will depend on:

- The conductor-to-wavelength ratio (influences Z_S).
- The physical length of the antenna (influences R_{rad}).
- The type, quality and placement of the loading devices (influences R_{rad}).
- The Q factor of the loading device(s) (influences R_{loss}).
- The height of the dipole above ground (influences R_{rad}).

For a given conductor length-to-diameter ratio and a given antenna height, the loaded antenna with the narrowest bandwidth will be the antenna with the highest efficiency. Indeed, large bandwidths can easily be achieved by incorporating pure resistors in the loading devices, The worst-radiating antenna one can imagine is a dummy load, where the resistor is a loading device while the radiating component does not exist. Judging by SWR bandwidth alone, the dummy load is a wonderful "antenna," since a good dummy load can have an almost flat SWR curve over thousands of megahertz!

2.5. The Efficiency of the Shortened Dipole

Besides the radiation resistance, the RF loss resistance of the shortened-dipole conductor is an important factor in the antenna efficiency. Refer to Table 8-4 for the RF loss resistances of common wire conductors used for antennas. For self-supporting elements, aluminum tubing is usually used. Both the dc and RF resistances are quite low, but special care should be taken to ensure that you make the best possible electrical RF contacts between parts of the antenna. Some makers of military-specification antennas go so far as to gold plate the contact surfaces for low RF resistance!

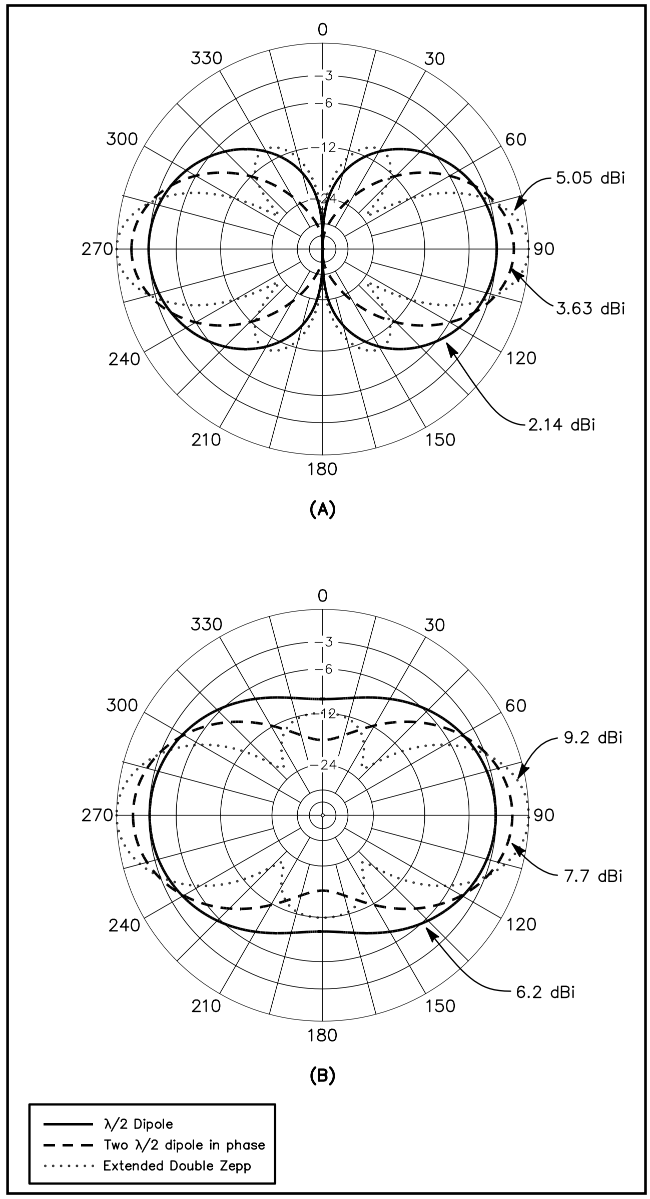


Fig 8-19 — Horizontal radiation patterns for three types of dipoles: The half-wave dipole, the collinear dipole (two half waves in phase) and the extended double Zepp. At A, the radiation patterns at a 0° wave angles with the antennas in free space. At B, the patterns at a 37° wave angle with the antennas ⅜ λ above good-quality ground. Notice the side-lobes apparent with the extended double Zepp antenna.

As a rule, loading coils are the most lossy elements, and capacitive end loading should be employed if possible. If you use coil loading, take care to construct loading coils with Qs in the range of 400 to 600, or better.

It is very important to minimize the contact losses at any point in the antenna, especially where high currents are present. Corroded contacts can turn a good antenna into a radiating dummy load. These aspects are covered in more detail in the chapter on vertical antennas.

3. LONG DIPOLES

Provided the correct current distribution is maintained, long dipoles can give more gain and increased horizontal directivity compared to a half-wave dipole. The "long" antennas discussed in this paragraph are not strictly dipoles, but arrays of dipoles. They are the double-sized equivalents of the "long-verticals" covered in the chapter on verticals. The following antennas are covered:

• Two half-waves in phase
• Extended double Zepp

3.1. Radiation Patterns

Center-fed dipoles can be lengthened to approximately 1.25 λ to achieve increased directivity and gain without introducing objectionable side lobes. **Fig 8-19** shows the horizontal radiation patterns for three antennas in free space: a half-wave dipole, two half-waves in phase (also called *collinear dipoles*), and the extended double Zepp, which is 1.25 λ long. Further lengthening of the dipole introduces major secondary lobes in the horizontal pattern unless phasing stubs are inserted to achieve the correct phasing between the half-wave elements.

As we know, a half-wave dipole has 2.14-dB gain over an isotropic antenna in free space. It is interesting to overlay the patterns of the two long dipoles on the same diagram, using the same dB scale. The extended double Zepp beats the dipole with almost 3 dB of gain. Note, however, how much more narrow the forward lobe on the pattern has become. This may be a disadvantage in view of varying propagation paths. The two-half-waves antenna is right between the dipole and the extended double Zepp, with 1.5-dB gain over the half-wave dipole.

Fig 8-20 and **Fig 8-21** show the horizontal radiation patterns for the half-waves-in-phase dipole and the extended double Zepp at various heights and wave angles. As with the half-wave dipole, the vertical radiation pattern depends on the height of the antenna above ground.

3.2. Feed-point Impedance of Long Dipoles

The charts from Chapter 9 (Figs 9-8 through 9-13) can be used to estimate the feed-point impedances of long dipoles. The values from the charts that are made for monopoles must be doubled for dipole antennas.

The easiest way to analyze the behavior of these antennas is to model them with the appropriate modeling software. I recommend not using *MININEC*, but rather a *NEC-2* (or *NEC-4*) based program such as *EZNEC*, especially when the antenna is less than 0.25 λ above ground. *NEC-2 or NEC-4-*based programs using the Sommerfeld-Norton ground model are accurate at any height above 0.001 wavelength.

Since center-fed long antennas are not loaded with lossy tuning elements that would reduce their efficiency, long dipoles

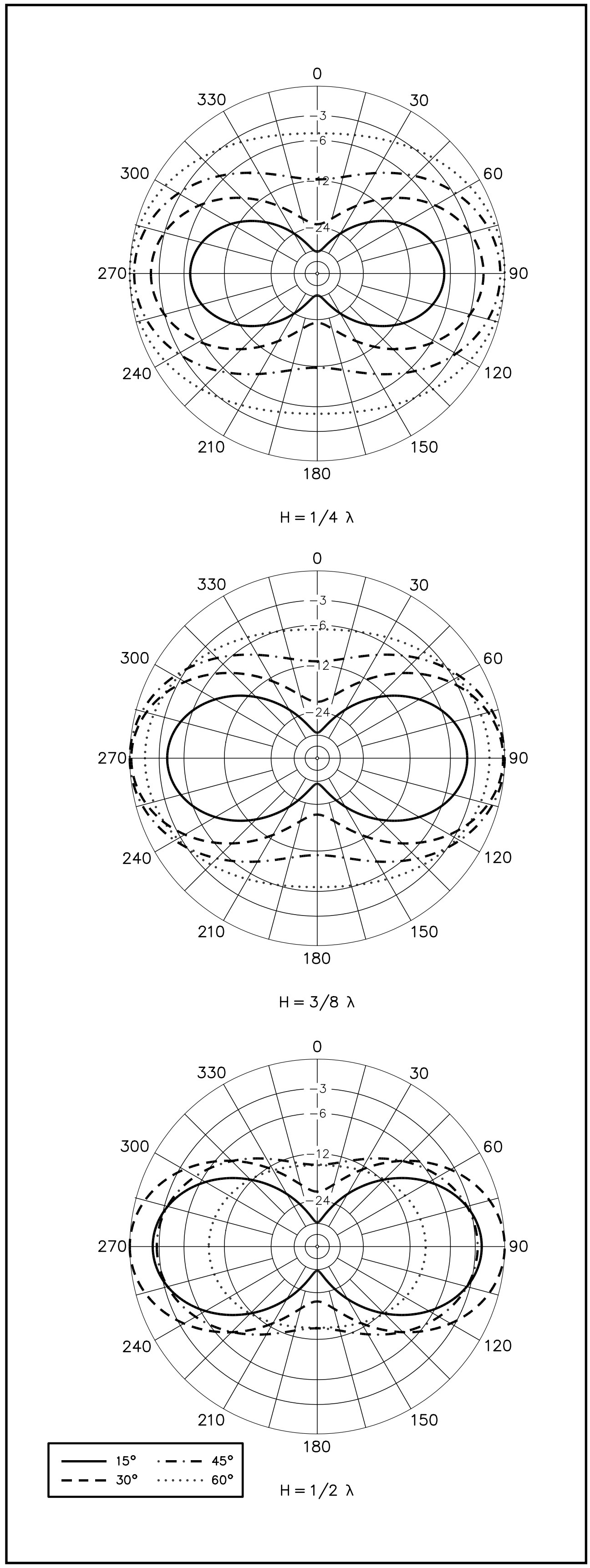


Fig 8-20 — Horizontal radiation patterns for collinear dipoles (two half-waves in phase) for wave angles of 15°, 30°, 45° and 60°. As with a half-wave dipole, directivity is not very pronounced at low heights and at high wave angles.

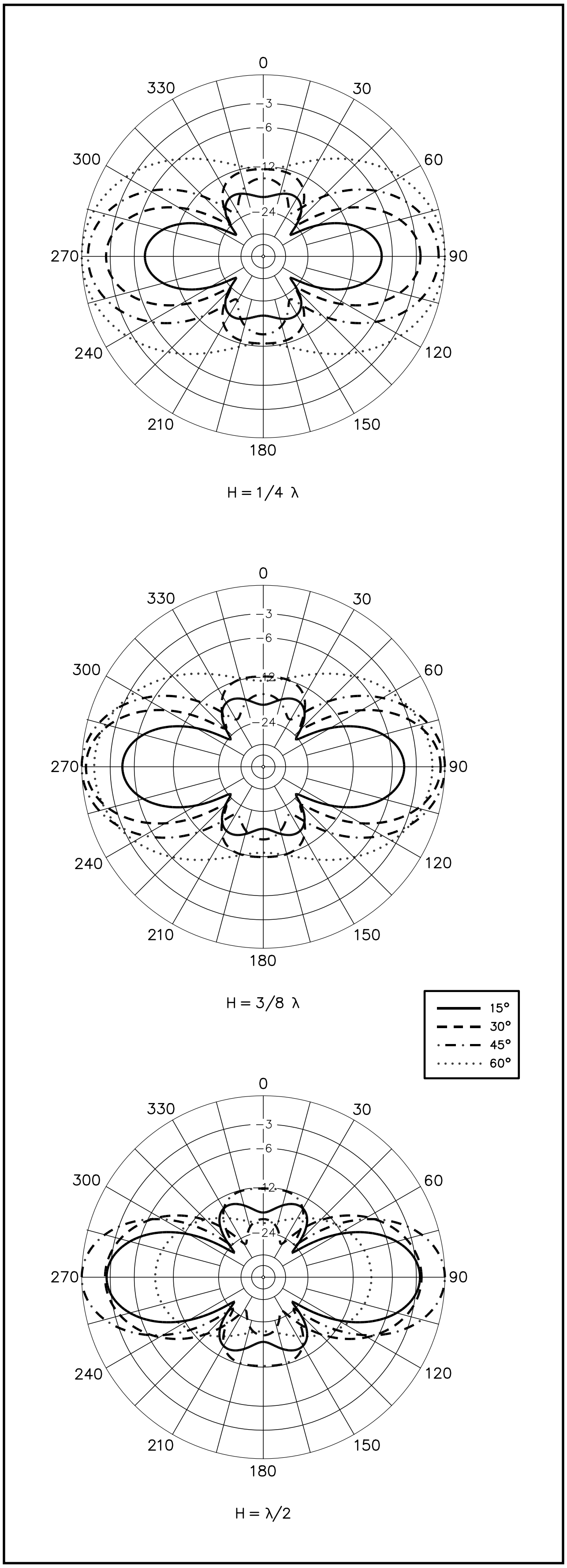


Fig 8-21 — Horizontal radiation patterns for the extended double Zepp for wave angles of 15°, 30°, 45° and 60°.

can have efficiencies very close to 100%, provided care is taken to use the best materials for the antenna conductor.

3.3. Feeding Long Dipoles

The software module Coax Transformer/Smith Chart from the *New Low Band Software* is an ideal tool for analyzing the impedances, currents, voltages and losses on transmission lines. The Stub Matching module can assist you in calculating a stub-matching system in seconds. In any case, we need to know the feed-point impedance of the antenna. Measuring the feed-point impedance is quite difficult, since you cannot use an impedance bridge unless it is specially configured for measuring balanced loads.

3.3.1. Collinear Dipoles (Two Half-Wave Dipoles in Phase)

The impedance at resonance for two half-waves in phase is several thousand ohms. With a 2-mm OD conductor (AWG #12), the impedance is approximately 6000 Ω on 3.5 MHz. The shortening factor in free space for this antenna is 0.952. The normalized SWR bandwidth of the two half-waves in phase is given in **Fig 8-22**. The antenna covers a frequency range from 3.5 to 3.8 MHz with an SWR of less than 2:1.

The antenna can be fed with open-wire feeders into a tuner, or via a stub-matching system and balun as shown in Chapter 6, Fig 6-15. Using tuned feeders with a tuner can, of course, ensure a 1:1 SWR to the transmitter (50 Ω) at all times.

3.3.2. Extended Double Zepp

The intrinsic SWR bandwidth of the extended double Zepp is much narrower than for the collinear dipoles. For an antenna made out of 2-mm OD wire (AWG #12) and with a total length of 1.24 λ, the feed-point impedance is approximately $200 - j\,1100\ \lambda$. The SWR curve (normalized to R_{rad} at the design frequency) is given in Fig 8-22. For lengths varying

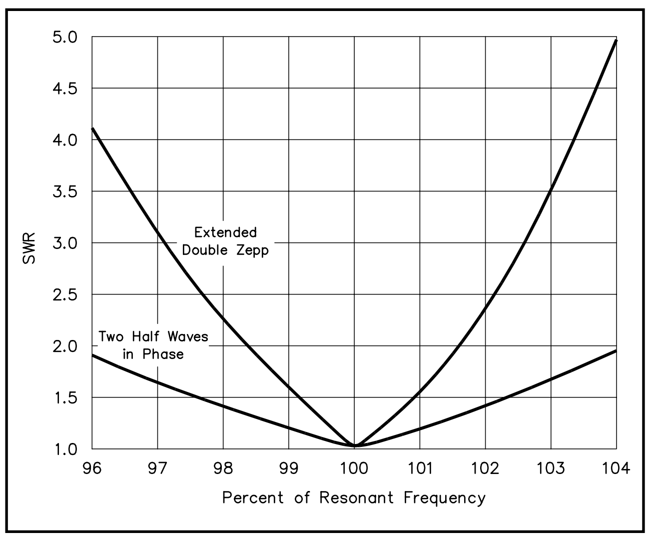


Fig 8-22 — SWR curves for an extended double Zepp and for two half waves in phase (collinear array). The calculation was centered on 3.65 MHz using a conductor of 2-mm OD (AWG #12), and the results normalized to the radiation resistances. The SWR bandwidth of the collinear array is much higher than for the double extended Zepp.

from 1.24 to 1.29 λ, the radiation resistance will vary between 130 and 200 Ω (decreasing resistance with increasing length).

The exact length of the antenna is not critical, but as we increase the length, the amplitude of the sidelobes increases. The magnitude of the reactance will depend on the length/diameter ratio of the antenna. An antenna made of a thin conductor will show a large reactance value, while the same antenna made of a large-diameter conductor will show much less reactance.

The impedance of the extended double Zepp also changes with antenna height, as with a regular half-wave dipole. For the 1.24-λ long extended double Zepp, the resistive part changes between 150 and 260 Ω, and settles at 200 Ω at very high heights.

In principle, we can feed this antenna in exactly the same way as the collinear, but since the intrinsic bandwidth is much more limited, it is better to feed the antenna with open-wire lines running all the way into the shack and to an open-wire antenna tuner.

3.4. Three-Band Antenna (40, 80, 160 Meters)

Refer to the three-band antenna of **Fig 8-23**. On 40 meters the antenna is a collinear array (two half-waves in phase) at a height of 24 meters (80 feet). On 80 meters, it is a half-wave dipole. For 160, we connect the two conductors of the open-wire feeders together, and the antenna is now a flat-top loaded vertical (T antenna). The disadvantage is that we must install a switchable tuning network at the base, right under the antenna.

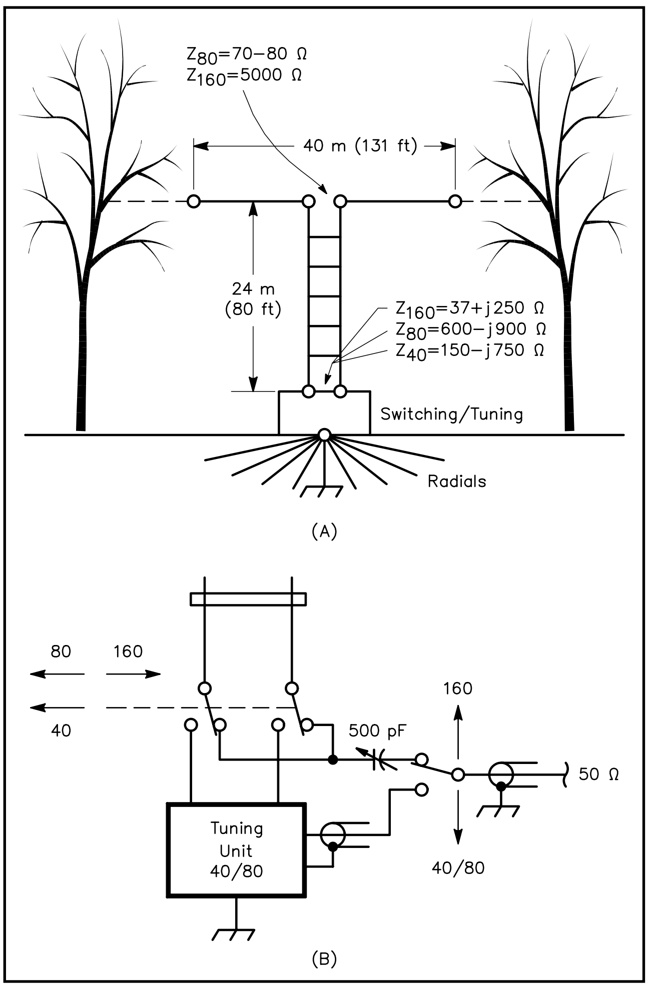


Fig 8-23 — Three-band antenna configuration (40, 80 and 160 meters). On 40 meters the antenna is a collinear (two half waves in phase); on 80 meters a half-wave dipole; and on 160 meters a top-loaded T vertical. The band switching arrangement is shown at B.

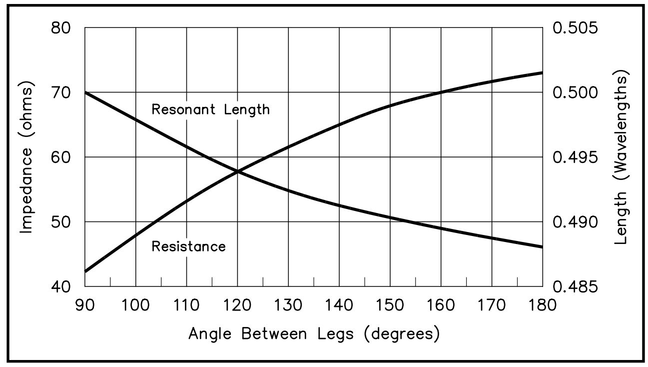


Fig 8-24 — Radiation resistance (resistance at resonance) of the inverted-V dipole antenna in free space as a function of the angle between the legs of the dipole (apex angle). Also shown is the physical length (based on the free-space wavelength) for which resonance occurs.

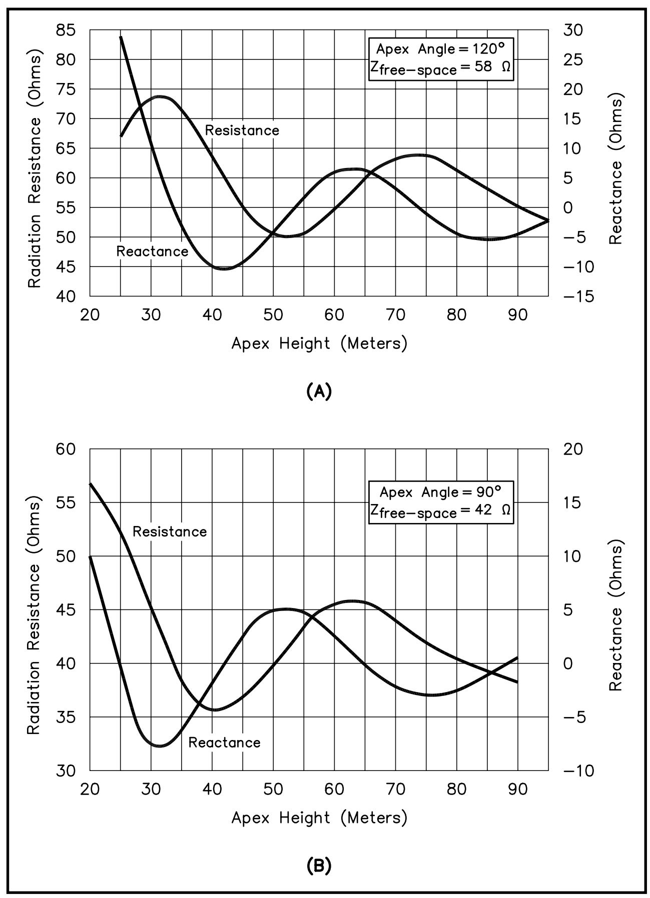


Fig 8-25 — Impedance (feed-point resistance and reactance) of inverted-V dipoles as a function of height above ground. Analysis frequency is 3.75 MHz, with a 2-mm OD wire (AWG #12). Resistances at resonance are: 120° apex angle, 58 Ω; 90° apex angle, 42 Ω. *NEC-2* was used for these calculations, since *MININEC* is unreliable for impedance at low heights.

Since the antenna is a vertical on 160 meters, its performance will largely depend on the quality of the ground and the radial system. Some slope away from vertical can, of course, be allowed in the feed line.

4. INVERTED-V DIPOLE

In the past, the inverted-V shaped dipole has often been credited with almost magical properties. The most frequently claimed special property is a low radiation angle. Some have more correctly called it a *poor man's dipole*, since it requires only one high support. Here are the facts.

4.1. Radiation Resistance

The radiation resistance of the inverted-V dipole changes with height above ground (as does a horizontal dipole) and as a function of the apex angle, which is the angle between the legs of the dipole. Consider the two apex-angle extremes. When the angle is 180°, the inverted-V becomes a flattop dipole, and the radiation resistance in free space is 73 Ω. Now take the case where the apex angle is 0°. The inverted-V dipole becomes an open-wire transmission line, a quarter-wavelength long and open at the far end. This configuration will not radiate at all. The current distribution will completely cancel all radiation, as it should in a well-balanced feed line. The input impedance of the line is 0 Ω, since a quarter-wave stub open at the end reflects a dead short at the input. This zero-angle inverted-V will have a radiation resistance of 0 Ω and consequently will not radiate at all.

I modeled a range of inverted-V dipoles with different apex angles at different apex heights. This was done using *NEC-2*. **Fig 8-24** shows the radiation resistance as a function of the apex angle for a range of angles between 90° and 180° (a flattop dipole). The curve also shows the physical length that produces resonance, where the feed point is purely resistive. Decreasing the apex angle raises the resonant frequency of the inverted-V.

Fig 8-25 shows the feed-point resistance and reactance for inverted-V dipoles with apex angles of 120° and 90°. The antennas were first resonated in free space. Then the reactances were calculated over ground at various heights. Notice that the shape of both curves is similar to the shape of the straight dipole curve in Fig 8-7. Bringing the inverted-V closer to ground lowers its resonant frequency. This is a fairly linear function between 0.25 λ and 0.5 λ apex height.

4.2. Radiation Patterns and Gain

So far we have compared the inverted-V to a straight dipole at the same apex height. It is clear that the inverted-V is a compromise antenna when compared to the straight horizontal dipole. At low heights (0.25 to 0.35 λ), the gain difference is minimal, but at heights that produce low-angle radiation the dipole performs substantially better.

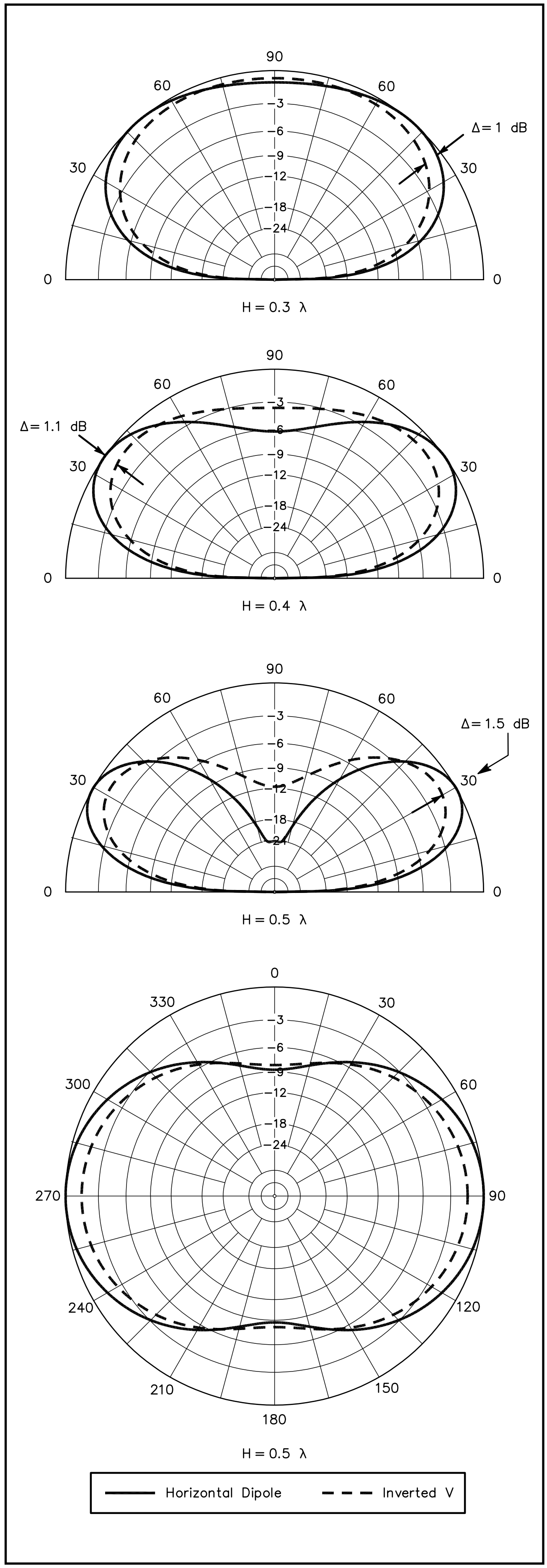


Fig 8-26 — Radiation patterns for an inverted-V dipole with an apex angle of 90°. For comparison, the radiation pattern of a horizontal dipole is included in each plot, on the same dB scale. The horizontal pattern is shown for the main wave angle (28° for the straight dipole and 32° for the inverted-V). The height of the inverted-V is the height at its apex, which is the same height as the flattop horizontal dipole.

The 90° apex angle inverted-V dipole

Fig 8-26 shows the vertical and horizontal radiation patterns for inverted-Vs with a 90° apex angle at different apex heights. Modeling was done over good ground. For comparison, I have included the radiation pattern for a straight dipole at the same apex height. In the broadside direction, the inverted-V dipole shows 1 to 1.5 dB less gain than the flattop dipole and also a slightly higher wave angle.

The 120° apex angle inverted-V dipole

The flat-top dipole is 0.6 dB better than the inverted-V at a height of 0.4 λ; 0.7 dB at 0.45 λ and 0.8 dB at 0.5 λ. In addition, the wave angle for the horizontal dipole is slightly lower than for the inverted-V, at approximately 3° for heights from 0.35 λ to 0.5 λ. The difference is not spectacular but it is clear that the inverted-V dipole has no magical properties.

4.3. Antenna Height

In many situations it will be possible to erect an inverted-V dipole antenna much higher than a flat top dipole, in most cases because there is only one high support structure available. In this respect the high inverted-V can be superior to a low horizontal dipole. The inverted-V loses compared to a flattop dipole at the same apex height, but not everyone has two such high supports. And if they do, are they in the right direction?

4.4. Length of the Inverted-V Dipole

The usual formulas for calculating the length of the straight dipole cannot be applied to the inverted-V dipole. The length depends on both the apex angle of the antenna and the height of the antenna above ground. Fig 8-25 shows the feed-point impedance for inverted-Vs of different configurations at different heights.

Closing the legs of the inverted-V in free space will increase the resonant frequency. On the other hand, the antenna will become electrically longer when closer to the ground due to the end-loading effect of the ground on the inverted-V ends.

4.5. Bandwidth

Fig 8-27 shows the SWR curves for three inverted-V dipoles with different apex angles: 90°, 120° and 180° (flattop dipole), for a conductor diameter of 2 mm (AWG #12) and a frequency of 3.65 MHz. As expected, the SWR bandwidth decreases with decreasing apex angle. The computed figures are for free space. The SWR values in Fig 8-27 are normalized figures. This means that the SWR at resonance is assumed to be 1:1, whatever the actual impedance (resistance) at resonance is. In practice, the SWR will almost never be 1:1 at resonance because the line impedance will be different from the feed-point impedance (see the impedance chart in Fig 8-7).

Over ground, the reactive part of the impedance remains almost the same value as in free space, after you have re-resonated the inverted-V at the center frequency. This means that the SWR bandwidth will be largest for heights where the radiation resistance is highest. For the inverted-V dipole this is at an apex height of approximately 0.35 to 0.4 λ. Practically speaking, it means that for an apex height of 0.3 λ to 0.5 λ, the SWR curve will be somewhat flatter over ground than in free space.

The SWR bandwidth of the inverted-V can be increased

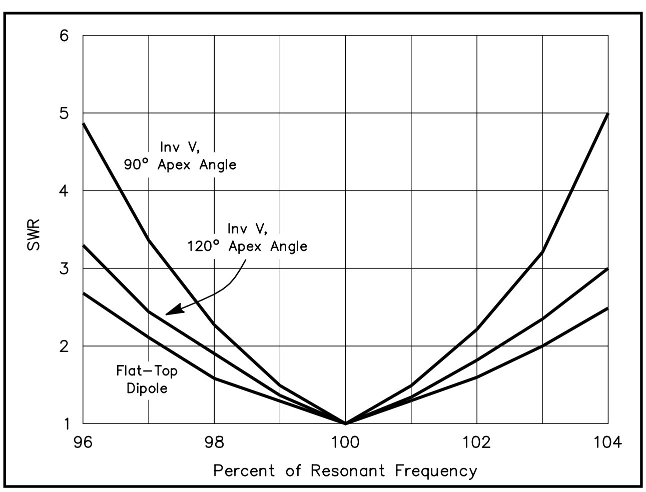


Fig 8-27 — SWR curves for three types of free-space half-wave dipoles: The horizontal (flattop) dipole, and inverted-V dipoles with apex angles of 120° and 90°. Each curve is normalized to the feed-point resistance at resonance.

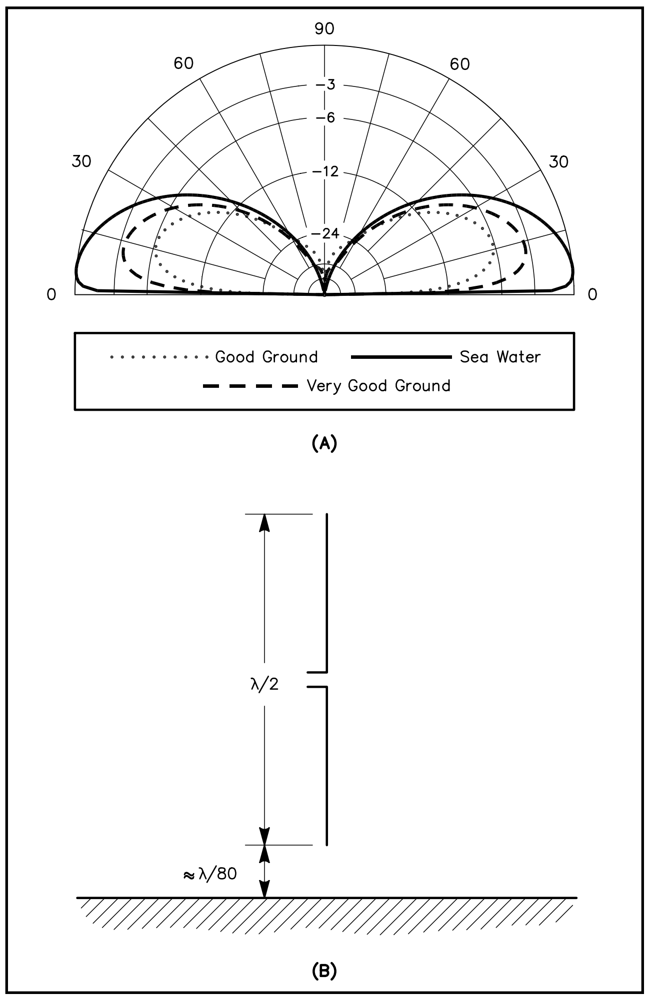


Fig 8-28 — At A, vertical radiation patterns over various grounds for a vertical half-wave center-fed dipole with the bottom tip just clearing the ground, as shown at B. The gain is as high as 6.1 dBi over ground. The feed-point impedance is 100 Ω.

significantly by making a folded-wire version of the antenna. The feed-point impedance of the folded-wire version is four times the impedance shown in Figs 8-24 and 8-25.

5. VERTICAL DIPOLE

The half-wave vertical is covered in detail in the chapter on vertical antennas. Whereas in that chapter we consider the half-wave vertical mainly as a base-fed antenna, we can of course use a dipole made of wire and feed it in the center. This is what we usually call a *vertical dipole*. In many practical cases a wire half-wave vertical will not be perfectly vertical, but will generally slope away from a tall support such as a tower or a building. Sloping half-wave verticals are covered in Section 6.

5.1. Radiation Patterns

Whether the half-wave vertical is base fed or fed in the center, the current distribution is identical, and hence the radiation pattern will be identical. Radiation patterns are shown in **Fig 8-28** when the lower end is near the ground. Over saltwater the half-wave vertical can yield 6.1-dBi gain, which drops to about 0 dBi over good soil. As with all verticals, it is mainly the quality of the ground in the Fresnel zone that determines how good a low-angle radiator the vertical dipole will be (see Section 4 of Chapter 9). Half-wave verticals produce excellent (very) low-angle radiation when erected in close proximity to saltwater. As a general-purpose DX antenna the vertical dipole may, however, produce too low an angle of radiation for some nearby DX paths.

Raising the half-wave vertical higher above the ground introduces multiple lobes. **Fig 8-29** shows the patterns for a half-wave center-fed vertical with the bottom ⅛ λ above ground.

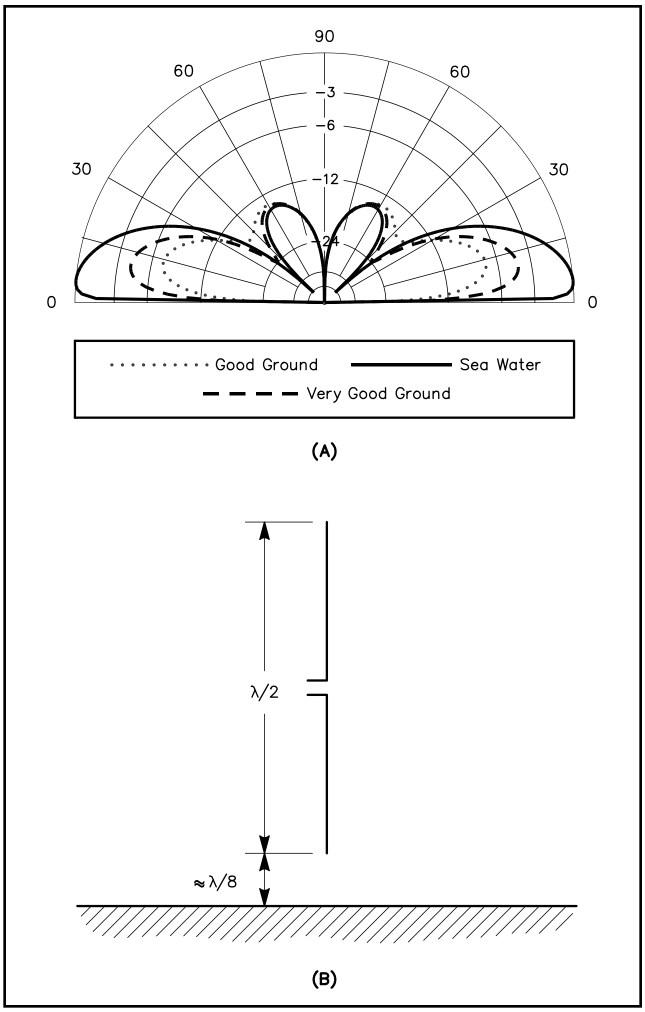


Fig 8-29 — At A, vertical radiation patterns of the half-wave vertical dipole with the bottom tip ⅛ λ off the ground, as shown at B.

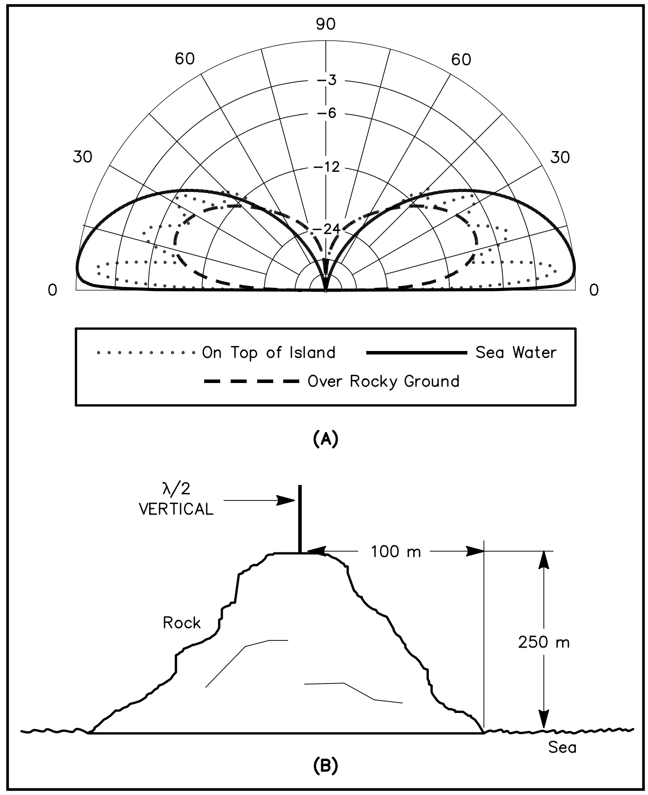


Fig 8-30 — At A, the modeled radiation pattern of a half-wave vertical overlooking a slope of very poor ground (an island with volcanic soil) next to the ocean, as shown at B. Because of the antenna height above the sea, multiple lobes show up in the pattern. The radiation patterns of the half-wave vertical at sea level and the pattern over very poor ground are superimposed for comparison.

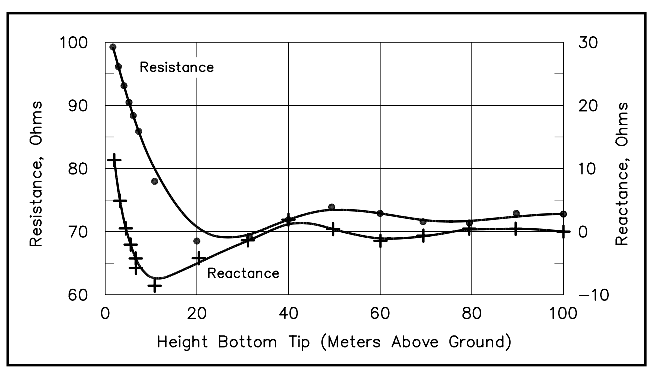


Fig 8-31 — Radiation resistance and reactance of the half-wave vertical as a function of height above ground. The height is taken as the height of the bottom tip. Calculations are for a design frequency of 3.5 MHz.

Note the secondary lobe, which is similar to the lobe we encountered with the horizontally polarized extended double Zepp.

I also modeled a half-wave vertical on top of a rocky island with very poor ground, 250 meters (820 feet) above sea level, and some 100 meters (330 feet) from the sea. **Fig 8-30** shows the layout and the radiation pattern. Superimposed on the pattern are the patterns for the same antenna at sea level, as well as over very poor ground. Note that the extra height does not give any gain advantage over sea level but the extra height does help low-angle rays shoot across the poor ground (the rocky island) and find reflection at sea level some 250 meters below the antenna.

5.2. Radiation Resistance

The radiation resistance of a vertical half-wave dipole, fed at the current maximum (the center of the dipole), is shown in **Fig 8-31** as a function of its height above ground. The impedance remains fairly constant except for very low heights. No current flows at the tips of the dipole, and hence the small influence of the height on the impedance, except at very low heights where the capacitive effect of the bottom of the antenna against ground lowers the resonant frequency of the antenna.

5.3. Feeding the Vertical Half-Wave Dipole

There are two main approaches to feeding a vertical half-wave dipole:

- Base feeding against ground (voltage feeding)
- Feeding in the center (current feeding)

Base feeding is covered in Section 4.4 of Chapter 6 on matching and feed lines. In most cases you will use a parallel tuned circuit on which the coax feed line is tapped. If the vertical is made using a sizable tower, the base impedance may be relatively low (600 Ω), and a broad-band matching system as described in Section. 4.5.2 in Chapter 6 on matching and feed lines (the W1FC broadband transformer) may be used.

A center-fed vertical dipole must be fed in the same way as a horizontal dipole. It represents a balanced feed point, and can be fed using open-wire line to a balanced tuner, or via a balun to a coaxial feed line (see Section 1.5).

Fig 8-32 — View from the top of I8UDB's tower in Naples. Such an awesome view needs no comment.

6. SLOPING DIPOLE

Sloping half-wave dipoles are used very successfully by a number of stations, especially near the sea. I8UDB is using a sloper on 160 from his mountaintop location near Naples, where electrical ground is nonexistent, but where the sea is only 100 meters away and a few hundred meters below the antenna. See **Fig 8-32**.

The half-wave sloper radiates a signal with both horizontal and vertical polarization components. Unless it is very high above the ground (such as at I8UDB), you need not bother with the horizontal component. Low-angle radiation will be produced only by the vertical component.

6.1. The Sloping Straight Dipole

Due to the weight of the feed line, a sloping dipole will seldom have two halves in a straight line. Let us nevertheless analyze the antenna as if it does.

Radiation patterns

All modeling in this section was done on 80 meters, over a very good ground. **Fig 8-33** shows the radiation patterns of sloping half-wave dipoles for apex angles of 15°, 30° and 45° over three types of ground (poor, good and sea). For the dipole

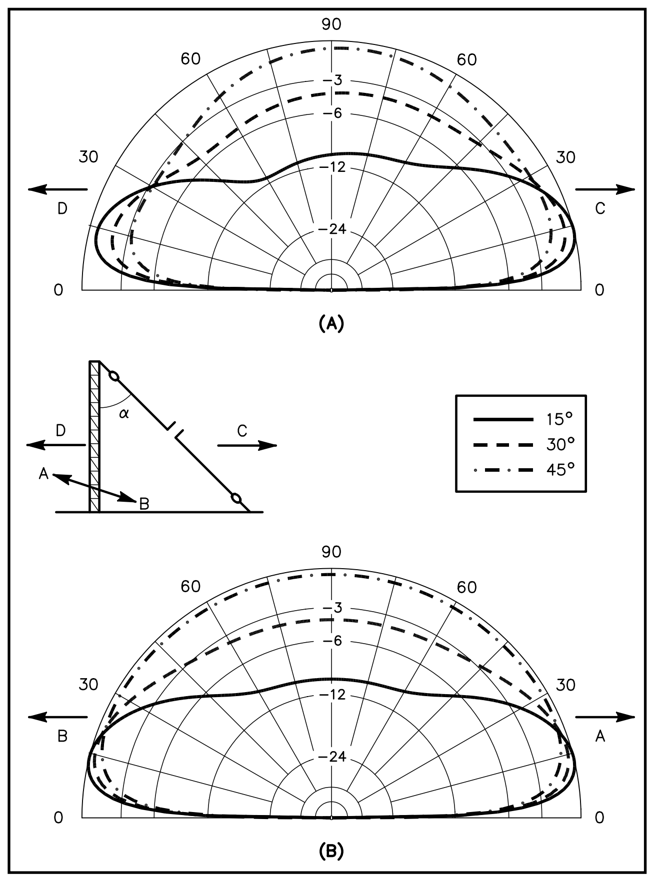


Fig 8-33 — Elevation-plane radiation patterns of sloping dipoles with various slope angles. At A, patterns in the plane of the sloper and its support (end-fire radiation), and at B, perpendicular to that plane (broadside radiation). End-fire radiation is 100% vertically polarized, while the broadside radiation contains a horizontal as well as vertical component. The horizontal pattern shows a very small amount of directivity.

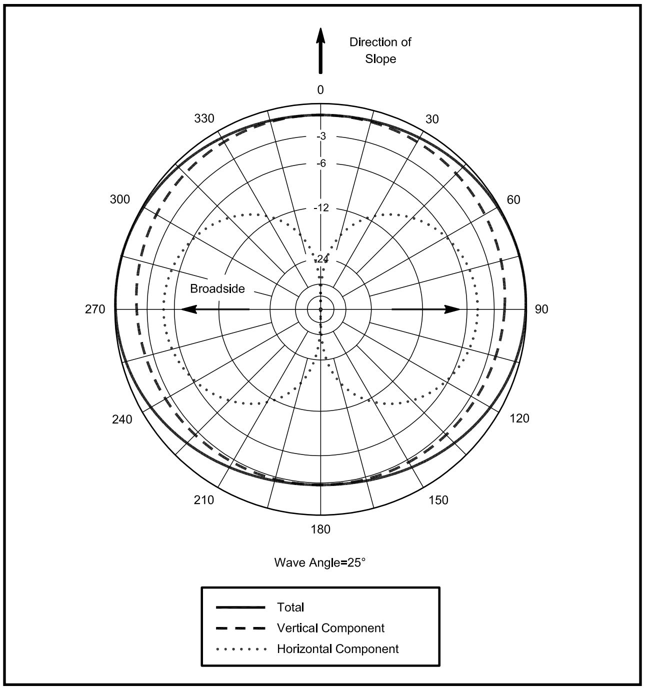


Fig 8-34 — Azimuth-plane radiation pattern for the sloping dipole with a 45° slope angle, taken at a 25° wave angle. Patterns for the vertical and horizontal components of the total are also shown. The directivity is very limited. Actually, the sloping dipole radiates best about 70° either side of the slope direction.

with a 45° slope angle I include the pattern showing the vertical and the horizontal radiation separately (**Fig 8-34**).

It is obvious that the steeper the slope, the smaller the horizontal radiation component will be. High-angle radiation is only due to the horizontal radiation component. For the vertical component the same rules apply as for the half-wave vertical: In order to exploit the intrinsic very low-angle capabilities, you must have an excellent ground around the antenna. Don't forget, the Fresnel zone (the area where the reflection at ground level takes place) can stretch all the way out to 10 wavelengths or more from the antenna.

Fig 8-34 shows the horizontal pattern for a sloping dipole with a 45° slope angle. The sloper is almost omnidirectional, but radiates best broadside (perpendicular to the plane going through the sloper and the support). In the end-fire direction (in the plane of the sloper and its support), it has less than 1 dB F/B at an elevation angle of 25°. The antenna radiates a little better in the direction of the slope. The fact that it radiates best in the broadside direction is due to the horizontal component, which only radiates in the broadside direction.

Impedance

The radiation resistance of the sloping dipole with the bottom wire approximately 1/80 λ above ground (1 meter for an 80-meter antenna) varies from 96 Ω for a 15° slope angle to 81 Ω for a 45° slope angle.

6.2. The Bent-Wire Sloping Dipole

Most real-life sloping half-wave dipoles have a bent-wire shape, because of the weight of the feed line. **Fig 8-35** and **Fig 8-36** analyze a sloping vertical with a slope angle of 20° for the top half of the antenna, and slope angles of 40° and 60° respectively for the bottom half of the dipole. Using a 60° slope angle reduces the height requirement for the support.

The sloping dipole with a relatively horizontal bottom quarter-wave wire yields almost the same signal as the straight sloping dipole. It is important to keep the top half of the sloping dipole as vertical as possible. Analysis shows the angle of the bottom half of the antenna is relatively unimportant.

Feed point

Is the feed point of such a bent sloping dipole a symmetrical feed point? Not strictly speaking. If you use such

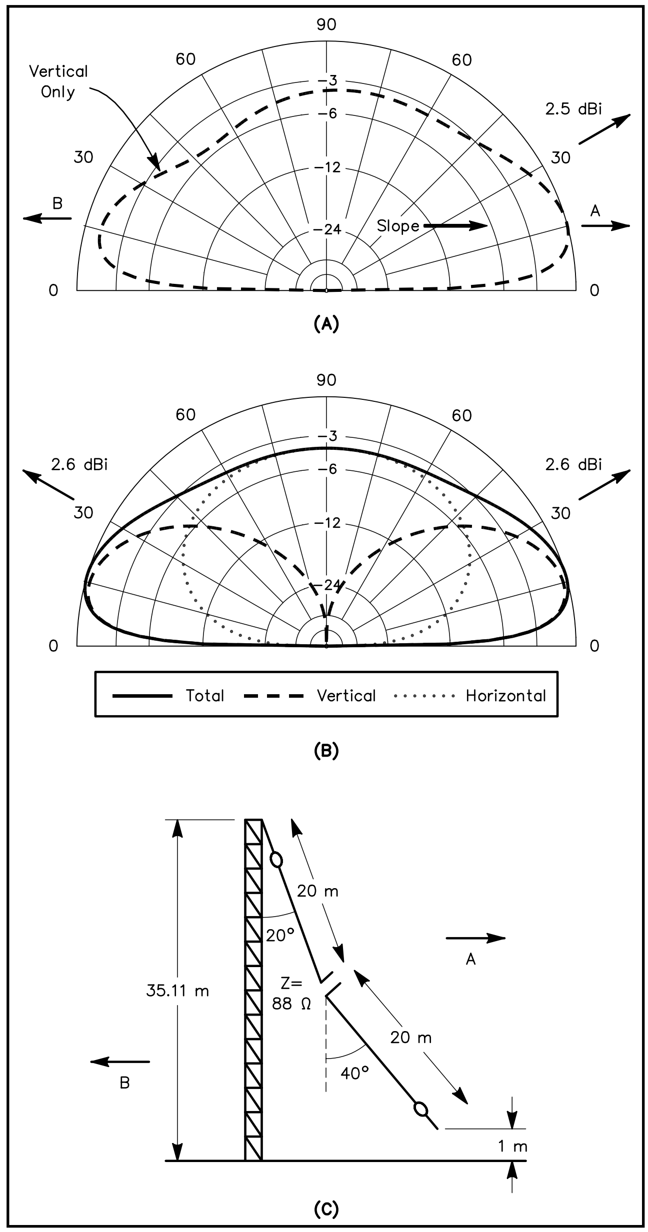


Fig 8-35 — At A, "end-fire" and at B, "broadside" vertical radiation patterns of a bent-wire half-wave sloper for 3.6 MHz. The horizontal and vertical components of the total pattern are also shown at B. The bottom 0.25-λ section slopes at an angle of 40°, as shown at C. Modeling is done over very good ground.

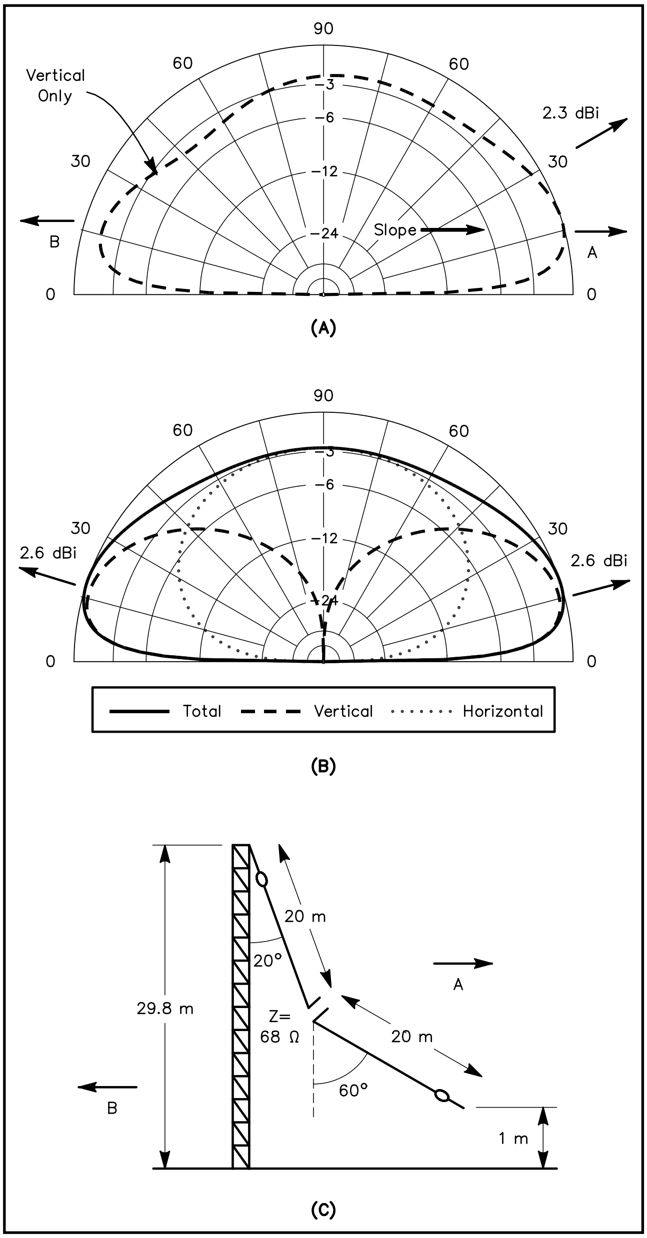


Fig 8-36 — At A, "end-fire" and at B, "broadside" vertical radiation patterns of a bent-wire half-wave sloper for 3.6 MHz. The horizontal and vertical components of the total pattern are also shown at B. The bottom 0.25-λ section slopes at an angle of 60°, as shown at C. This configuration and that of Fig 8-35 are just as valid as the configuration using a straight sloper. The loss in gain is negligible. This arrangement requires less support height than that of the straight sloper or that of Fig 8-35.

an antenna, don't take any chances. It does not hurt to put a current balun at a load even when the load is asymmetric. Use a current-type choke (common-mode) balun to remove any current from the outside of the coaxial cable. A coiled coax or a stack of ferrite beads is the way to go (see Section 1.5).

6.3. Evolution into the Quarter-Wave Vertical

We can go one step further and bring the bottom quarter-wave all the way horizontal. If the top half were fully vertical, we now would have a quarter-wave vertical with a single elevated radial. This configuration is described in detail in the Chapter 9 on vertical antennas (see Fig 9-18).

To transform the half-wave sloper into a quarter-wave vertical, more often called a ground-plane antenna, we first replace the sloping bottom half of the antenna with two wires, now called *radials*. Both radials are "in line" and slope toward the ground, as shown in **Fig 8-37C.** A and B of Fig 8-37 show the radiation patterns for this configuration. Note that the high-angle radiation has been attenuated some 10 dB, and we pick up 0.5 to 0.8 dB of gain. The little horizontally polarized radiation left over is, of course, caused by the sloping radials. The configuration shows gain in the direction of the sloping wire of approximately 0.4 dB.

Next we move the radials up, so they are horizontal, and move the antenna down so the base is now 5 meters (16 feet) above ground (Fig 8-37E). All the horizontal radiation is gone, and the gain has settled halfway between the forward and the backward gain of the previous model, which is to be expected. We now have a quarter-wave vertical with two radials, which is how the original ground plane was developed (see Section 1.3.3 of the chapter on verticals).

The quarter-wave vertical with two radials definitely has an asymmetrical feed point. The feed line is exposed to the strong fields of the antenna and often is run on the ground under the two radials. So you should fully decouple the feed line from the feed point by using a current-type balun (coiled coax or stack of ferrite beads).

6.4. Conclusion

Some 6.1 dBi gain can be obtained with a half-wave vertical only over nearly perfect ground (such as saltwater). Even over very good soil, the half-wave vertical will not be any better than a quarter-wave vertical (3 dBi gain). This means that unless you are near the sea, you may as well stick with a quarter-wave vertical. The sloping vertical (make the sloping wire as vertical as possible) with two radials (5 meters high for 3.6 MHz) will produce as good a signal as a half-wave vertical or sloping half-wave vertical over very good ground. It will, however, only require a 25-meter support instead of a 35- or 40-meter support for a half-wave 80-meter vertical dipole.

7. MODELING DIPOLES

MININEC can be used successfully for a large number of modeling aspects with dipoles. I would recommend using *NEC-2 (or NEC-4)*-based modeling programs however, especially when we deal with antennas relatively close to ground.

Straight dipoles can be modeled accurately with a total of 10 to 20 pulses. Inverted-V dipoles require more pulses, depending on the apex angle, to obtain accurate impedance data. **Table 8-8** shows impedance data for a straight dipole, and **Table 8-9** for an inverted-V dipole as a function of the pulses, wires and segments. An inverted-V with a 90° apex angle requires at least 50 equal-length segments for accurate impedance data. By using the tapering technique (see the chapter on Yagi and quad antennas), accurate results can be

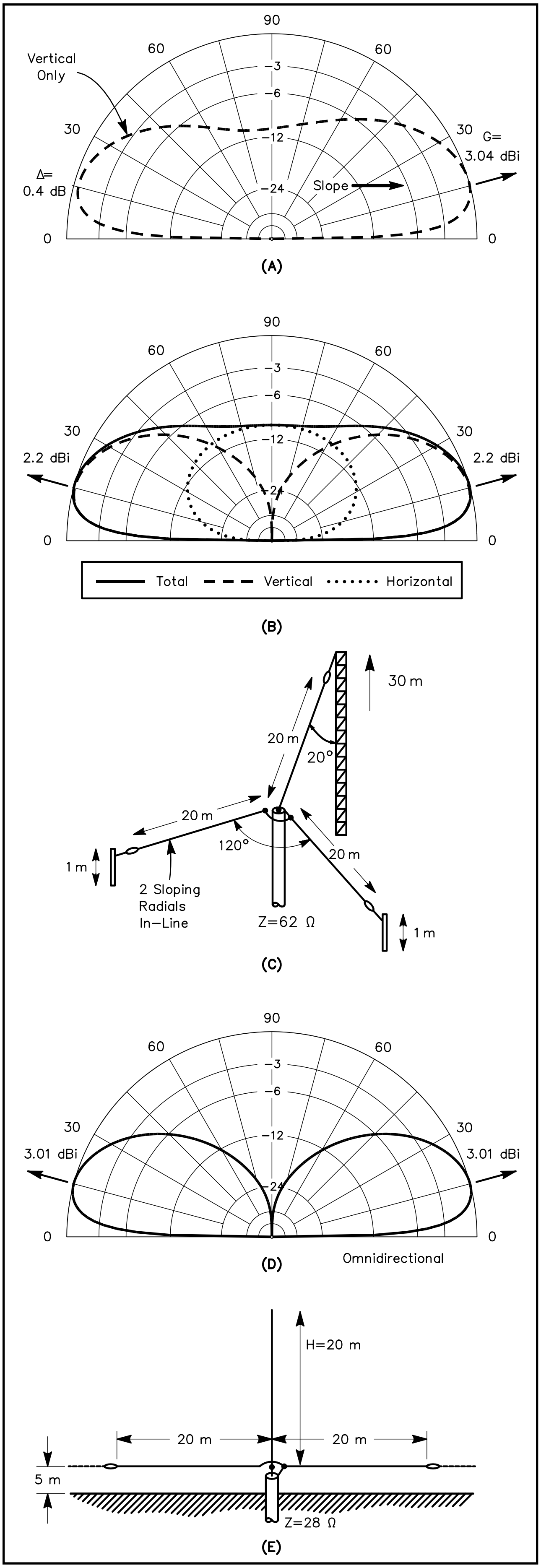


Fig 8-37 — Transition from a sloping dipole to a 0.25 λ vertical with two radials. At C, the bottom half of the dipole is replaced by two 0.25 λ wires sloping to the ground; the resulting patterns are shown at A and B. At E the radials are lifted to the horizontal, with the resulting pattern at D. This changes eliminates all the horizontal radiation component that was originated by the sloping wires. Analysis frequency: 3.65 MHz.

Table 8-8
***MININEC* Pulses Versus Calculated Impedance for a Straight Dipole Antenna**

Pulses	*Impedance*
5	71 – *j* l4
10	67 – *j* 26
20	68 – *j* 28
30	68.5 – *j* 28
50	68.6 – *j* 27.3
80	68.7 – *j* 27.1
100	68.7 – *j* 27.0

Table 8-9
***MININEC* Pulses Vs Calculated Impedance for an Inverted-V Dipole Antenna**

Pulses	*Impedance*
5	43.6 – *j* 23.7
10	44.3 + *j* 10.3
20	44.6 + *j* 28.4
30	44.6 + *j* 34.3
50	44.7 + *j* 38.1
80	44.7 + *j* 39.8
100	44.8 + *j* 42.0
20 tapered, min 0.4 m, max 3.0 m	44.2 + *j* 36.1
26 tapered, min 0.3 m, max 2.0 m	44.4 + *j* 37.6
26 tapered, min 0.4 m, max 2.0 m	44.4 + *j* 36.3
28 tapered, min 0.4 m, max 2.0 m	44.4 + *j* 38.9
46 tapered, min 0.2 m, max 1.0 m	44.2 + *j* 40

obtained with a total of only 26 segments. *EZNEC* provides an automatic feature for generating tapered segment lengths, which is a great asset when you model antennas with bent conductors.

Knowing the exact impedance is important only if you want to calculate the exact resonant length (or frequency) of a dipole, or if the dipole is part of an array. To obtain reliable results using a *MININEC*-based program the dipoles should not be modeled too close to ground. For half-wave horizontal dipoles, the antenna should be at least 0.2 λ high. For longer dipoles, the minimum height ensuring reliable results is somewhat higher. Vertical dipoles and sloping dipoles (with a steep slope angle) can be modeled quite close to the ground, as there is very little radiation in the near-field toward the ground (a dipole does not radiate off its tips).

A *NEC-2 or NEC-4*-based modeling program is required if accurate gain and impedance data are required for dipoles close to ground.

Contents

CHAPTER 9

Vertical Antennas

Uli Weiss, DJ2YA, is an all-around radio amateur. His more-than-casual interest and in-depth knowledge of antenna matters and his eminent knowledge of the English language (Uli teaches English at a German "Gymnasium") has made him one of the few persons who could successfully translate the *Low Band DXing* book into the German language without any assistance from the author. It also makes him a very successful antenna builder and contest operator. Uli was, with Walter Skudlarek, DJ6QT, cofounder of the world-renowned RRDXA Contest Club, which has led the CQ World Wide DX Contest club championships for many years. Uli has been an editor, advisor, helping hand and supporter for this chapter on vertical antennas, and that for several editions of this book.

Thank you for your help, Uli.

The effects of the earth itself and the artificial ground system (if used) on the radiation pattern and the efficiency of vertically polarized antennas is often not understood. The influence of the ground and the ground system on a vertical antenna is twofold.

Near the antenna (in the *near field*), you need a good ground system to collect the antenna return currents without losses. This will determine the *radiation efficiency* of the antenna.

At distances farther away (in the *far field*, also called the *Fresnel zone*), the wave is reflected from the earth and combines with the direct wave to generate the overall radiation pattern. The absorption of the reflected wave is a function of the ground quality and the incident angle. This mechanism determines the *reflection efficiency* of the antenna.

Vertical monopole antennas are often called *ground-mounted verticals*, or simply verticals. They are, by definition, mounted perpendicular to the earth, and they produce a vertically polarized signal. Verticals are popular antennas for the low bands, since they can produce good low-angle radiation without the very high supports needed for horizontally polarized antennas to produce the same amount radiation at low takeoff angles.

1. THE QUARTER-WAVE VERTICAL

1.1. Radiation Patterns

1.1.1. Vertical Pattern of Vertical Monopoles Over Ideal Ground

The radiation pattern produced by a ground-mounted quarter-wave vertical antenna is basically one-half that of a half-wave dipole antenna in free space. The dipole is twice the physical size of the vertical and has a symmetrical current distribution. A vertical antenna is frequently referred to as a "monopole" to distinguish it from a dipole. The radiation pattern of a quarter-wave vertical monopole over perfect ground is half of the figure-8 shown for the half-wave dipole in free space. See **Fig 9-1**.

The relative field strength of a vertical antenna with sinusoidal current distribution and a current node at the top is given by:

$$E_f = k \times I \left[\frac{\cos (L \sin \alpha) - \cos L}{\cos \alpha} \right] \quad \text{(Eq 9-1)}$$

where
k = constant related to impedance
E_f = relative field strength
α = elevation angle above the horizon
L = electrical length (height) of the antenna
I = antenna current

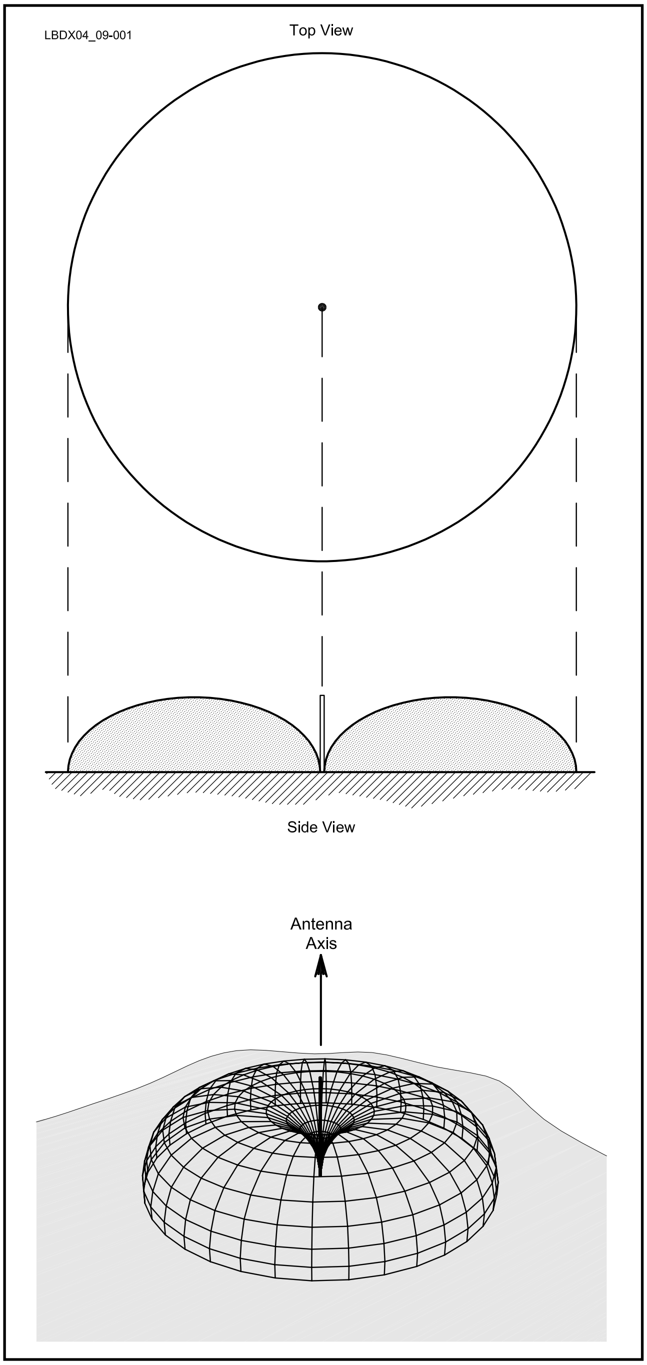


Fig 9-1 — The radiation patterns produced by a vertical monopole over perfect ground. The top view is the horizontal pattern, and the side view is the vertical (elevation plane) pattern.

This equation does not take imperfect ground conditions into account, and is valid for antenna heights between 0° and 180° (0 to λ/2). The "form factor" inside the square brackets containing the trigonometric functions is often published by itself for use in calculating the field strength of a vertical antenna. If used in this way, however, it appears that short verticals are vastly inferior to tall ones, since the antenna length appears only in the numerator of the fraction.

Replacing the current I in the equation with the term

$$\sqrt{\frac{P}{R_{rad} + R_{loss}}}$$

gives a better picture of the actual situation. For short verticals, the value of the radiation resistance is small, and this term largely compensates for the decrease in the form factor. This means that for a constant power input, the current into a small vertical will be greater than for a larger monopole.

The radiation resistance R_{rad} does not determine the current — the sum of the radiation resistance and the loss resistance(s) does. With a less-than-perfect ground system and short, less-than-perfect loading elements (lossy coils used with short verticals), the radiation can be significantly less than the case of a larger vertical (where R_{rad} is large in comparison to the ground loss and where there are no lossy loading devices).

Interestingly, short verticals are almost as efficient radiators as are longer verticals, provided the ground system is good and there are no lossy loading devices. When the losses of the ground system and the loading devices are brought into the picture, however, the sum $R_{rad} + R_{loss}$ will get larger, and as a result part of the supplied power will be lost in the form of heat in these elements. For instance, if $R_{rad} = R_{loss}$, half of the power will be lost. Note that with very short verticals, these losses can be much higher.

1.1.2. Vertical Radiation Pattern of a Monopole Over Real Ground

The three-dimensional radiation pattern from an antenna is made up of the combination of the direct wave and the wave resulting from reflection from the earth. The following explanation is valid only for reflection of vertically polarized waves. See Chapter 8 on dipole antennas for an explanation of the reflection mechanism for horizontally polarized waves.

For perfect earth there is no phase shift of the vertically polarized wave at the reflection point. The two waves add with a certain phase difference, due only to the different path lengths. This is the mechanism that creates the radiation pattern. Consider a distant point at a very low angle to the horizon. Since the path lengths are almost the same, reinforcement of the direct and reflected waves will be maximum. In case of a perfect ground, the radiation will be maximum just above a 0° elevation angle.

1.1.2.1. The Reflection Coefficient

Over real earth, reflection causes both amplitude and phase changes. The reflection coefficient describes how the incident (vertically polarized) wave is being reflected. The reflection coefficient of real earth is a complex number (such as $\angle\alpha°$)

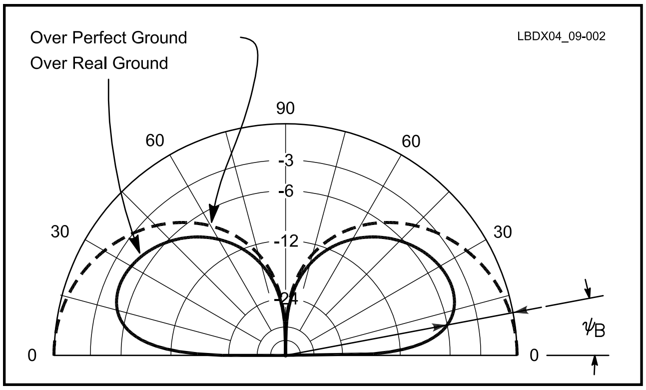


Fig 9-2 — Vertical radiation patterns of a λ/4 monopole over perfect and imperfect earth. The pseudo-Brewster angle is the radiation angle at which the real-ground pattern is 6 dB down from the perfect-ground pattern.

with magnitude and phase, and it varies with frequency. In the polar-coordinate system the reflection coefficient consists of:

- The magnitude of the reflection coefficient (a): It determines how much power is being reflected, and what percentage is being absorbed in the lossy ground. A figure of 0.6 means that 60% will be reflected and 40% absorbed.
- The phase angle (α): This is the phase shift that the reflected wave will undergo as compared to the incident wave. Over real earth the phase is always lagging (minus sign). At a 0° elevation angle, the phase is always –180°. This causes the total radiation to be zero (the incident and reflected waves, which are 180° out-of-phase and equal in magnitude, cancel each other). At higher elevation angles, the reflection phase angle will be close to zero (typically –5° to –15°, depending on the ground quality).

1.1.2.2. The Pseudo-Brewster Angle

The magnitude of the vertical reflection coefficient is minimum at a 90° phase angle. This is the reflection-coefficient phase angle at which the so-called *pseudo-Brewster wave* angle occurs. It is called the pseudo-Brewster angle because the RF effect is similar to the optical effect from which the term gets its name. At the pseudo-Brewster angle the reflected wave changes sign. Below the pseudo-Brewster angle the reflected wave will subtract from the direct wave. Above the pseudo-Brewster angle it adds to the direct wave. At the pseudo-Brewster angle the radiation is 6 dB down from the perfect ground pattern (see **Fig 9-2**).

All this should make it clear that knowing the pseudo-Brewster angle is important for each band at a given QTH. Most of us use a vertical to achieve good low-angle radiation.

Fig 9-3 shows the reflection coefficient (magnitude and phase) for 3.6 MHz and 1.8 MHz for three types of ground. Over seawater the reflection-coefficient phase angle changes from –180° at a 0° wave angle to –0.1° at less than 0.5° wave angle! The pseudo-Brewster angle is at approximately 0.2° over saltwater.

1.1.2.3. Ground-Quality Characterization

Ground quality is defined by two parameters: the dielectric constant and the conductivity, expressed in milliSiemens per meter (mS/m). Table 5-2 in Chapter 5 shows the characterization of various real-ground types. The table also shows five distinct types of ground, labeled as very good, average, poor, very poor and extremely poor. These come from Terman's classic *Radio Engineers' Handbook*, and are also used by Lewallen in his *EZNEC* modeling programs. The denominations and values listed in Table 5-2 are the standard ground types used throughout this book for modeling radiation patterns. In the real world, ground characteristics are never homogeneous, and extremely wide variations over short distances are common. Therefore any modeling results based on homogeneous ground

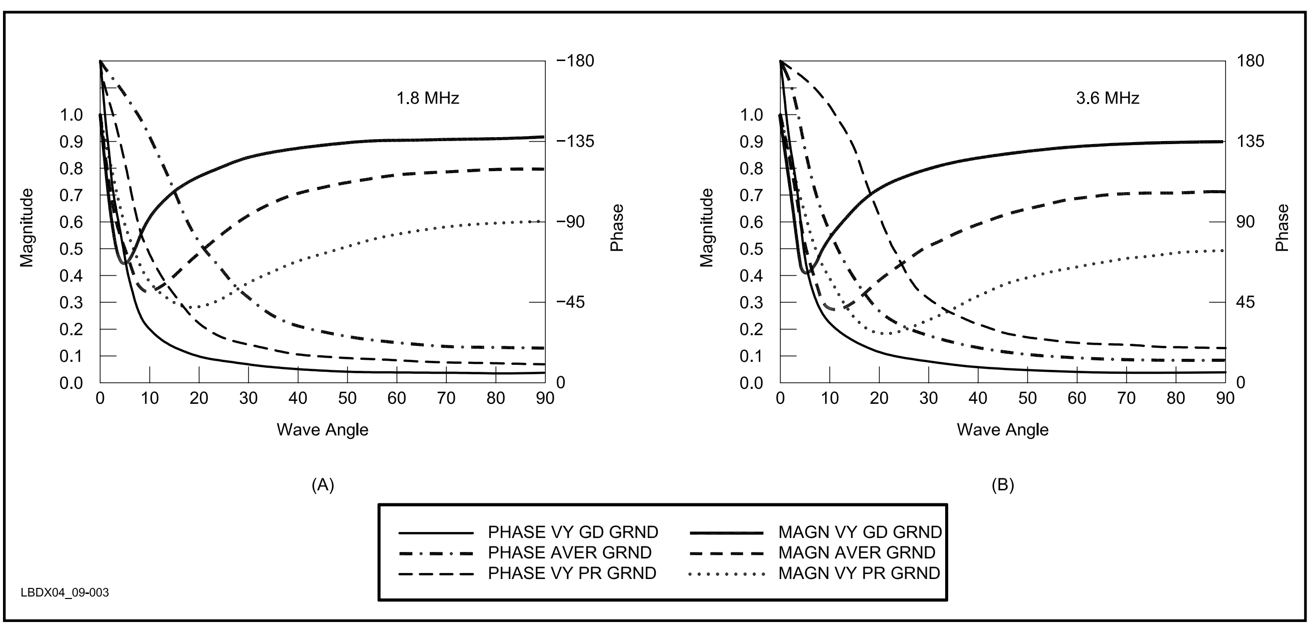


Fig 9-3 — Reflection coefficient (magnitude and phase) for vertically polarized waves over three different types of ground (very good, average and very poor).

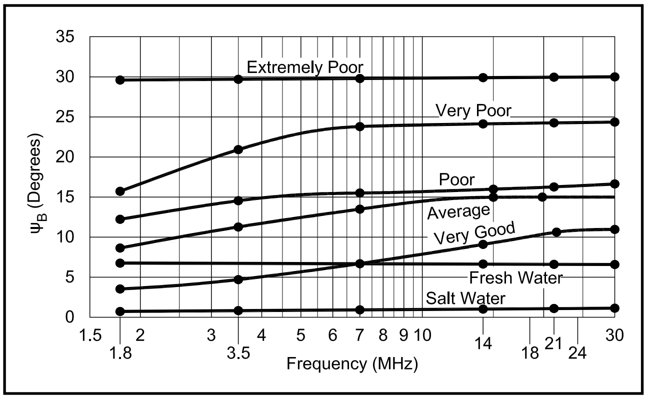


Fig 9-4 — Pseudo-Brewster angle for different qualities of reflecting ground. Note that over saltwater the pseudo-Brewster angle is constant for all frequencies, at less than 0.1°! That's why vertical antennas located right at the saltwater shore get out so well.

characteristics will only be as accurate as the homogeneity of the ground itself.

1.1.2.4. Brewster Angle Equation

Terman (*Radio Engineers' Handbook*) publishes an equation that gives the pseudo-Brewster angle as a function of the ground permeability, the conductivity and the frequency. The chart in **Fig 9-4** uses the Terman equation. Note especially how saltwater has a dramatic influence on the low-angle radiation performance of verticals. In contrast, a sandy, dry ground yields a pseudo-Brewster angle of 13° to 15° on the low bands, and a city (heavy industrial) ground yields a pseudo-Brewster angle of nearly 30° on all frequencies! This means that under such circumstances the radiation efficiency for angles under 30° will be severely degraded in a city environment.

1.1.2.5. Brewster Angle and Radials

Is there anything you can do about the pseudo-Brewster angle? Very little. Ground-radial systems are commonly used to reduce the losses in the near field of a vertical antenna. These ground-radial systems are usually 0.1 to 0.5 λ long, too short to improve the earth conditions in the area where reflection near the pseudo-Brewster angle takes place.

For quarter-wave verticals the Fresnel zone (the zone where the reflection takes place) is 1 to 2 λ away from the antenna. For longer verticals (such as a half-wave vertical) the Fresnel zone extends up to 100 wavelengths away from the antenna (for an elevation angle of about 0.25°).

This means that a good radial system improves the efficiency of the vertical in collecting return currents and shielding from lossy ground, but will not influence the radiation by improving the reflection mechanism in the Fresnel zone. Of course you could add 5 λ long radials, and keep the far ends of these radials less than 0.05 λ apart by using enough radials. But that seems rather impractical for most of us! In most practical cases radiation at low takeoff angles will be determined only by the real ground around the vertical antenna.

Conclusion

This information should make it clear that a vertical may not be the best antenna if you are living in an area with very poor ground characteristics. This has been widely confirmed in real life. Many top-notch DXers living in the Sonoran desert or in mountainous rocky areas on the US West Coast swear by

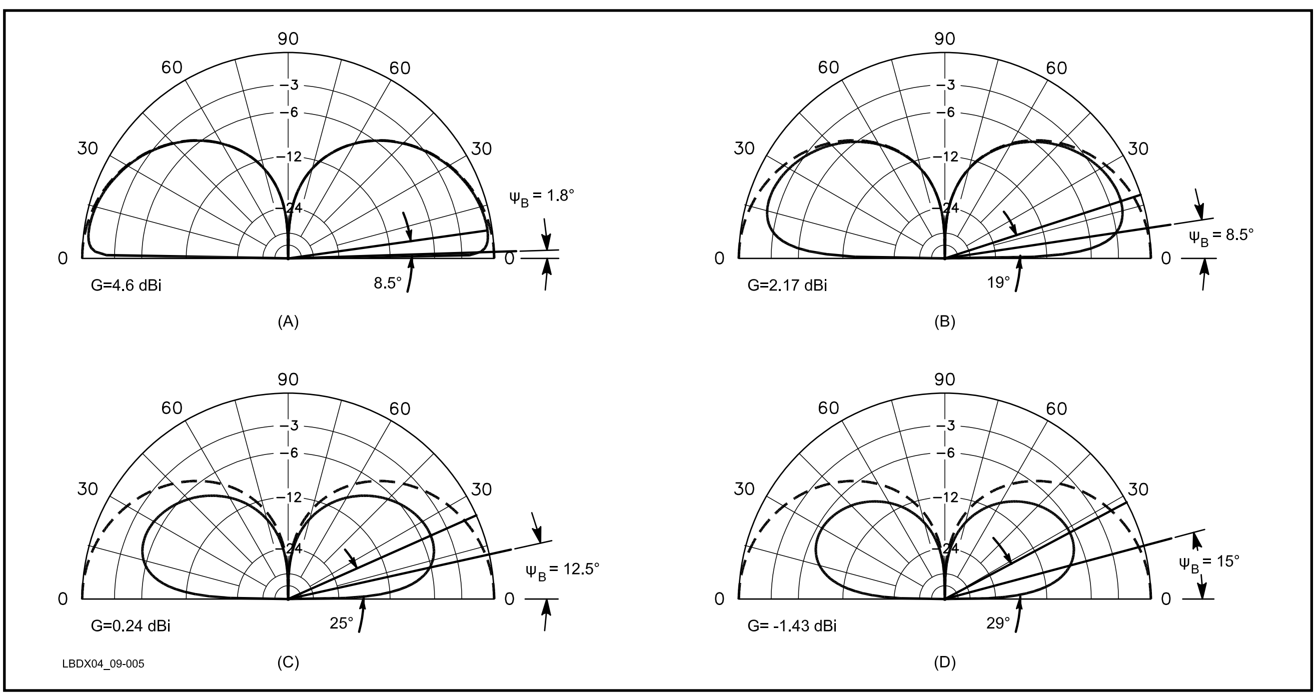


Fig 9-5 — Vertical-plane radiation patterns of 80-meter λ/4 verticals over four standard types of ground. At A, over saltwater. At B, over very good ground. At C, over average ground. At D, over very poor ground. In each case using 64 radials, each 20 meters long. The perfect ground pattern is shown in each pattern as a reference (broken line, with a gain of 5.0 dBi). This reference pattern also allows us to calculate the pseudo-Brewster angle. Modeling was done by N6BV using *NEC-4*. All patterns are all plotted on the same scale.

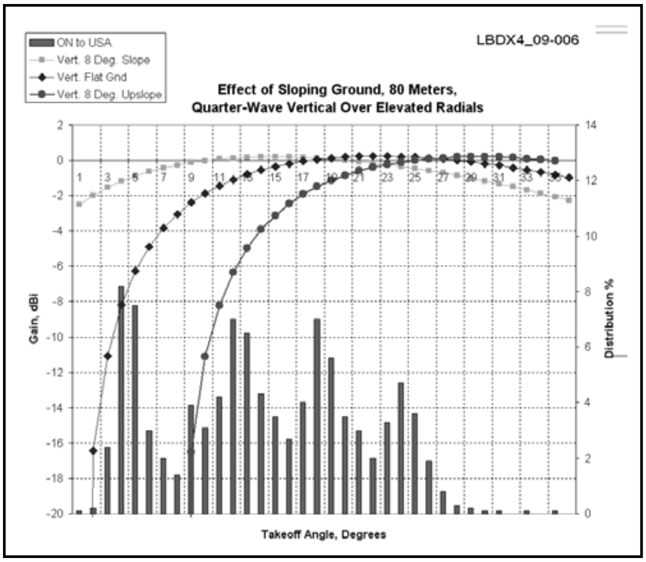


Fig 9-6 — The bar graph represents the distribution of the wave angles encountered on 80 meters on a Europe to USA path. Modeling was done over good ground. The wave angles are shown for a λ/4 vertical over flat ground, over an uphill slope of 8° and over a downhill slope of 8°. The downhill slope is very helpful when it comes to very low angles.

horizontal antennas for the low bands, at least on 80 meters, while some of their colleagues living in flat areas with rich fertile soil, or even better, on such a ground near the seacoast, will be enthusiastic advocates for vertical antennas and arrays made of vertical antennas.

On Top Band another mechanism enters into the game — the effect of power coupling (see Chapter 1, Section 3.5), which makes a vertically polarized antenna the better antenna in most places away from the equator (such as North America and Europe) due to the influence of the Earth's magnetic field. In addition, horizontally polarized antennas producing a low radiation angle on 160 meters are out of reach for all but a few who have antenna supports that are several hundred meters high!

1.1.2.6. Vertical Radiation Patterns

It is important to understand that gain and directivity are two different things. A vertical antenna over poor ground may show a good wave angle for DX, but its gain may be poor. The difference in gain at a 10° wave angle for a quarter-wave vertical over very poor ground, as compared to the same vertical over seawater, is an impressive 6 dB. **Fig 9-5** shows the vertical-plane radiation pattern of a quarter-wave vertical over four types of "real" ground:

- Seawater
- Excellent ground
- Average ground
- Extremely poor ground

1.1.2.7. Vertical Radiation Patterns Over Sloping Grounds

So far all our discussions about radiation patterns assumed we have perfectly homogeneous flat ground stretching for tens of wavelengths around the antenna. In Section 1.1.2 of Chapter 5, I discussed the influence of sloping terrain on vertical radiation patterns of antennas on the low bands.

Fig 9-6 shows that a terrain that slopes downhill in the direction of the target is as helpful for vertical antennas as it is for horizontally polarized antennas. On the other hand, an upward-sloping terrain works the other way!

1.1.2.8. Horizontal Pattern of a Vertical Monopole

The horizontal radiation pattern of both the ground-mounted monopole and the vertical dipole is a circle.

1.2. Radiation Resistance of Monopoles

The IRE definition of radiation resistance says that radiation resistance is the total power radiated as electromagnetic radiation, divided by the *net* current causing that radiation.

The radiation resistance value of any antenna depends on where it is fed. I'll call the radiation resistance of a vertical antenna at a point of current maximum as $R_{rad(I)}$ and the radiation resistance of a vertical antenna when fed at its base as $R_{rad(B)}$. For verticals greater than one quarter-wave in height, these two are not the same. Why is it important to know the radiation resistance of our vertical? The information is required to calculate the efficiency of the vertical:

$$\text{Eff} = \frac{R_{rad}}{R_{rad} + R_{loss}}$$

The radiation resistance of the antenna plus the loss resistance R_{loss} is the resistive part of the feed-point impedance

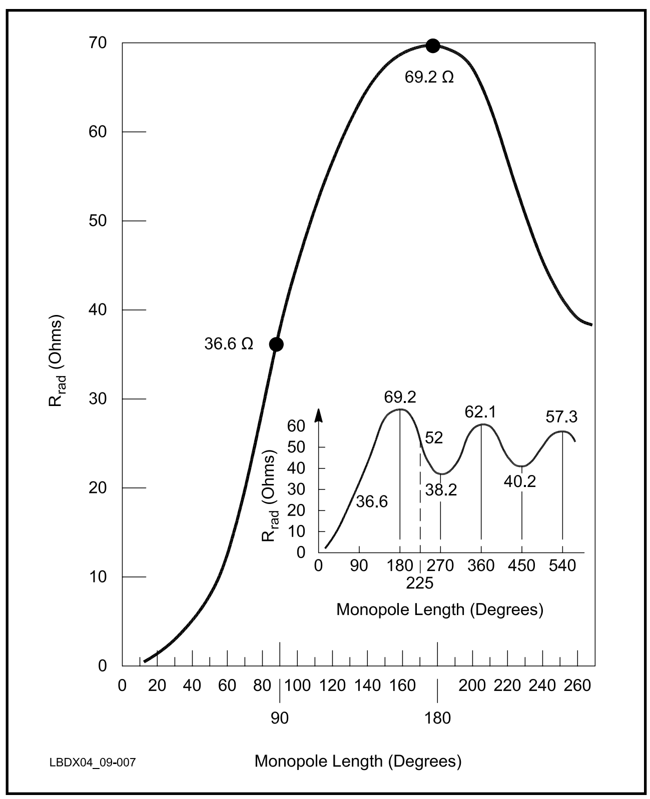


Fig 9-7 — Radiation resistances (at the current maximum) of monopoles with sinusoidal current distribution. The chart can also be used for dipoles, but all values must be doubled.

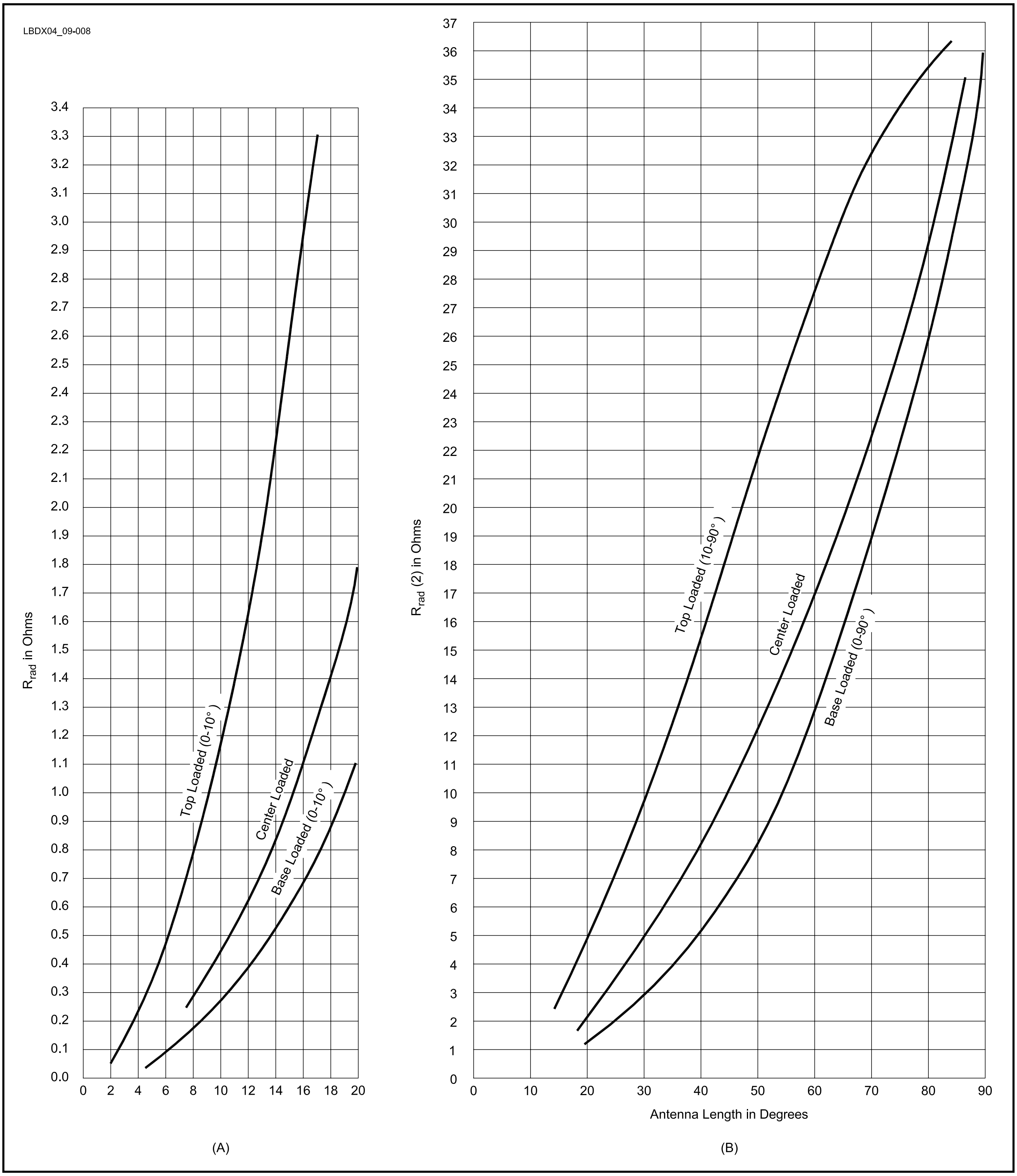


Fig 9-8 — Radiation resistance (R_{rad}) charts for verticals up to 90° or λ/4 long. At A, for lengths up to 20°, and at B, for greater lengths.

of the vertical. The feed-point resistance (and reactance) is required to design an appropriate matching network between the antenna and the feed line.

Fig 9-7 shows $R_{rad(I)}$ of verticals ranging in electrical height from 20° to 540°. (This is the radiation resistance referred to the current maximum.) The radiation resistance of a vertical shorter than or equal to a quarter wavelength and fed at its base [thus $R_{rad(I)} = R_{rad(B)}$] can be calculated as follows:

$$R_{rad} = \frac{1450\ h^2}{\lambda^2} \qquad \text{(Eq 9-2)}$$

where

h = effective antenna height in meters

λ = wavelength of operation, meters (= 300/f_{MHz})

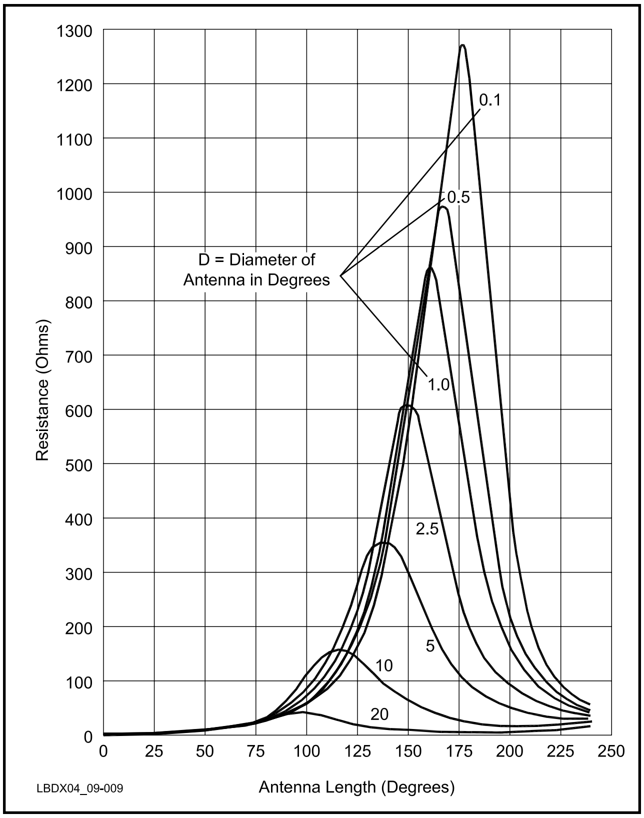


Fig 9-9 — Radiation resistances for monopoles fed at the base. Curves are given for various conductor (tower) diameters. The values are valid for perfect ground only.

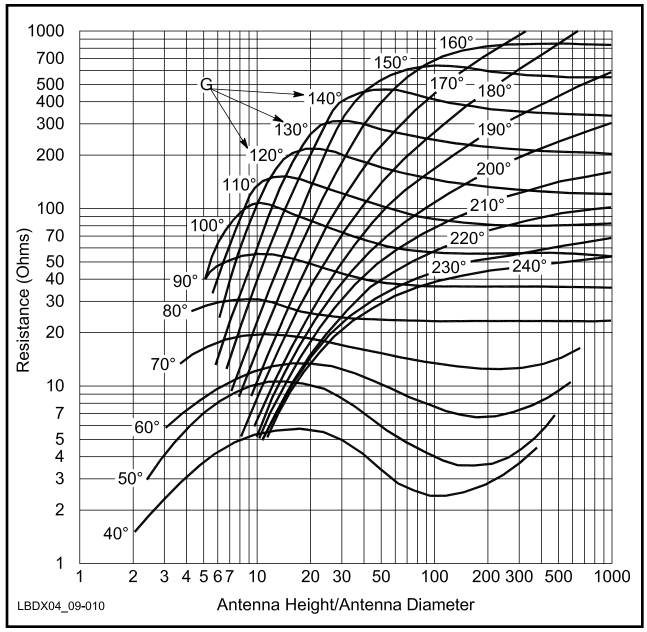


Fig 9-10 — Radiation resistances for monopoles fed at the base. Curves are given for various height/diameter ratios over perfect ground.

The *effective height* of the antenna is the height of a theoretical antenna having a constant current distribution all along its length. The area under this current distribution line is equal to the area under the current distribution line of the "real" antenna. Equation 9-2 is valid for antennas with a ratio of antenna length to conductor diameter of greater than 500:1 (typical for wire antennas).

For a full-size, quarter-wave antenna the radiation resistance is determined by:

- Current at the base of the antenna = 1 A (given)
- Area under sinusoidal current-distribution curve = 1 A × 1 radian = 1 A ×180/π = 57.3 A-degrees
- Equivalent length = 57.3° (1 radian)
- Full electrical wavelength = 300/3.8 = 78.95 meters
- Effective height = (78.95 × 57.3)/360 = 12.56°

$$R_{rad} = \frac{1450 \times 12.56^2}{78.95^2} = 36.6\ \Omega$$

Fig 9-8 shows the radiation resistance for a short vertical (valid for antennas with diameters ranging from 0.1° to 1°). For antennas made of thicker elements, **Fig 9-9** and **Fig 9-10** can be used. These charts are for antennas with a constant diameter. The same procedure can be used for calculating the radiation resistance of various types of short verticals.

For verticals with a *tapering diameter*, large deviations have been observed. W. J. Schultz describes a method for calculating the input impedance of a tapered vertical (Ref 795). It has also been reported that verticals with a large diameter exhibit a much lower radiation resistance than the standard 36.6 Ω value. A. Doty, K8CFU, reports finding values as low as 21 Ω during his extensive experiments on elevated radial systems (Ref 793). I have measured a similar low value on my quarter-wave 160-meter vertical (see Section 6.6.)

Vertical antennas longer than ¼ λ are usually not fed at the current maximum, but rather at the antenna base, so that $R_{rad(I)}$ is no longer the same as $R_{rad(B)}$ for long verticals in Figs 9-9 and 9-10. (Source: Henney, *Radio Engineering Handbook*, McGraw-Hill, NY, 1959, used with permission.) $R_{rad(I)}$ is illustrated in **Fig 9-11**. The value can be calculated from the following formula (Ref 722):

$$R_{rad(I)} = \varepsilon - 0.7\,L + 0.1\,\left[20 \sin\left(12.56637L - 4.08407\right)\right] + 45 \qquad \text{(Eq 9-3)}$$

where

ε = the base for natural logarithms, 2.71828.

L = antenna length in radians (radians = degrees × π/180° = degrees divided by 57.296). The length must be greater than π/2 radians (90°).

Fig 9-11C shows the case of a 135° (3λ/4) antenna. Disregarding losses, $R_{rad(B)} = R_{feed} \approx 300\ \Omega$, but the value of 2R, the theoretical resistance at the maximum current point, will be lower (57 Ω). If P1 (radiated power) = P2 (power dissipated in 2R), then $R_{rad(I)} = 2R$.

These values of $R_{rad(I)}$ are given in Fig 9-6, while $R_{rad(B)}$ can be found in Figs 9-8 and 9-9. **Fig 9-12** and **Fig 9-13** show the reactance of monopoles (at the base feed point) for varying antenna lengths and antenna diameters (Source: E. A. Laport, *Radio Antenna Engineering*, McGraw-Hill, NY, 1952, used by permission.).

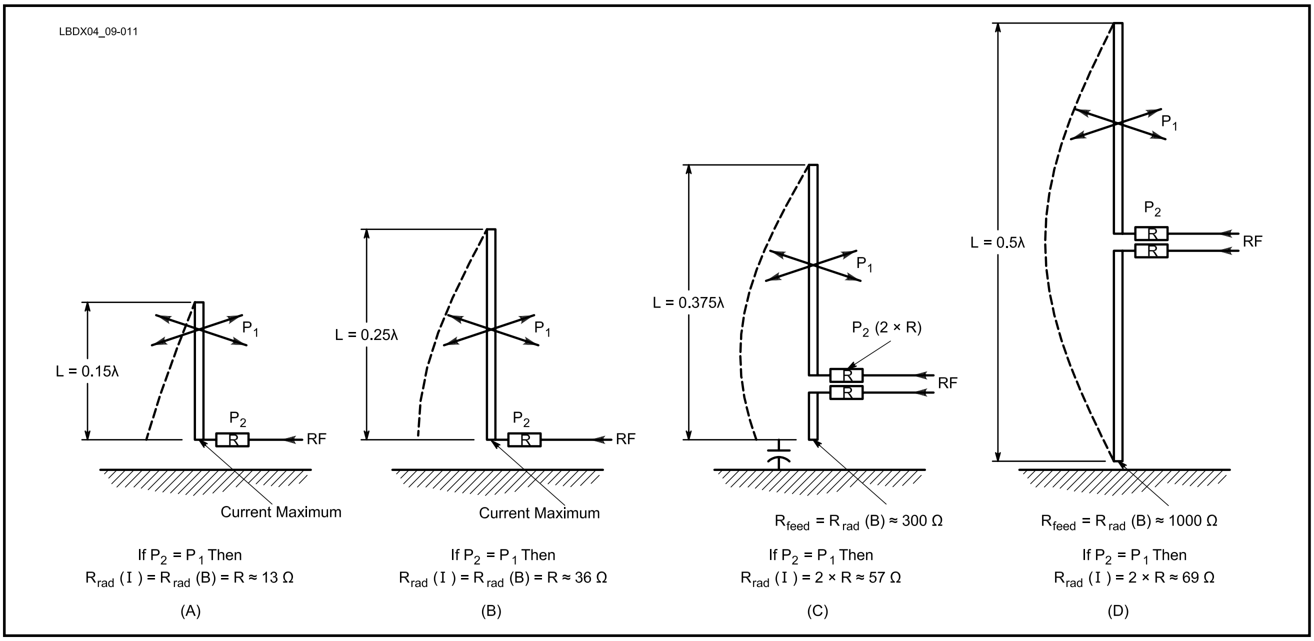
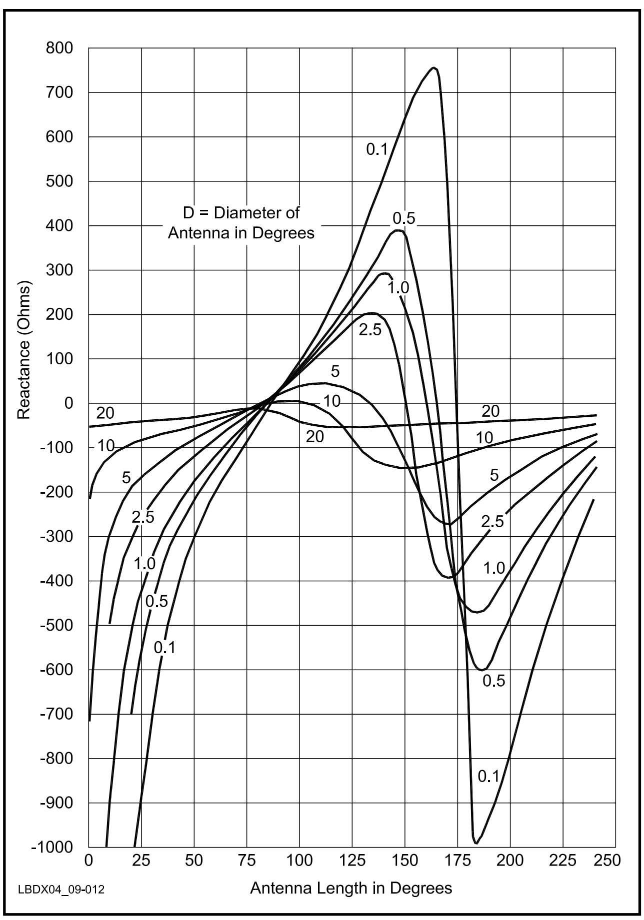


Fig 9-11 — Radiation resistance terminology for long and short verticals. See text for details. The feed point resistances indicated assume no losses.

Fig 9-12 — Feed-point reactances (over perfect ground) for monopoles with varying diameters.



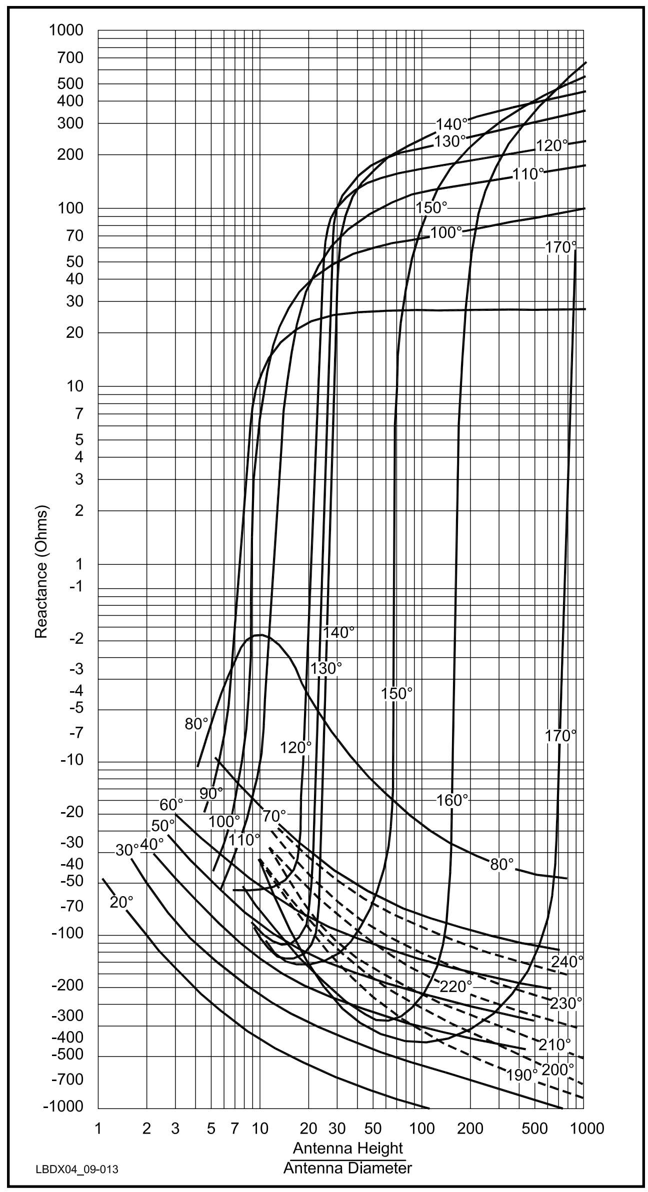


Fig 9-13 — Feed-point reactances (over perfect ground) for monopoles with different height/diameter ratios.

1.3. Radiation Efficiency of the Monopole Antenna

The radiation efficiency for short verticals has been defined as

$$\mathrm{Eff} = \frac{R_{rad}}{R_{rad} + R_{loss}}$$

For the case of any vertical, short or long, when fed at its base this equation becomes

$$\mathrm{Eff} = \frac{R_{rad}}{R_{rad(B)} + R_{loss}} \qquad \text{(Eq 9-4)}$$

The loss resistance of a vertical is composed of:

- Conductor RF resistance
- Parallel losses from insulators
- Equivalent series losses of the loading element(s)
- Ground losses part of the antenna current return circuit
- Ground absorption in the near field

1.3.1. Conductor RF Resistance

When multi-section towers are used for a vertical antenna, care should be taken to ensure proper electrical contact between the sections. If necessary, a copper braid strap should interconnect the sections. Rohrbacher, DJ2NN, provided a formula to calculate the effective RF resistance of conductors of copper, aluminum and bronze:

$$R_{loss} = (1 + 0.1\,L)\left(f^{0.125}\right)\left(0.5\frac{1.5}{D}\right) \times M \qquad \text{(Eq 9-5)}$$

where

L = length of the vertical in meters
f = frequency of operation in MHz
D = conductor diameter in mm
M = material constant (M = 0.945 for copper, 1.0 for bronze, and 1.16 for aluminum)

1.3.2. Parallel Losses in Insulators

Base insulators often operate at low-impedance points. For monopoles near a half-wavelength long, however, care should be taken to use high-quality insulators, since very high voltages can be present. There are many military surplus insulators available for this purpose. For medium and low-impedance applications, insulators made of nylon stock (turned down to the appropriate diameter) are excellent, but a good old glass Coke bottle may do just as well!

1.3.3. Ground Losses

Efficiency means: How many of the watts I deliver to the antenna are radiated as RF. *Effectiveness* means: Is the RF radiated where I want it? That is, at the right elevation angle and in the right direction. Your antenna can be very efficient but at the same time be very ineffective. Even the opposite is possible (killing a mouse with a cruise missile).

A large number of articles have been published in the literature concerning ground systems for verticals. The ground plays an important role in determining the efficiency as well as effectiveness of a vertical in two very distinct areas: the near field and the far field. Losses in the near field are losses causing the radiation efficiency to be less than 100%.

- **I^2R losses**: Antenna return currents travel through the ground, and back to the feed point, right at the base of the antenna (see Fig 9-42 later in this chapter). The resistivity of the ground will play an important role if these antenna RF return currents travel through the (lossy) ground. Unless the vertical antenna uses elevated radials, the antenna return current will flow through the ground. These currents will cause I^2R losses. Even for elevated radials, return currents can partially flow through the ground if a return path exists (can be by capacitive coupling if raised radials are close to ground). With a small elevated system, loss increases with any RF ground path at the antenna base, including the path back by the coax shield. This why the feed line should be decoupled for common modes at the antenna feed point with an elevated radial system.
- **Absorption losses**: The conductivity and the dielectric properties of the ground will play an important role in absorption losses, caused by an electromagnetic wave penetrating the ground. These losses are due to the interaction of the near-field energy-storage fields of the antenna (or radials) with nearby lossy media, such as ground. These types of losses are present whether elevated radials are used or not. The radials should shield the antenna from the lossy soil and distribute the field evenly around the antenna. Most often elevated radials don't help much here, since they normally aren't dense enough to make an effective screen. Four radials are far from a screen! The field is concentrated near the radials, and other areas are directly exposed to the antenna's induction fields.

In the far field (efficiency and effectiveness issues):

- Up to many wavelengths away, the waves from the antenna are reflected by the ground and will combine with the direct waves to form the radiation at low angles, the angles we are concerned with for DXing. The reflection mechanism, which is similar to the reflection of light in a mirror is described in Section 1.1.1. The real part of the reflection coefficient determines what part of the reflected wave is absorbed. The absorbed part is responsible for Fresnel-zone reflection losses (efficiency).
- The ground characteristics in the Fresnel zone will also determine the low-angle performance of the vertical, and this is an effectiveness issue.

The effect of ground in these two different zones has been well covered by P. H. Lee, N6PL (Silent Key), in his excellent book, *Vertical Antenna Handbook*, p 81 (Ref 701). The next section will cover these and various other aspects of the subject.

2. GROUND AND RADIAL SYSTEM FOR VERTICAL ANTENNAS: THE BASICS

2.0.1. Ground-Plane Antennas

We all know that a VHF vertical antenna usually employs four radials as a "ground-plane," hence its popular name. But in fact, two radials would do the same job. All you need with a $\lambda/4$ vertical radiator is a $\lambda/4$ wire connected to the feed-line outer conductor in order to have an RF ground at that point. The radial provides the other terminal for the feed line to "push" against. Unless the feed line is radiating, you will have exactly the same current into the radial (system) as you have in the form of common-mode current exciting the vertical. That is the "push against" effect of the radials. This is also how the antenna return currents are collected.

But if you have only one radial, this radial would radiate a horizontal wave component. Two λ/4 radials in a straight line have their current distributed in such a way that radiation from the radials is essentially canceled in the far field, at least in an ideal situation. This is similar to what happens with top-wire loading (T antennas). Using three wires (at 120° intervals) or four radials at right angles does the same also.

It was George Brown himself, Mr. 120-buried-radials, who invented elevated resonant radials. He invented the ground-plane antenna. The story goes that when Brown first tried to introduce his ground-plane antenna it had only two radials, but he had to add two extra radials because few of his customers believed that with only two radials the antenna would radiate equally well in all directions! In the case of a VHF ground plane mounted at any practical height above ground, there is no "poor ground" involved and all return currents are collected in the form of displacement currents going through the two, three or four radials.

The VHF case is where detrimental effects of real ground are eliminated by raising the antenna high above ground, electrically speaking. There are no I^2R losses, because the return currents are entirely routed through the low-loss radials. There also are no near-field absorption losses, since the real ground is several wavelengths away from the antenna. On VHF/UHF we are not counting on reflection from the real earth to form our vertical radiation pattern; we are not confronted by losses of Fresnel reflection in the far field either. In other words, we have totally eliminated poor earth.

2.0.2. Verticals with an On-Ground (or In-Ground) Radial System

The other approach in dealing with the poor earth is going to the other extreme — bring the antenna right down to ground level, and, by some witchcraft, turn the ground into a perfect conductor. This is what you try to do in the case of grounded verticals.

I^2R Losses

You can put down radials to improve the conductivity of the ground, and to reduce the I^2R losses (I is the *return current* of the antenna, R is the equivalent ground losses) as much as possible. (Also see Section 1.3.3.) This mechanism is well-known. You can also measure its effect: You know that as you gradually increase the number and the length of radials, the feed-point impedance is lowered, and with a fairly large number of long radials (for example, 120 radials, λ/2 long) you will reach the theoretical value of the radiation resistance of the vertical. In the worst case, when no measures are taken to improve ground conductivity, losses can be incurred that range from 5 to well over 10 dB with λ/4 long radiators, and much higher with shorter verticals.

Absorption Losses

Absorption losses (see Section 1.3.3.) by the lossy earth are less well-known in amateur circles. This is partly because you cannot directly measure the effects, as you can for I^2R losses. But the effect is nevertheless there and can result in 3 to 6 dB of signal loss if not properly handled. The RF energy radiation downward toward the ground hits the ground. This can be very close to the antenna (waves that will contribute to high angle radiation) or very far from the antenna (low angle waves). The ground as such is not a perfect reflector. There are different ways to tackle this problem:

- Move the base of the vertical well above ground (at least ⅜ λ) as is done on higher HF and VHF bands, so that the lossy ground is not in the antenna near field.
- Put a screen on the ground so the antenna does not see the lossy ground (hiding the lossy ground).

This means that in the case of buried or on-the-ground radials, their number and length must be such that the ground underneath is effectively made invisible to the antenna. It has been established experimentally that for a λ/4 vertical you must use at least λ/4-long radials, in sufficient number so that the tips of the radials are separated no more than 0.015 λ (1.2 meters on 80 meters and 2.4 meters on 160 meters). This means approximately 100 radials to achieve this goal. With half that number, you will lose approximately 0.5 dB due to near-field absorptive losses. This is RF "seeping" through an imperfect ground screen. In real life, taking good care of the I^2R losses with buried radials also means taking good care of the near-field absorption losses.

2.0.3. Verticals with a Close-to-Earth Elevated Radial System

In some cases it is difficult or impossible to build an on- or in-the-ground radial system that meets this requirement, in most cases because of local terrain constraints. In this case a vertical with a radial system *barely* above ground may be an alternative. The question is: how good is this alternative and how should we handle this alternative? With radials at low height (typically less than 0.1 λ above ground) you still must deal with effectively collecting return currents and with absorption losses in the real ground.

It is clear that if you raise an almost-perfect on-ground radial system higher above ground, it should yield an almost-perfect elevated-radial system. The *almost perfect* on-ground system would consist of 50 to 100 λ/4-long radials. In fact, the screening effect that is good for radials laying directly *on* the lossy ground will be even better if the system is raised somewhat above ground. That the screening of such a dense radial system is close to 100% effective was witnessed by Phil Clements, K5PC, who reported on the Internet that while walking below the elevated radial system (120 elevated radials) of a broadcast transmitter in Spokane, Washington, he could hardly hear the transmitted signal on a small portable receiver. The question, of course, is: Do we really need so many elevated radials, or can we live with many less? This question is one of the topics that I deal with in detail in Section 2.2 on elevated radial systems.

When dealing with the antenna return currents, it is clear that simple radial systems (in the most simple form a single radial) can be used. This has proven true for ages in VHF and UHF ground planes. The only issue here is the possible radiation of these radials in the far field, which could upset the effective radiation pattern of the antenna. This will also be dealt with in Section 2.2.

2.1. Buried Radials

Dr Brown's original work (Ref 801) on buried ground-radial systems dates from 1937. This classic work led to the still common requirement that broadcast antennas use at least 120 radials, each at least 0.5 λ long.

Table 9-1
Equivalent Resistances of Buried Radial Systems

Radial	*Number of Radials*				
Length (λ)	*2*	*15*	*30*	*60*	*120*
0.15	28.6	15.3	14.8	11.6	11.6
0.20	28.4	15.3	13.4	9.1	9.1
0.25	28.1	15.1	12.2	7.9	6.9
0.30	27.7	14.5	10.7	6.6	5.2
0.35	27.5	13.9	9.8	5.6	2.8
0.40	27.0	13.1	7.2	5.2	0.1

Table 9-2
Wave Angle and Pseudo-Brewster Angle for Ground-Mounted Vertical Antennas Over Different Grounds.

The Wave angle and the Pseudo Brewster angle are essentially independent of the radial system used, unless the radials are several wavelengths long.

Band/Ground Type	*Wave Angle*	*Pseudo-Brewster Angle*
80 meters		
Very Poor Ground	29°	15.5°
Average Ground	25°	12.5°
Very Good Ground	17°	7.0°
Sea Water	8.5°	1.8°
160 meters		
Very Poor Ground	28°	14.5°
Average Ground	23°	11°
Very Good Ground	19.5°	8.5°
Sea Water	8.5°	7.0°

2.1.1. Near-Field Radiation Efficiency

The effect of I^2R losses can be assessed by measuring the impedance of a λ/4 vertical, as a function of the number and length of the radials. Many have done this experiment. **Table 9-1** shows the equivalent loss resistance computed by deducting the radiation resistance from the measured impedance.

2.1.2. Modeling Buried Radials

Antenna modeling programs based on *NEC-4* (or later) can model buried radials. These programs address both the I^2R losses and the absorption losses in the near field, plus of course any far-field effects. These powerful new tools can be dangerous. They would make you believe you can now model everything, and that there is no need for validation. In the real world, mainly due to the non-homogeneous nature of the ground surrounding our antennas, the slight variations we sometimes see from modeling results are totally meaningless. (Some authors would rank modeled ground systems by quoting gains specified to a 1/100 of a dB!) At best modeling under such circumstances indicates a trend. Let's have a look at these trends.

R. Dean Straw, N6BV, ran a large number of models using *NEC-4* for me (at the time *NEC-4* was not available to non-US citizens). Separate computations were done for 80 and 160 meters. The radiators were λ/4 long and the radials were buried 5 cm in the ground. The variables used were:

- Ground: very poor, average, very good
- Radial length: 10, 20 and 40 meters (for 80 meters), and 10, 40 and 80 meters (for 160 meters)
- Number of radials: 4, 8, 16, 32, 64 and 120.

We computed the gain, the elevation angle and the pseudo-Brewster angle. Although we ordinarily talk about λ/4 buried radials, buried radials do not have to be resonant. A λ/4 wire that is resonant above ground is no longer resonant in the ground — not even on or near the ground. Typically for a wire on the ground, the physical length for λ/4 resonance will be approximately 0.14 λ and the exact length depending on ground quality and height over ground. Quarter-wave radials, in the context of buried radials, are wires measuring λ/4 over ground (typically 20 meters long on 80 meters and 40 meters long on 160 meters).

The gains of the modeling are shown in **Figs 9-14** through **9-19**. The wave angle as well as the Brewster angle are almost totally independent of the radial system in the near field. The values are listed in **Table 9-2**.

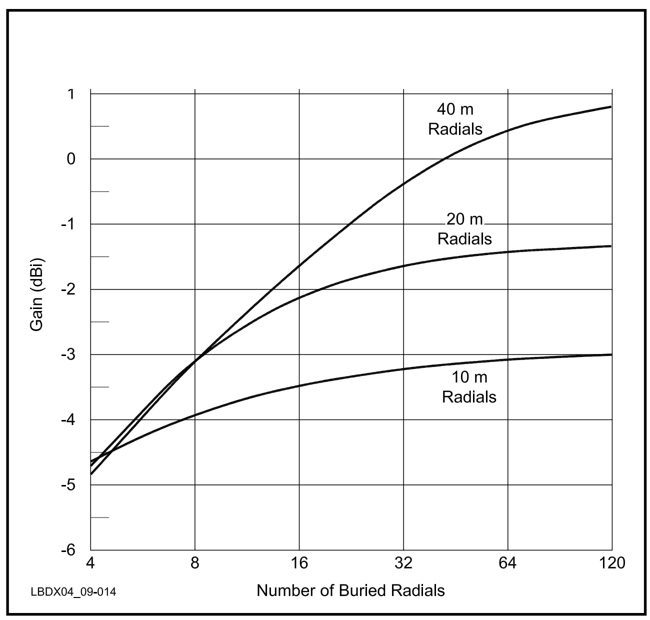


Fig 9-14 — Gain of 0.25-λ 80-meter vertical over very poor ground as a function of radial length and number of radials. For short (10-meter long) radials there is not much point in going above 16 radials. With 20-meter radials you are within 0.5 dB of maximum gain with 32 radials. If you want maximum benefit from 0.5-λ radials (40 meters), 120 radials are for you.

When modeling the antenna over poor ground using only four buried radials, it was apparent that the gain was slightly higher using 15-meter long radials rather than 20 meter or even 40-meter long radials (the gain difference being 0.7 dB, quite substantial). It happens that the resonant length of a λ/4 radial in such lossy ground is 10 to 15 meters (and not ≈ 20 meters as it would be in air). In case of a small number of radials, there is hardly any screening effect, and antenna return currents flow back through lossy, high-resistance earth to the antenna, as well as through the few radials. There are two parallel return circuits: a low-resistance one (the radials) and a high-resistance one (the lossy ground). If the radials are made resonant, their

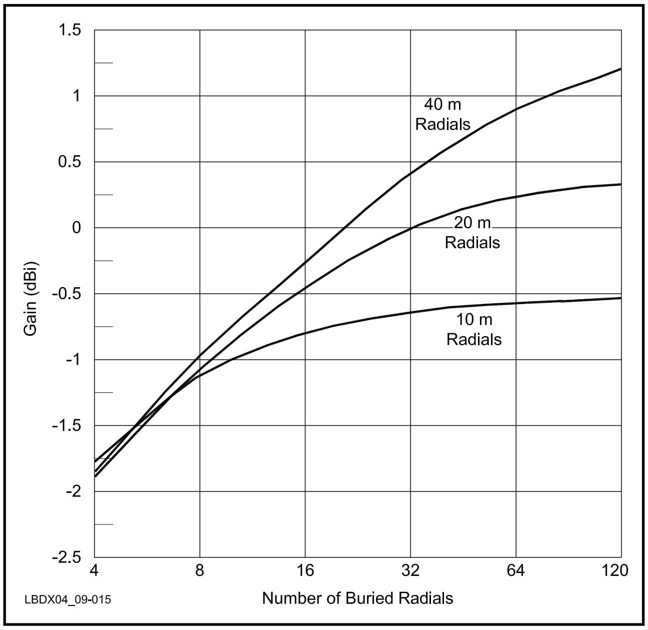


Fig 9-15 — Gain of a λ/4 80-meter vertical over average ground, as a function of radial length and number of radials. Note that for 10-meter long radials there is practically no gain beyond about 52 radials. For quarter wave radials there is little to be gained beyond 104 radials, and the difference between 26 λ/4 radials and 104 λ/4 radials is only 0.5 dB. These are exactly the same number N7CL came up with by experiment (see Section 2.1.3).

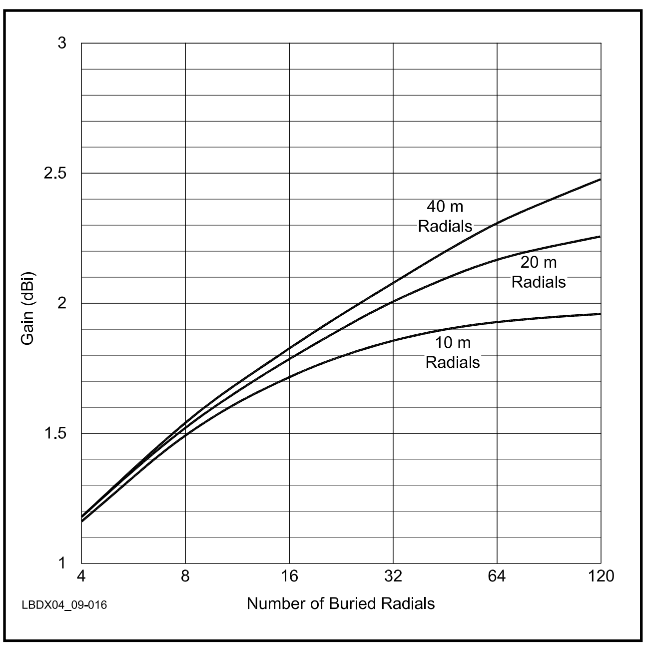


Fig 9-16 — Gain of λ/4 80-meter vertical over very good ground as a function of radial length and number of radials.

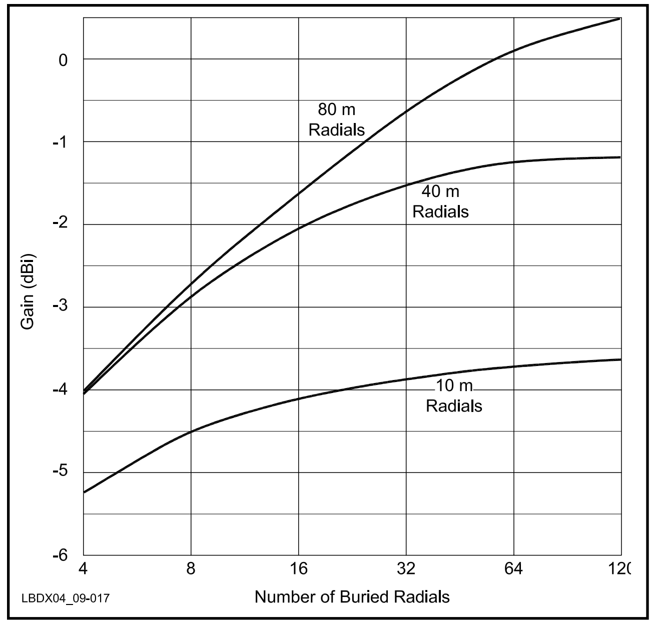


Fig 9-17 — Gain of λ/4 160-meter vertical over very poor ground as a function of radial length and number of radials. Note that 10-meter radials, no matter how many, are really too short for 160 meters.

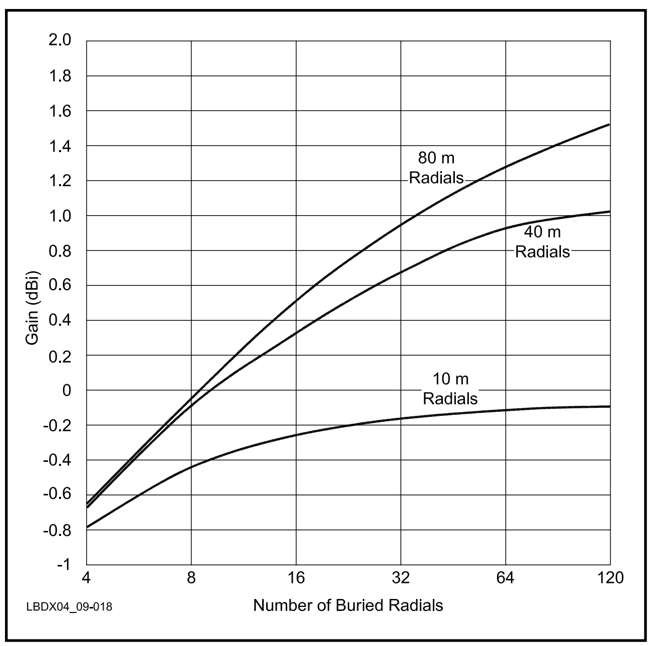


Fig 9-18 — Gain of λ/4 160-meter vertical over average ground as a function of radial length and number of radials.

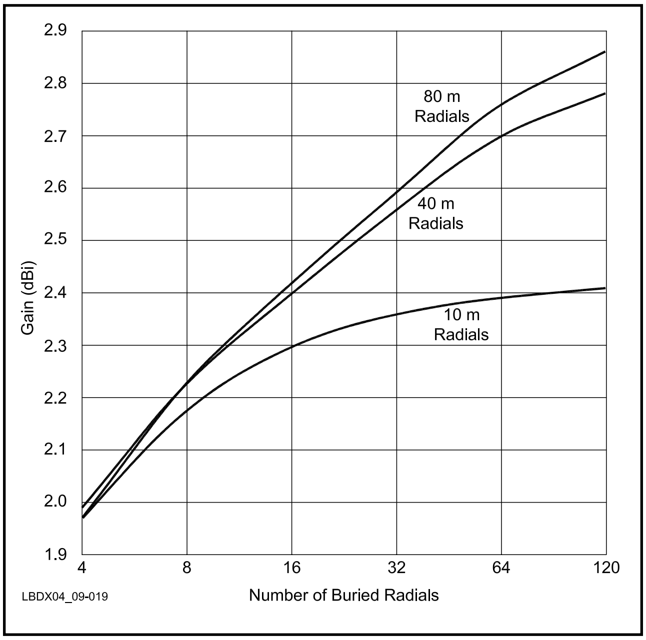


Fig 9-19 — Gain of λ/4 160-meter vertical over very good ground as a function of radial length and number of radials. The λ/2 radials are really a waste over very good ground.

impedance at the antenna feed point will be low, thereby forcing most of the current to return through the few radials. If the impedance is high (such as with 20 or 40-meter long radials), a substantial part of the return currents can flow back through the lossy earth. (See Section 2.1.3.)

The same phenomenon is marginally present with radials in average ground as well, but has disappeared completely in good ground. These observations tend to confirm the mechanism that originates this apparent anomaly. All of this is of no real practical consequence, since four radials are largely insufficient, in whatever type of ground (except saltwater).

We also modeled radials "in"/"on" seawater. With seawater you need no radials but you need to be able to connect the shield of the coaxial feed line to the saltwater, and that's not so easy. Carefully read Section 2.1.7.10. **Fig 9-20** shows

Fig 9-20 — XZØA had an ideal location for far-field reflection efficiency: Saltwater all around. Four Squares were used on 80 and 160 meters, resulting in signals up to S9+20 dB in Europe on Top Band, quite extraordinary from that part of the world.

an operation where saltwater was used to great advantage.

Years ago B. Edward, N2MF, modeled the influence of buried radials (Ref 816), and discovered that for a given number of radial wires, there is a corresponding length beyond which there is no appreciable efficiency improvement. This corresponds very well with what we find in Figs 9-14 through 9-19. Brian found that this length is (maybe surprisingly at first sight) nearly independent of earth conditions. This indicates that it is the screening effect that is more important than the return-current I^2R loss effect. Indeed, the effectiveness of a screen only depends on its geometry and not on the quality of the ground underneath. **Table 9-3** shows the optimum radial length as a function of the number of radials. This was also confirmed through the experimental work by N7CL (see Section 2.1.3).

Conclusion

To me, the results obtained when modeling verticals using buried radials with *NEC-4* seem to be rather optimistic, but the trends are clearly correct. Take the example of an 80-meter vertical over average ground: going from a lousy eight 20-meter long radials to 120 radials would only buy you 1.4 dB of gain, which is less than what I think it is in reality. In very good ground that difference would be only 0.7 dB!

There has been some documented proof that *NEC-4* does not handle very low antennas correctly, and that the problem is a problem associated with near-field losses (see Section 2.2.2.). Maybe this same limitation of *NEC-4* causes the gain figures calculated with buried radials to be optimistic as well. The future will tell. No doubt further enhancements will be added to future *NEC* releases, which may well give us gain (loss) figures that I would feel more comfortable with for verticals with buried radials.

2.1.3. How Many Buried Radials Now, How Long, What Shape?

When discussing radial lengths, I usually talk about λ/4 or λ/8 radials. Mention of a λ/4 radial leads most of us to think of a 20-meter long radial on 80 meters. A wire up in the air at heights where you normally have an antenna has a velocity factor (speed of travel vs speed of light) of about 98%. When you bring that same wire close to ground, the velocity factor starts dropping rapidly below a height of about 0.02 wavelength. On the ground, the velocity factor is on the order of 50-60%, which means that a radial that is physically 20 meters long is actually a half-wave long electrically!

If you use just a few on-the-ground radials over poor ground, the radials may act like they are somewhat resonant. The resonance vanishes if you have many radials or if the ground is good to excellent. For these cases it is best to use radials that are an *electrical* quarter-wave long. On 80 meters you should use 10-meter long radials, and on 160 meters you should use 20-meter long radials if you are only using a few (up to four). But that's bad practice anyhow: four is far too few radials.

As soon as you use a larger number of equally spread radials the resonance effect disappears, and the radials form a disk, which becomes a screen with no resonance characteristics. In this case we no longer talk about length of radials but about the diameter of a disk hiding the lossy ground from the antenna.

Assume we have 1 km of radial wire and unrestricted space. How should we use it? Make one radial that is 1000 meters long, or 1000 radials that are 1-meter long? It's quite obvious the answer is somewhere in the middle.

2.1.3.1. Early Work

Brown, Lewis, and Epstein in the June 1937 *Proceedings of the IRE* published measured field strength data at 1 mile (versus number and length of radials). Measurements were done at 3 MHz. The measured field strength was converted to dB vs the maximum measured field strength (for 113 radials of 0.411 λ). See **Table 9-4**.

2.1.3.2. Some Observations

- For short radials (0.137 λ), there is negligible benefit in having more than 15 radials.
- For radial lengths of 0.274 λ and greater, continuous improvement is seen up to 60 radials. Note that doubling the number and doubling the length of radials from the above case (15 short radials of 0.137 λ) only gains 1 dB greater field strength, with four times the total amount of wire.
- Lengthening radials 50% from 0.274 λ to 0.411 λ and keeping the same number hardly represents an improvement (0.24 dB). Raising the number to 113 radials represents a gain of 0.66 dB over the second case, but uses nearly three times as much wire.

From these almost 70-year-old studies, we can conclude that 60 quarter-wave long radials is a cost-effective optimal solution for amateur purposes.

N7CL's Work

The following rule was experimentally derived by N7CL and seems to be a very sound and easy one to follow. Put radi-

Table 9-3
Optimum Length Versus Number of Radials

Number of Radials	*Optimum Length (λ)*
4	0.10
12	0.15
24	0.25
48	0.35
96	0.45
120	0.50

This table considers only the effect of providing a low-loss return path for the antenna current (near field). It does not consider ground losses in the far field, which determine the very low-angle radiation properties of the antenna.

Table 9-4
From Brown, Lewis and Epstein
Signal Strength vs Length of Radials in Wavelengths

Number Radials	*Length 0.137 λ*	*Length 0.274λ*	*Length 0.411λ*
2	–4.36	–4.36	–4.05 dB
15	–2.4	–1.93	–1.65 dB
30	–2.4	–1.44	–0.97 dB
60	–2.0	–0.66	–0.42 dB
113	–2.0	–0.51	0 dB (Ref)

als down in such a way that the distance between their tips is not more than 0.015 λ. This is 1.3 meters for 80 meters and 2.5 meters on 160 meters.

The circumference of a circle with a radius of λ/4 is 2 × π × 0.25 = 1.57 λ. At a spacing of 0.015 λ at the tips, this circumference can accommodate 1.57/0.015 = 104 radials. With this configuration you are within 0.1 dB of maximum gain over average to good ground. If you space the tips 0.03 λ you will lose about 0.5 dB.

For radials that are only λ/8 long, a 0.03-λ tip spacing requires 52 radials. Here too, if you use only half that number, you will give up another 0.5 dB of gain. In general, the number that N7CL came by experimentally closely follow those from Brown, Lewis and Epstein. Let us apply this simple rule to some real-world cases:

Example 1

Assume your lot is 20 by 20 meters and that you want to install a radial system for 80 and 160 meters. Draw a circle that fits your lot. This circle has a radius of

$$\sqrt{20^2 / 2} = 14 \text{ meters}$$

On each 20-meter long side of your lot you would space the ends evenly by 1.3 meters. This means you can fit 16 radials on your property. The longest will be 14 meters; the shortest will be 10 meters long. The average radial length is 12 meters. You can install a total of 16 (radials) × 4 (sides) × 12 meters (average length) = 768 meters of radial wire, with a total of 64 radials. A radial system using 32 evenly spread radials, and using only 385 meters of wire, would compromise your system by about 0.5 dB.

In actual practice, when laying radials on an irregular lot where the limits are the boundaries of the lot, the most practical way to make best use of the wire you have is just walk the perimeter of the lot and start a radial from the perimeter (inward toward the base of the antenna) every 0.015 λ (1.3 meters for 80 meters or 2.5 meters for 160) as you walk along the perimeter.

Example 2

You have only 500 meters of wire and space is not a problem. How many radials and how long should they be to be used on both 80 and 160 meters?

The formula to be used is:

$$N = \frac{\sqrt{2 \times \pi \times L}}{A} \qquad \text{(Eq 9-6)}$$

where

N = number of radials
L = total wire length available
A = distance between wire tips (1.3 meters for 80, 2.5 meters for 160, or twice that if 0.5 dB loss is tolerated).

For this example use L = 500 meters, A = 1.3 meters, and you calculate:

$$N = \frac{\sqrt{2 \times \pi \times 500}}{1.3} = 43 \text{ radials.}$$

Each radial will have a length of 500/43 = 11.6 meters.

You could also use A = 2.6 meters, in which case you wind up with 22 radials, each 18 meters long. However, the first solution will give you slightly less loss.

For a given length of wire, it is better to use a larger number of short radials than a smaller number of long radials, the limit being that the tips should not be closer than 0.015 λ.

Example 3

How much radial wire (number and length) is required to build a radial system (for a λ/4 vertical) that will be within 0.1 dB of maximum gain. How much to be within 0.5 dB?

The answer to the first question is 104 radials, each λ/4 long. The total wire length for 80 meters is 2080 meters (4000 meters for 160). With 52 radials, each λ/4 long, you are within 0.5 dB of maximum gain. This translates to 1000 meters of radial wire required for 80 meters and 2000 meters for 160 meters.

Example 4

I can put down 15-meter long radials in all directions. How many should I put down, and how much radial wire is required?

The circumference of a circle with a radius of 15 meters is: 2 × π × 15 = 94.2 meters. With the tips of the radials separated by 1.3 meters, we have 94.2/1.3 = 72 radials. In total I would use 72 × 15 = 1080 meters of radial wire. There is no point in using more than 72 radials.

2.3.1.3. K3NA's Work

In private correspondence ("Effects of Ground Screen Geometry on Verticals"), Eric Scace, K3NA, explained a simple rule of thumb he derived from an extensive modeling study he conducted using *NEC-4*. His conclusions are applicable for radials up to 3λ/8 in length:

- Measure R, the real component of the feed-point impedance.
- Double the number of radials.
- Measure R again.
- Continue doubling the number of radials until R changes by less than 1 Ω.

K3NA's detailed modeling study to evaluate the effectiveness of various radial configurations was similar to what N6BV did years ago for the Third Edition of this book. The main difference between the two studies is that K3NA calculated the gain versus the total amount of radial wire used for different configurations. He calculated the *sky gain* (Gsky) to assess the quality of the radial system. Gsky is the total power radiated to the entire sky, covering all elevation angles, all azimuths.

K3NA was concerned with two aspects: the *efficiency* issue, which is related to the task of collecting return currents in the vicinity of a lossy ground and doing so with the smallest possible losses. (See definitions in Section 1.3.3.) The second issue is that of *effectiveness*, which means putting the radiated power where we want it. For a single vertical this means obtaining appropriate vertical angles of radiation, which is actually formed in the far field by the combination of the direct and the reflected waves.

Over Very Good Ground

K3NA used as a starting point in his studies the available quantity of radial wire. For up to 3 λ of available wire, the most efficient solution is to use λ/16 radials, even if there is space for longer ones. Beyond 48 radials, he found hardly any improvement. This confirms what we show in Fig 9-16. Not everything in his study is, however, a perfect match with the modeling results done several years ago by N6BV.

Figs 9-21 through **9-23** show the results for very good ground, good ground and very poor ground respectively. These confirm that any improvement in efficiency by improving the radial system improves radiation at all elevation angles equally. For regular-shaped radials laid out as the spokes of a wheel K3NA came to the conclusion that N7CL's rule of thumb, which says to *separate the tips of the radials by no more than 0.015* λ, is confirmed by modeling, at least for radials up to λ/4 in length.

In a later study (Ref 7801 and Ref 7802) K3LC confirmed these findings.

Other Configurations

K3NA also investigated the possibility of using radials that split out along their way: fork-shaped radials. He found out that for a given amount of available wire, these fork-type radials do not perform any better than regular straight radials.

A third alternative he examined was alternating long (λ/4) and short (λ/8) radials. Here too this radial geometry reduces Gsky compared to a system using the same total length of radial wire used as uniform-length straight radials.

Eric went on to assess the performance of ground screens in square and triangular grids. Here again, for a given amount of radial wire, the performance did not meet that of a classical radial configuration.

Looking at all these very detailed modeling results you must ask yourself: "Is it really like this in real life?" We are playing with very minute changes in inputs and obtaining even smaller changes in results. Can you really trust these models? Earth is a very difficult thing to model, and it is very non-homogeneous.

It's obvious that we should be conscious of trends, and the modeling results confirm the trends revealed by N7CL's experimental work. There's an even simpler rule: Put in as many radials as you can, until you feel satisfied. If you think you can do better, do better. If you think "this is as far as I can go," be happy with it!

Tom, W8JI, wrote this interesting observation for the Top Band reflector: "Even a very small limited space antenna like an inverted L will do very well if some effort is put into the ground system. My friend K8GIJ was always within a few dB of my signal (I used a ¼ λ vertical tower with 100 radials), and all he had was a 15 by 100 foot back yard! But then Harold filled his small yard with radials, and even tied the fences and everything else in to his ground system." So, you guys on a city size lot, there is no reason not to be loud on 160 meters.

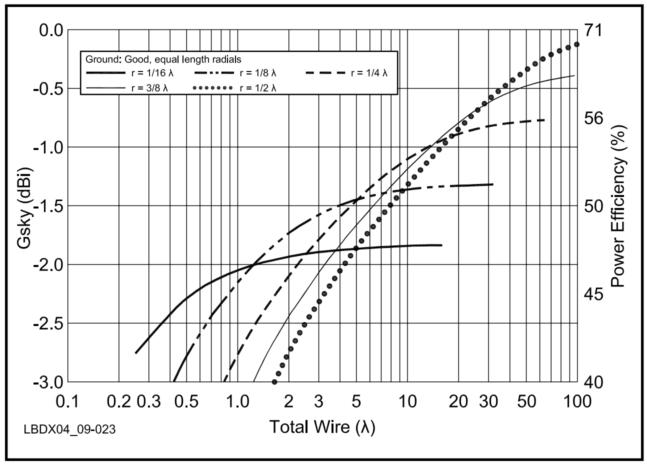


Fig 9-22 — The same graph as in Fig 9-21 but for good ground. Unless you only have 4 λ of wire, λ/8 radials are really too short; λ/4 radials are just fine for up to about 20 λ of wire (this is about 3.3 km or 10,000 ft of radials on 160 meters). Notice that this study also shows that there is little to be gained beyond approx 100 λ/4 radials. 300 λ/2 radials only gain about 0.7 dB (a power increase of only 20%) over 100 λ/4 radials — not really a whole lot!

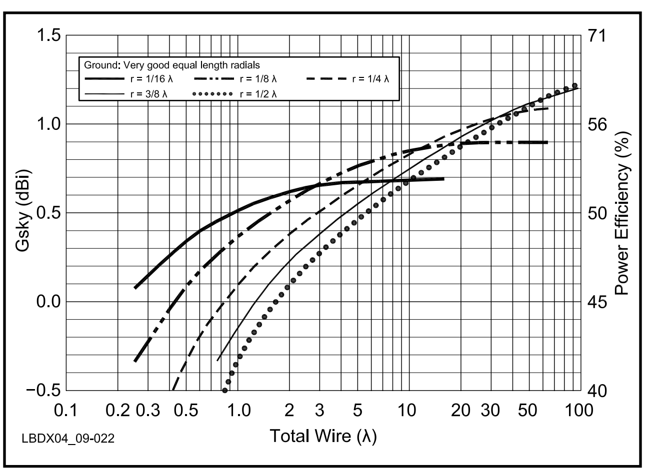


Fig 9-21 — Total sky-gain results over very good ground for various radials systems using standard radials, shaped like the spokes of a wheel. The graph shows clearly that with small amounts of wire, many short radials are the answer. It also tells us that 10 λ of radial wire used to make 80 λ/8 radials is only 0.2 dB down from 30 λ of radial wire used to make 120 λ/4 radials.

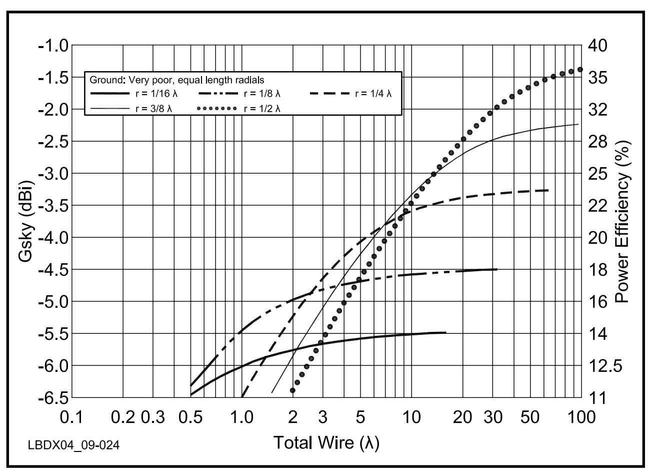


Fig 9-23 — Have a look at the gain axis: No matter what you do (lots of λ/2 long radials), –1.3 dBi is the limit for very poor ground (as compared to 0 dBi for good ground and approximately +1.5 dB over very good ground). The nearly 3 dB difference is due to the Fresnel-zone reflection efficiency.

Of course, to be able to *hear* as well as Tom, W8JI, is another challenge.

2.1.4. Two-Wavelength-Long Radials and the Far Field

Everything that happens in the near field determines the radiated field strength in the far field. Radials, screens, and I^2R losses have very little influence on the radiation pattern of the vertical, except maybe at very high angles, which don't interest us anyhow. Any method of improving ground conductivity in the near field (up to $\lambda/4$ from the base of a $\lambda/4$ vertical) improves the entire radiation pattern, not just favoring certain radiation angles more than others.

In the far field, however, ground characteristics greatly influence the low-angle characteristics of a vertical antenna. For $\lambda/4$ verticals the area where Fresnel reflection occurs starts about 1 λ from the antenna and extends to a number of wavelengths.

For current collecting and near-field screening there is really no point in installing radials longer than $\lambda/4$. With 104 such radials you are within 0.1 dB of what is theoretically possible. The Brown rule (120 radials, 0.5-λ long) shoots for less than 0.1 dB and has some extra reserve built in. Watch out: All of this is mathematics, nobody will be able to tell the difference, and who can measure the difference? Calculating is easy; measuring is a different game.

If you want to influence the far field and pull down the radiation angle somewhat, or reduce the reflection loss, then we are talking about radials that are at least 2 λ long (and much longer if you consider really low angle). For this you would need a terrain measuring 660 × 660 meters (43 hectares or 100 acres) for Top Band, which is hardly practical, of course.

The only practical way to influence the far-field reflection efficiency and effectiveness is to install your vertical in the middle of or along saltwater. In that case you will have a peak radiation angle of between 5 and 10° and a pseudo-Brewster angle of less than 1°! The elevation pattern becomes very flat, showing a –3 dB beamwidth ranging from 1 to 40°. All this is due to the wonderful conductivity properties of saltwater. If the antenna is erected on a saltwater shore, the benefit will evidently only apply in the direction shooting across the sea.

Tom Bevenham, DU7CC (also SM6CNS), wrote: "At my beach QTH on Cebu Island, I use all vertical antennas standing out in saltwater. Also, at high tide, water comes all the way underneath the shack. On Top Band, I use a folded monopole attached alongside a 105-foot bamboo pole. This antenna is a real winner. I use not much of a ground system, only a few hundred feet of junk wire at sea bottom. At the other QTH, less than half a mile from the beach, the same antennas with ground radials don't work at all."

Of course, we have all heard how well the over-saltwater vertical antennas perform. I remember the operation from Heard Island (VKØIR) for one. The Battle Creek Special (see Section 6.7) was standing with its base right in the saltwater (**Fig 9-24**). In recent years, DXpeditions have discovered that even on the higher bands, verticals near saltwater are better alternatives to Yagis at very low heights that radiate at very high angles.

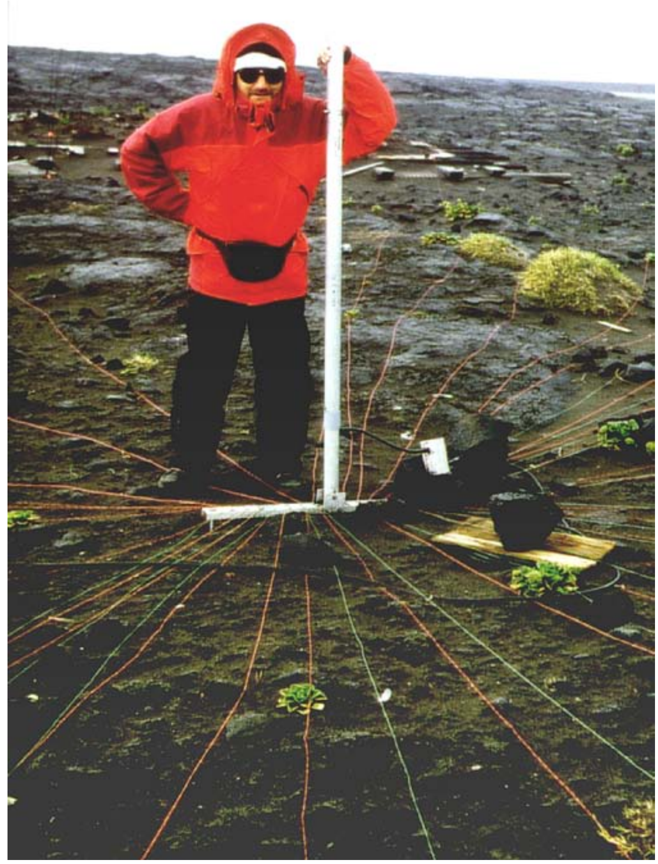


Fig 9-24 — The Battle Creek Special that made Heard Island available on 160 for over 1000 different stations. Ghis, ON5NT, is not holding up the antenna; it is very well capable of standing up by itself (how about the OM?...). The antenna was located near the ocean's edge, on saltwater-soaked lava ash.

2.1.5. Ground Rods and Static Discharge

Ground rods are important for a good dc ground, which is necessary for adequate lightning protection, even if ground rods contribute very little to the RF ground system. If you use a series-fed (insulated-base) vertical, a lightning arrestor spark gap with a good dc ground is a good idea. In addition, you can install a 10 to 100-kΩ resistor or an RF choke between the base of the antenna and the dc ground to drain static charges.

2.1.6. Depth of Buried Radials

C. J. Michaels, W7XC (Silent Key), calculated the depth of penetration of RF current in different types of ground. He defined the depth of penetration as the depth at which the current density is 37% of what it is at the surface. On 80 meters he calculated a depth of penetration of 1.5 meters for very good ground. For very poor ground the depth reaches 12 meters!

From the point of view of I^2R loss, you can bury the radials "deep" without any ill effects. However, from near-field screening effect point of view, we need to have the radial system above the lossy material.

Bob Leo, W7LR, in Ref 808 reports that burying the radials a few inches below the surface does not detract from their performance. Al Christman, K3LC (ex KB8I), confirmed this when modeling his elevated radial systems using *NEC-4*. He found a difference of only hundredths of a dB between burying radials at 5 cm or 15 cm (ever hear the difference in signal strength of 0.01 dB…?). I would not bury them much deeper though. The sound rule here is "the closer to the surface, the better."

2.1.7. Some Practical Hints

2.1.7.1. Local Ground Characteristics

It is impossible to make a direct measurement of ground characteristics. The most reliable source of information about local ground characteristics may be the engineer of your local AM broadcast station. The so-called "full proof-of-performance" record will document the average soil conductivity for each azimuth out to about 30 km (20 miles). But unfortunately this is hardly what you need to know. What you need is the ground characteristics in a circle with a $\lambda/4$ radius around the base of your vertical! In your modeling program you plug in a single set of values that supposedly characterize your ground. In the real world, the soil around an antenna is virtually never homogeneous — and almost always not even remotely close to homogeneous. Real-world earth is a widely varying mix of moisture, as well as different types of soil. Because of this, any model that treats the earth as a uniform medium will not be accurate. Verification by field-strength measurement is the only way to know for sure what's going on!

2.1.7.2. Radial Bus-Bar/Low-Loss Connections

There are two good ways to collect the currents in the many radials at the base of the vertical. You could use a radial plate (see **Fig 9-25**) and use stainless-steel hardware to connect the radials. Using solder lugs and stainless steel hardware makes it possible to disconnect the radials so that individual radial-current measurements can be made.

Another method is to make a heavy gauge bus-bar made of a large diameter copper ring, and solder all (copper) radials to the bus (see **Fig 9-26**).

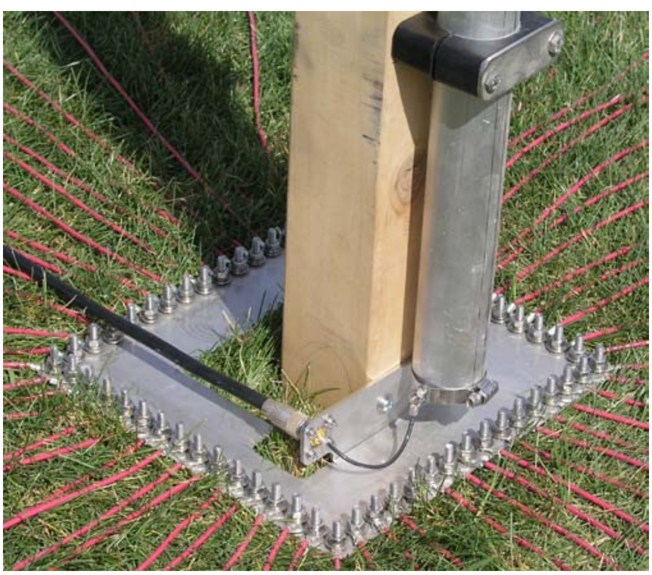


Fig 9-25 — The stainless steal radial plate made by DX Engineering has 64 holes drilled around its perimeter. All the stainless steel hardware is provided to make a quality radial connecting system using crimped lugs at the ends of the radials.

Fig 9-26 — W8LRL uses a copper tube (about 10 mm in diameter) bent in a circle, to which he solders all his radials. For a permanent installation this is probably the best way to go, provided you use silver solder or protect your connections made with regular electronics solder with liquid rubber.

2.1.7.3. Soldering/Welding Radial Wires

Tin-lead (Sn-Pb) or lead-free solder normally used in electronics construction is often used to join copper radial wires. Regular solder will deteriorate in the ground and may be the source of bad contacts. Therefore you should use silver-solder on all copper radials, or even better yet, weld the radials. Information about CADWELD welding products from The RF Connection is available on their Web page: **www.therfc.com**.

If you decide to use regular electronics solder, cover all soldered joints with several layers of liquid rubber, so that the acidity of the ground cannot reach the solder joint.

2.1.7.4. Sectorized Radial Systems

Very long radials (several wavelengths long) in a given direction have been evaluated and found to be effective for lowering the wave angle in that direction, but seem to be rather impractical for just about all amateur installations. A similar effect occurs when verticals are mounted right at the saltwater line. Similar in result to a sectorized radial system is the situation where an elevated radial system is used with only one radial (see Section 2.2.6).

2.1.7.5. Radial Wire Material

Use copper wire if at all possible. Galvanized-steel wire is not good, as it has poor conductivity and will rust away in just a few years in wet acidic ground. Aluminum is okay as far as conductivity is concerned, but aluminum gradually turns to a white powder as it reacts with the soil. Soldering aluminum wire is not easy, and crimp-on lugs are the only way to go if you decide to use aluminum.

2.1.7.6. Radial Wire Gauge

When less than six radials are used, the gauge of the wires is important for maximum efficiency. The heavier the better — #16 AWG wires are certainly no luxury when only a few buried radials are used. With many radials, wire size becomes unimportant since the return current is divided over a large number of conductors. DXpeditions using temporary antennas often take a small spool of #24 or #26 AWG (0.5 or 0.4-mm diameter) enameled magnet wire. This is inexpensive and can be used to establish a very efficient RF ground system.

2.1.7.7. Bare or Insulated Wire?

Experience has shown that you can use insulated as well as bare copper wire for buried radials. The *NEC-4* modeling program finds no noticeable difference between insulated and bare buried radials. This relates to the capacitive coupling between the radial wire and the earth around it. Experience is what counts, and the modeling program gives the correct answer on this issue!

2.1.7.8. A Radial Plow

Installing radials can be quite a chore. Hyder, W7IV, (Ref 815) and Mosser, K3ZAP, (Ref 812) have described systems and tools for easy installation of radials.

2.1.7.9. Radials on the Ground

Radials can also be laid on the ground (instead of being buried in the ground) in areas that are suitable. A neat way of installing radials in a lawn-covered area is to cut the grass really short at the end of the season (October), and lay the radials flat on the ground, anchored here and there with metal hooks (clothespins, doll pins, gutter nails or fencing staples). By the next spring, the grass will have covered up most of the wires, and by the end of the following year the wires will be completely covered by the grass. This will also guarantee that your radials are "as close as possible" to the surface of the ground, which is ideal from a near-field screening point of view!

2.1.7.10. Radials and Saltwater

The conductivity of saltwater is excellent. But you should also remember that the skin depth of saltwater is very limited, and you better keep that in mind when you install radials "in" saltwater. Throwing radials in saltwater and letting them sink to the bottom is like installing radials "under" a copper plate: not much use! It seems best to have many short radials dangling from the base of the antenna into the saltwater or better yet, have a few copper plates extending into the saltwater to ensure a large contact surface with the saltwater.

If your antenna is exposed to the tide it seems like a good idea to have a floating device with large copper fins extending under the device in the saltwater. If the area gets dry at low tide, you should also have regular radials lying on the ground. Over saltwater, two in-line elevated radials make a very valid alternative, that's much easier to install! As a matter of fact you can have the vertical and the two in-line elevated radials as much as ¼ λ from the saltwater and still get all the extra "saltwater gain."

2.2. Elevated Radials

With elevated VHF or UHF ground-plane antennas the three or four radials are more an electrical *counterpoise* (a zero ohm connection point high above ground, to "push against"), than a ground plane. The ground is so far away that any term including the word "ground" is really not applicable. The radials of such antennas radiate in the near field (radiation from the radials only cancels in the far field), but they do not suffer near-field absorption losses in the ground, because of their relative height above ground.

Using a small number of elevated radials does not prevent the antenna and its radials from coupling heavily to the feed line and from inducing common-mode currents onto the feed line (see also Chapter 8, Section 1.5). There will be substantial feed-line radiation unless you isolate the feed line from such common-mode currents.

Such HF and VHF/UHF ground planes have been in use for many years. We are concerned however with vertical antennas using radials at much lower heights, typically 0.01 to 0.04 λ above ground. It is no secret to insiders that there is still quite a bit of controversy on this subject. It appears that a number of real-life results do confirm the current modeling results, while others do not. The jury is still out. I will try to represent both views in this book.

A. Doty, K8CFU, concluded from his experimental work (Ref 807 and 820) that a λ/4 vertical using an elevated counterpoise system can produce the same field strength as a λ/4 vertical using buried bare radials. The reasoning is that in the case of an elevated radial or counterpoise system, the return currents do not have to travel for a considerable distance through high-resistance earth, as is the case when buried radials are used. His article in April 1984 *CQ* also contains a very complete reference list of just about every publication on the subject of radials (72 references!).

Frey, W3ESU, used the same counterpoise system with his Minipoise short low-band vertical (Ref 824). He reported that connecting the elevated and insulated radial wires together at the periphery definitely yields improved performance. If a counterpoise system cannot be used, Doty recommends using *insulated* radials lying right on the ground, or buried as close as possible to the surface.

Quite a few years after these publications, A. Christman, R. Redcliff, D. Adler, J. Breakall and A. Resnick used computer modeling to come to conclusions which are very similar to the findings brought forward after extensive field work by A. Doty. The publication in 1988 by A. Christman, K3LC (ex KB8I), has since become the standard reference work on elevated radial systems (Ref 825), work that has stirred up quite a bit of interest and further investigation.

The results from Christman's study were obtained by computer modeling using *NEC-GSD*. It is interesting to understand the different steps he followed in his analysis (all modeling was done using average ground):

1) Modeling of the λ/4 vertical with 120 buried radials (5-cm deep). This is the 1937 Brown reference. (See Section 2.1.3.1.)
2) The λ/4 vertical was modeled using only four radials at different radial elevations. For a modeling frequency of 3.8 MHz, Christman found that 4.5 meters was the height at which the four-radial systems equaled the 120-buried-radial systems so far as low-angle radiation performance is concerned.
3) Christman's studies also revealed that as the quality of the soil becomes worse, the elevated radial system must be raised progressively higher above the earth to reach performance on par with that of

the reference 120-buried-radial vertical monopole. If the soil is highly conductive, the reverse is true.

The elevated-radial approach has become increasingly popular with low-band DXers since the publication of the above work, and it appears that elevated radials represent a viable alternative to digging and plowing, especially where the ground is unfriendly for such activities.

It is important to critically analyze the elevated-radial concept and therefore to understand the mechanism that governs the near-field absorptive losses (see Section 1.3.3) connected with elevated radials. In the case of an elevated-radial system these near-field losses can be minimized in only three ways:

1) By raising the elevated radials as high as possible (move the near field of the antenna away from the real lossy ground).
2) By installing many radials, so that these radials screen the near fields from "seeing" the underlying lossy earth.
3) By improving ground conductivity of the real ground below the raised radials.

Although the experts all agree on the mechanisms, there appears to be a good deal of controversy about the exact quantification of the losses involved (see Section 2.2.1).

Gull Wing Radials

Incidentally, an elevated radial system does not imply that the base of the vertical must be elevated from the ground. The radials can, from ground level, slope up at a 45º angle to a support a few meters away, and from there run horizontally all the way to the end. This is what we commonly call "gull wing fashion radials." It is a good idea to keep the radials high enough so no passersby can touch them. This is also true when radials are quite high. In an IEEE publication (Ref 7834) it was reported that significantly better field strengths were obtained with elevated radials 10 meters high than with radials 5 meters high. In both cases the radials were sloping upward at a 45° angle from the insulated base of the vertical at ground level.

2.2.1. Modeling vs Measuring? Elevated vs Ground Radials

The *comparative* performance of an elevated radial system or a buried radial system can be assessed by computer modeling. I am convinced that the *operational* performance of radial systems can only be assessed by real-life testing and field-strength measuring. In the ideal case the results should match. This is not the case, as has been reported over and over.

Ideally each study based on computer modeling should be followed by real-life field testing. Being in a critical mood, if the modeling study has not been confirmed by measurements, one could wonder if we are studying the performance of elevated radials or studying the performance of the modeling software and a given modeling approach.

In the pre-*NEC-4* days when we could not yet model radials buried in the ground, it was common to calculate the gain of the antenna by using tables providing data on how the equivalent ground resistance evolved as a function of number and length of radials over different sorts of ground (see Table 9-1). These data were not obtained through modeling but through real life field strength measurements. I still use this method extensively in this book, mainly as it is based on measured data.

With the advent of *NEC-4*, modelers can now include the buried radials in their antenna model. I have seen a lot of publications based on *NEC-4* modeling where I have serious doubts when I see the results. Was the modeling not done correctly, or does *NEC-4* still need some refinement? Only a lot of real life verification over time will tell us.

Al Christman, K3LC, used *NEC-4* to study the influence of the number of elevated radials and their height on antenna gain and antenna wave angle (Ref 7825) and came to the conclusion that if the height of the radials is at least 0.0375 λ (3 meters on 80, 6 meters on 160) there is very little gain difference between using four or up to 36 radials. He also concluded that the gain of antennas with an elevated radial system compared in gain to the same antenna with about 16 buried radials. Incidentally, the modeling also showed that for buried λ/4 radials the difference in gain between 16 radials and 120 radials is *only* about 0.74 dB (although almost 1 dB on 160 when signals are riding in, on or under the noise can be a lot). When raising the elevated radials to a height of 0.125 λ (20 meters on 160), the gain actually approached the gain of a vertical with 120 buried radials. The publication of these results (1988) gave a tremendous impetus in the use of elevated-radial systems.

In another study, Jack Belrose, VE2CV (Ref 7821 and 7824) also concluded that there was a good correlation between measured and computed results. In this study Belrose used a λ/4 vertical, as well as λ/4 (resonant) radials.

A good correlation between the modeled results and field strength measurements was established in several study cases. One of them was an extremely well-documented case, with thousands of field strength measurements, which matched very well the figures obtained with modeling (*NEC-4*). Belrose's studies revealed that radials should be at least 0.03-λ high (2.5 meters on 80 meters, 5 meters on 160 meters) to avoid excessive near-field absorption ground losses, especially so if fewer than eight radials are used. With a large number of radials (>16) the radials can be much lower.

Another well-documented case was reported in a technical paper delivered by Clarence Beverage (nephew of Harold Beverage) at the 49th NAB Broadcast Engineering Conference entitled: "New AM Broadcast Antenna Designs Having Field Validated Performance." The paper covered antenna tests done in Newburgh, NY, under special FCC authority. The antenna system consisted of a tower 120 feet in height with an insulator at the 15-foot level and six elevated radials a quarter wavelength in length spaced evenly around the tower and elevated 15 feet above the ground. The system operated on 1580 kHz at a power of 750 W. The efficiency of the antenna was determined by radial field-intensity measurements (in 12 directions) extending out to distances up to 85 km. The measured RMS efficiency was 287 mV/m (normalized) to 1 kW at 1 km, which is the same measured value as would be expected for the tower above with 120 buried radials.

In a number of other cases however, it was reported that field-strength measurements indicated a discrepancy of 3 to 6 dB with the *NEC-4* computed results. Tom Rauch, W8JI, published the following measured results:

Number of Radials	*On the Ground*	*Elevated 0.03 λ*
4	–5.5 dB	–4.3 dB
8	–2.7 dB	–2.4 dB
6	–1.3 dB	–0.8 dB
32	–0.8 dB	–0.7 dB
60	0 dB (Reference)	–0.2 dB

In other words, Tom's measurements indicated that four elevated radials are down 4.3 dB vs 60 radials on the ground.

Calculations with *NEC-4* show a difference of only about 2 dB going from 4 to 60 buried radials, which is 3.5 dB less that Rauch's experiment showed. The 5 dB he found inspired the following comment: "Consider that going from a single vertical to a four square only gained me 5 dB! I got almost that just by going from four radials to 60 radials."

Eric Gustafson, N7CL, reported (on the Top Band reflector) that several experiments comparing signal levels of a ground mounted λ/4 vertical with 120 radials with those produced by the same radiator with an elevated radial system (using a few radials) have been done a number of times by various researchers for various organizations ranging from the broadcast industry and universities to the military. He reported that the results of these studies always have returned the same results: "the correctly sized, sufficiently dense screen is superior to four resonant radials in close proximity to earth". The quantification of the difference has varied. The largest difference Eric personally measured during research for the military was 5.8 dB, the smallest difference 3 dB. The latter one was measured over really good ground, being a dry salt-lake bed (measured conductivity approximately 20 mS during the test). It is clear that the quality of the ground plays a very important role in the exact amount of loss. These figures seem to confirm the values W8JI reported.

For those who would like to duplicate these tests, understand that you cannot do these tests on one and the same vertical, switching from elevated radials to ground-mounted radials, *unless* you remove (physically) the ground-mounted radials when you use the elevated ones. If not, you have an elevated radial system *plus* a screen, effectively screening the near fields from the underlying real ground.

It seems to me that elevated-radial systems are indeed a valid alternative for buried ones, especially if buried ones are not possible or very difficult to install for whatever reason. Even the broadcast industry now uses elevated-radial systems quite extensively and successfully where local soil conditions make it impossible to use the classic 120 buried λ/2 radials. It must be said though that most of these systems use more than just a few radials. I also know of many amateur antenna systems successfully using elevated radial systems. Whether they get optimum performance or lose maybe 2 to 5 dB because of near-field absorption losses, is hard to tell. As a matter of fact, there is still the possibility of improving the ground conductivity *under* the elevated radial system. More on that in Section 2.2.13.

The discrepancy between measured and modeled gain figures has been recognized by a number of expert *NEC* users. All of the current modeling programs have flaws, but most are known and can be compensated for by experienced users. It seems to me that modeling of very low wires even with current versions of *NEC-4* may be affected by such a flaw.

We should also recognize that the total losses due to mechanisms in the near field can amount to much more than 5 dB. Antenna return-current losses (sometimes also called "connection" losses) can range easily from 10 to even 40 dB over poor ground. These losses can, however, easily be mastered with elevated radials and reduced to zero. The remaining 4 or 5 dB, accountable to near-field absorption losses, are indeed somewhat more difficult to deal with using elevated radials.

2.2.2. Modeling Vertical Antennas with Elevated Radials

As mentioned before only *NEC*-based programs can model antennas with elevated radials close to ground. Roy Lewallen's *EZNEC* program (using the *NEC-2* engine) incorporates the "high-accuracy" (*NEC* Sommerfeld) ground model, which should be accurate for low horizontal wires down to 0.005-λ high (about 0.8 meters on 160 meters). Using the *NEC-4* engine, one can now also model antennas with buried radials (see Chapter 4, Section 1.4).

Still, many cases have been reported indicating a difference of up to 6 dB in gain for antennas very close to ground. A similar flaw was already present in *NEC-2* and has been documented by Jack Belrose, VE2CV, who compared the experimentally obtained results, published by Hagn and Barker in 1970 ("Gain Measurements of a Low Dipole Antenna Over Known Soil"), with the *NEC-2* predictions. At 0.01 λ above ground, *NEC-2* showed 5 dB more gain than the actual measured values.

All of this goes to say that modeling software is a *mathematical tool.* Most modeling programs have limitations. Some are well-known, but sometimes they have little-known or barely documented limitations.

Field-strength measurements are the real thing (eating is the proof of the pudding). But we should be thankful for having access to antenna-modeling programs. They have undoubtedly helped the non-professionals to gain an enormous amount of insight they would miss without these tools. It is the role of the professionals and the experts to show non-expert users how to use them correctly, and make corrections if necessary.

2.2.3. How Many Elevated Radials?

Through antenna modeling, Christman, K3LC, calculated (for 80 meters), the λ/4 antenna gains for elevated radial heights of 5, 10, 15, 20, 25 and 30 meters, while varying the number of λ/4 radials between 4 and 36 (Ref 7825). According to these calculations, at a height of 4.5 meters (which is roughly what I have) it made less than 0.1 dB of difference between 4 and 32 radials, and this was within 0.3 dB of a buried radial system using 120 quarter-wave radials. These results were confirmed by Belrose, VE2CV (Ref 7821) *also through antenna modeling*.

Eric Gustafson, N7CL, in a well documented e-mail addressed to the Top Band reflector, explained that for a λ/4 vertical radiator, a radial system with 104 λ/4-long radials (resulting in wire ends separated not more than 0.015 λ at their tips) achieves 100% shielding effectiveness. His experimental work (radials about 5 meters high) further indicates that the screening effectiveness of a λ/8-long radial system does not improve above 52 radials. See Fig 9-14, where we note that the experimental work by N7CL confirms the modeling results. Beyond 104 λ/4 radials there hardly is any increase in gain, and the same is true beyond 52 radials that are λ/8 long. This means that the shielding effectiveness of the λ/8 radial system with 52 radials by itself is 100%, but that some loss will be caused by near fields "spilling over" the screen at its perimeter. (In other words, the screen is dense enough, but not large enough.) Using just 26 λ/4-long radials, you will typically lose about 0.5 dB due to near-field absorption losses in the ground.

N7CL goes on to say that a λ/4 vertical with only four elevated radials can indeed produce the same signal as a ground-mounted vertical with 120 radials λ/4 long, provided that:

- The base of the vertical is at least 3λ/8 high.
- Or that the quality of the ground under the elevated radials has been improved so that it acts as an efficient screen, preventing the nearby field from interacting with the underlying lossy ground.

Unless such measures are effectively taken, N7CL calculated that the extra ground absorption losses can be as high as 5 or 6 dB. Loss figures of this order have been measured in a number of cases (eg, by Tom Rauch, W8JI) reported on the Top Band Reflector (see Section 2.2.1).

Conclusion

According to the *NEC*-based modeling results, there should be no point in using more than four elevated radials. With four radials over good ground the gain of a λ/4 monopole is –0.1 dBi. Two such radials gives an average of –0.15 dBi (+0.14 and –0.47 dBi due to slight pattern squeezing). One elevated radial gives a gain of +1.04 dBi in the direction of the radial, and –2.3 dBi off its back, resulting in an integrated gain of 0.65 dBi. These optimistic figures drove many people to use four elevated radials on their verticals, convinced that they would be as loud as their neighbors using 120 buried radials. Over the years, though, the enthusiasm for elevated radials seems to have somewhat settled down, and many have returned to the old-fashioned large numbers of radials on the ground, at least where feasible.

The *NEC*-based modeling programs are overly optimistic when it comes to dealing with near-field absorption losses. Three or four elevated radials over a poor ground, in my humble opinion, can never be as good as 120 ground-mounted (or elevated for that matter) radials. There is simply no free lunch! If you need to use an elevated radial system, maybe it's not a bad idea after all to use 26 radials, which according to N7CL would put you within 0.5 dB of the Brown standard.

2.2.4. Radial Layout

If you use a limited number of elevated radials (two, three or four), a symmetrical layout is necessary for the radiation from radials to cancel "as much as possible" in the far field. One radial is not symmetrical, but two and more are symmetrical, provided the radials are spread out evenly over 360°. When using more than four radials the exact layout as well as the exact radial length is not important when considering high-angle radiation.

2.2.5. Elevated Radials by the Sea

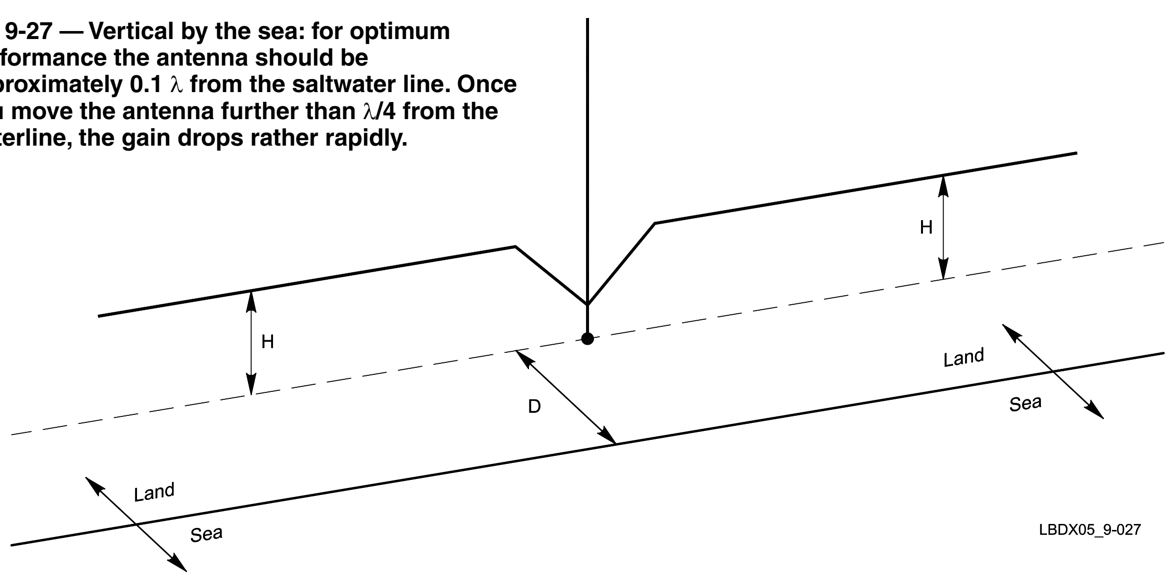
In order to get the maximum benefit from lossless far field reflection on saltwater, the vertical should be not further than λ/4 from the saltwater line. In a typical setup, two inline gull wing elevated radials run parallel to the saltwater line at a height of approximately 2 meters (**Fig 9-27**). The base insulator of the antenna should be high enough to keep it above water at all times.

An advantage of using just two elevated radials is that you can "tune" the vertical (adjust its resonant frequency) by merely changing the length of the radials. To change from the phone to the CW portion of the band, use a short piece of wire with an alligator clip to extend the radials. Make sure you have two radials of the same length to avoid low angle radiation from the radials.

Do not forget to install a common mode choke on the feed line. On 160 meters you can use an inverted-L vertical with the top loading section sloping away from the water.

The extra gain in the direction of the water is 5 to 6 dB, as compared to the opposite direction. The peak radiation angle in the direction of the sea is typically between 2° and 5° (vs 26° in the opposite direction).

Fig 9-27 — Vertical by the sea: for optimum performance the antenna should be approximately 0.1 λ from the saltwater line. Once you move the antenna further than λ/4 from the waterline, the gain drops rather rapidly.



Al Christman, K3LC modeled this setup in detail (Ref 7806, 7807 and 7808). Real life experiences, mainly during DXpeditions, confirm the modeling conclusions (remember VP6DX?).

2.2.6. Only One Radial

In his original article on elevated radials (Ref 825) Christman showed the model of a λ/4 vertical using a single elevated radial. This pattern shown in **Fig 9-28** is for a radial height of 0.05 λ over average ground. He showed this vertical, with a single elevated radial, as having (within a minor fraction of a dB) the same gain in its favored direction as a ground-mounted vertical with 120 buried radials.

Note however that the pattern is non-symmetrical. The radiation favors the direction of the radial, resulting in a 3 to 4 dB F/B over average ground. Modeling the same vertical over very good ground results in much less directivity, and over saltwater the antenna becomes perfectly omnidirectional.

I expect that it is sufficient to install radials on the ground under the antenna to improve the properties of the ground in the near field of the antenna to a point where the directivity, due to the single radial, is reduced to less than 1 dB. The slight directivity can be used to advantage in a setup (Ref 7824) where the vertical is part of a fixed array, and where you make use of the initial directivity of each element to provide some added directivity (see **Fig 9-29**).

The single radial not only creates some horizontal directivity, but also introduces some high-angle radiation, caused by the radiation from the single radial. If two or more radials are used, they can be set up in such a way that the horizontal radiation of these radials is effectively canceled. Notice from Fig 9-27 that most of the high-angle pattern energy is at or near 90°.

If you are looking for maximum low-angle radiation (which is normally the case for DXing), using only one radial is not the best choice, especially if the antenna is going to be used for reception as well. In a contest-station environment, however, creating some high-angle radiation, to give some "presence" on the band with locals can be desirable. If separate directive low-angle receiving antennas (eg, Beverages) are used, using a single radial on a vertical may well be a logical choice. I am using a single 5-meter high elevated radial on my 80-meter Four Square (radials pointing out of the square). At the same time I have a decent shielding effect on the real ground because of the more than 200 radials for the 160-meter vertical, which supports the 80-meter wire Four Square (see Chapter 11).

A vertical with a single radial can also be a logical choice for a DXpedition antenna (near/over saltwater or over a good ground screen) for two reasons:

1) Ease of adjusting resonance from the CW to the phone end of the band, by just lengthening the radial.
2) Extra gain by putting the radial in the wanted direction (toward areas of the world with high amateur population density).

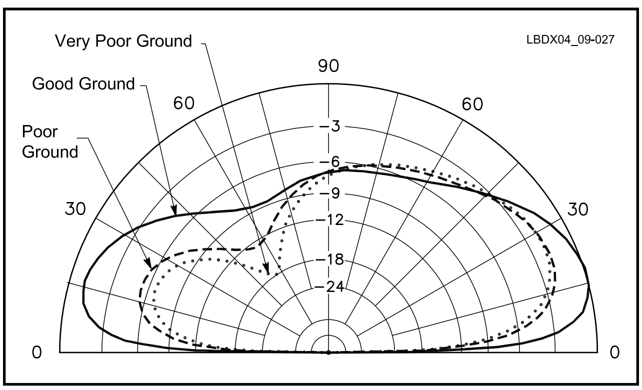


Fig 9-28 — Vertical radiation pattern of a quarter-wave vertical with one horizontal λ/4 radial at a height of 0.05 λ over different types of ground.

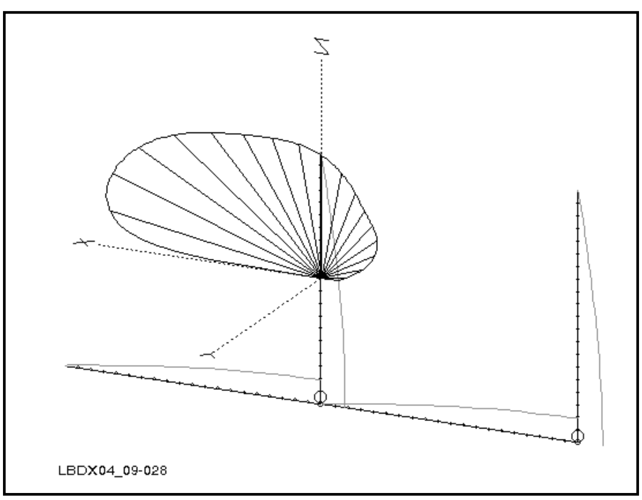


Fig 9-29 — Two λ/4 verticals are used in an end-fire configuration (see Chapter 10), producing a cardioid pattern. By placing the single radial in the forward direction of the array, some additional gain can be achieved. This technique makes it impossible to switch directions.

2.2.7. How High Should the Radials Be?

The *NEC*-modeling results, published by Christman, K3LC, indicate that radials above a height of approximately 0.03 λ achieve gains within typically 0.2 dB of what can be achieved with 64 buried radials. In other words, there is no point in raising the radials any higher than 6 meters on 160 or 3 meters on 80 meters.

Measurements done by Eric Gustafson, N7CL, however, tell us a totally different and very logical story. To prevent the near fields created by the radial currents from causing absorption losses in the underlying ground, the radials must be high enough so that the near fields do not touch ground. With up to six radials, this is between λ/8 and λ/4. Below λ/8 the losses are very considerable (if no other screen is available). For amateur purposes with four radials, a minimum height of λ/4 would be a reasonable limit to use. The minimum height decreases as the density of the radial screen is increased. With a density of about 100 quarter-wave long radials (in which case the distance between the tips of the radials is 0.015 λ) the radial plane can be lowered all the way *onto* the ground without incurring significant near-field absorption loss. This is shown in Fig 9-15, where beyond 100 radials there is little to gain. At a height of about 0.03 λ, 26 radials will result in an absorption loss of not more than 0.5 dB, according to N7CL.

Conclusion

If you want to play it extra safe, and if you have the tower height, get the radials up as high as possible and add a few more. Having more radials will make their exact length much less critical as well.

Equally effective is to use a ground screen in addition to a few elevated radials. However, in actual practice we often see elevated radials used because the ground conditions do not allow installing wires, which means that installing a screen may be equally as difficult.

It all is very logical. Get away from the lossy ground by raising the radials higher above the ground or hide the lossy ground with a dense screen using many radials.

Again, there is no free lunch!

2.2.8. Why Quarter-Wave Radials in an Elevated Radial System?

In modeling it is quite easy to create perfectly resonant quarter-wave radials. Why do we want them to be exactly λ/4 long? Let's examine this issue. What we really want is the vertical plus the radials to be resonant, not because this would make the antenna radiate better, but only because that makes it easier to feed the antenna.

Dick Weber, K5IU, found through a lot of measuring and testing of real-life verticals with elevated radials that using λ/4-long elevated radials has a certain disadvantage. In his models he used four radials (one per 90° of azimuth) because he wanted the radiation from these radials to be completely canceled: no pattern distortion and no high-angle horizontally polarized radiation. He found out though that this is very difficult, if not impossible, to achieve in the real world. Of course, λ/4 radials works fine on a computer model, since you can define four radials that have exactly the same electrical length. But this is not always the case in the real world. One radial will always be, perhaps by only a minute amount, electrically longer or shorter than another one. And therein lies the problem. We want these four radials all to carry exactly the same current, in order for the radiation to balance out.

The real question is how important are equal currents in the radials? I modeled several cases of intentional radial current imbalance. **Fig 9-30** shows the vertical radiation pattern of a λ/4-vertical (F = 3.65 MHz), with two elevated radials, 4 meters high. Pattern A is for two radials showing no reactance (both perfectly 90°, which can never be achieved in real life). For pattern B, I have intentionally shortened one radial about 20 cm (approximately 1% of the radial length). This introduced a reactance of $-j\,8\,\Omega$ for this radial. One radial now carried 62% of the antenna current, the other the remaining 38%. Over good ground this imbalance causes the horizontal pattern to be skewed about 0.6 dB (an inconsequential amount), but we see a fill-in of the high-angle rejection (around 90° elevation) that we would expect to have when the currents are really equal. Pattern C is for a case where one radial is 20 cm too short, and the other one 20 cm too long (reactance $-j\,8\,\Omega$ and $+j\,8\,\Omega$). In this case the relative current distribution was very similar as in the first case (63% and 36%). The horizontal pattern skewing was the same as well. Pattern D is for a rather extreme case

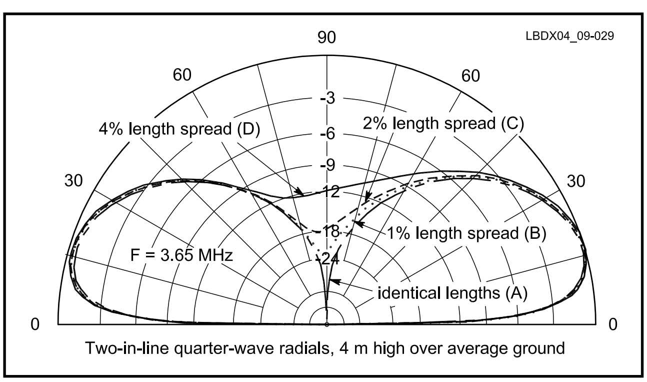


Fig 9-30 — Vertical radiation patterns (over good ground) for a λ/4 long 80-meter vertical, with two in-line radials 4 meters high, for various radial lengths around λ/4. See text for details.

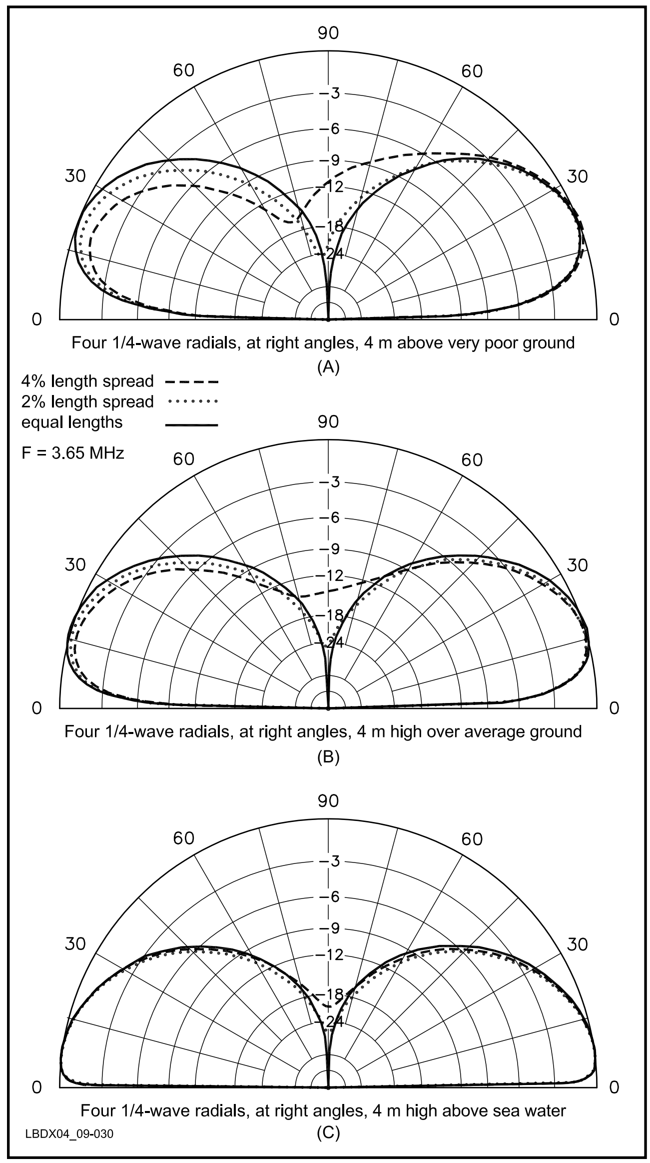


Fig 9-31 — Vertical radiation patterns of an 80-meter λ/4 vertical with four elevated radials (4 meters high) over various types of ground. Patterns are for: (A) average ground, (B) very good ground and (C) saltwater. See text for details.

where radials differ 80 cm in length ($+j$ 16 Ω and $-j$ 16 Ω). Current imbalance has now increased to 76% versus 24%.

I did a similar computer analysis for a vertical using four elevated radials. In this case, I did the analysis over three different types of ground: good ground, very good ground and seawater (ideal case).

Fig 9-31 shows the results of these models. Case A is for equal currents in the four radials (theoretical case); case B is for radials showing reactances of $+j$ 8 Ω, 0 Ω, $-j$ 8 Ω, and $+j$ 10 Ω. The relative current distribution in the four radials was 51%, 39%, 5% and 5%, which are values very similar to what has been measured experimentally by K5IU. Pattern C shows a rather extreme imbalance with radial reactances of $-j$ 16 Ω, 0 Ω, $+j$ 16 Ω and $+j$ 8 Ω (a total length spread of 4% of the nominal radial length). In this case the relative currents in the radials are 54%, 28%, 8% and 10%. Plot 1 is for the antenna over good ground, Plot 2 over very good ground, and Plot 3 over seawater.

Note that the pattern deformation depends to a very high degree on the quality of the ground under the antenna! Over seawater the current imbalances practically cause no pattern deformation at all. The horizontal pattern squeeze is at maximum 1.6 dB over good ground, and 0.6 dB over very good ground, computed at the main elevation angle.

From this it appears that in addition to using a few (typically less than 10) elevated radials, it is a good idea to improve the ground conductivity right under the radials by installing a ground screen using radials there as well. This is for two different reasons: To form a screen hiding the lossy ground from the antenna, and to reduce the effect of high-angle radiation from the radials.

You should understand that if you have enough elevated radials any variation in the exact electrical length will *not* result in high-angle radiation or pattern squeezing. With 16 radials, length variations of ±1.5%, and angular variations of ±5° (not evenly spaced in azimuth), the effect is of no consequence, resulting in horizontally polarized radiation components down >40 dB). The radials now form a screen that no longer shows resonance, just like the case with radials on the ground.

You also need a large number of elevated radials to avoid excessive near-field losses. You can kill two birds with one stone with a raised radial system using at least 16 radials.

Dick Weber, K5IU, measured many real-life installations with either two, three or four elevated radials, and it was not uncommon to find one radial taking 80% of the antenna current, one radial 20% and the other two almost zero! The recorded variations in radial currents were used to calculate the patterns shown in Fig 9-30.

The question now is whether or not you can live with the high-angle fill in (mostly around the 90° elevation angle) and slight pattern-squeeze (typically not more than 1 dB). If you want maximum low-angle radiation, and if you don't want to lose a fraction of a dB, and if you don't want to put up a few more radials, then equal-radial currents may be for you. Or maybe you would like some high-angle radiation? Maybe you are not using your vertical or vertical array for reception, and you want some high-angle radiation? If you are a contest operator, this is a good idea (you want some local presence as well). In that case, don't bother with equal radial currents, maybe just one radial is the answer for you, as I did.

However, even a small number of radials that are laid out perfectly symmetrically and that carry identical currents are no guarantee of 100% cancellation of the horizontal high-angle radiation in the far field. Slight differences in ground quality under the radial wires (or environment, trees, bushes, buildings) can result in different near-field absorption losses under radials that would otherwise carry identical RF currents. The result will be incomplete cancellation of their radiated fields in the far field. Measuring radial currents does not, indeed, tell you the full story!

It is interesting though to understand why slight differences in radial lengths can cause such large differences in radial current. A λ/4 radial is equivalent to an open-circuited λ/4 transmission line that uses the ground as the second conductor. This acts like a dead short at its resonant frequency. When this short is connected in parallel with another λ/4 radial, it's like connecting a short circuit across another short circuit, and then expecting that both shorts will take exactly the same current.

We have similar situations in electronics when we parallel devices such as power transistors in power supplies, or when we parallel stubs to reject harmonics on the output of a transmitter. If one stub gives us 30 dB of attenuation, connecting a second one right across the first one will increase the attenuation by 3 dB at the most. If we take special measures (λ/4 lines at the harmonic frequency) between the two stubs, then we get greater attenuation (almost double that of the single stub, an additional 6 dB). **Fig 9-32** shows the equivalent schematic of the situation using λ/4 radials.

Conclusions

1) For elevated radial systems using two, three or four (resonant) λ/4 radials, slight differences in electrical length cause radial current imbalances, resulting in some high-angle radiation as well as some pattern squeezing, especially over less than very good ground. However, even perfectly balanced currents are not a 100% guarantee for

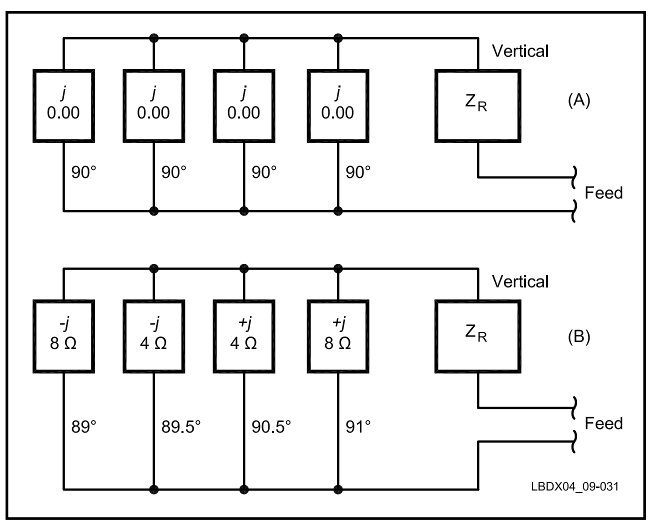


Fig 9-32 — At A the ideal (not of this world) case where all four radials are exactly 90° long. They all are a perfect short and exhibit zero reactance. At B the real life situation, where it now is clear that in this circuit, where the current divides into four branches, these currents are now very unequal.

zero high-angle radiation (due to unequal near-field ground losses under different radials).

2) Starting with eight radials (or more) the influence of unequal radial current on the generation of high-angle radiation is almost nonexistent. If you are greatly concerned about a little high-angle radiation, you should simply increase the number of elevated radials to eight.

3) Adding a good ground screen under the antenna totally annihilates the effects of unequal radial currents, and in addition it will raise the gain of the antenna by up to 5 dB!

By the way, you need not be concerned about any of these issues with a classic in-ground (or on-the-ground) radial system using 60 radials.

2.2.9. Quarter-Wave Radials of Equal Length

Despite all of that, it's nice to know how you can make λ/4 radials of identical electrical length! In the past, one of the standard methods of making resonant radials was to connect them as a (low) dipole and prune them to resonance. It is evident that resonance does not mean that both halves of the dipole have the same electrical length, even if both halves are the same physical length. One half could exhibit + *j* 20 Ω reactance, while the other half could exhibit a so-called conjugate reactance, – *j* 20 Ω. At the same time the dipole would be perfectly resonant.

Nevertheless, there is a more valid method of constructing radials that have the same electrical length. Whether these are perfect λ/4 radials is not so important, we can always tune out any remaining reactance with a small series coil or a capacitor (if too long). This method is as follows:

- Model the length of the vertical to be λ/4 at the design frequency.
- Put up an elevated vertical of the computed length.
- Use one of the charts in **Fig 9-33** to determine the theoretical radial length. Note that the length is very dependent on radial height.
- Connect one radial.
- Trim the radial to bring the vertical to resonance.
- Disconnect the radial.
- Put up the second radial in line with number one.
- Trim this second radial for resonance.
- If you use four radials, do the same with the remaining two radials.

Then connect all radials to the vertical and check its resonant frequency. It is likely that the vertical will no longer be resonant at the design frequency. Is it necessary to have the vertical at exactly λ/4? No, but if you want, here are two procedures to make the antenna plus radials perfectly resonant on your design frequency.

2.2.9.1. First Method

This requires changing the length of the vertical to bring the system to resonance. Do not change any radial length, but change the length of the vertical to achieve resonance at the desired frequency.

2.2.9.2. Second Method

Change all radials in length by exactly the same amount (all together, not one at a time) until you establish resonance. Neither of these two methods guarantees that both the radial system and the vertical are exactly a quarter wavelength; they only guarantee that both connected together are resonant. Again, it is totally irrelevant whether both are 90° long or not. It is not unusual that radials of different physical length result in identical electrical lengths. This is mainly due to the variation of ground conductivity, which can vary to a wide degree over small distances. Other causes are coupling to nearby conductors.

On the other hand, radials of exactly the same electrical length are still no guarantee of identical radial current because of near-field losses being different under different radials.

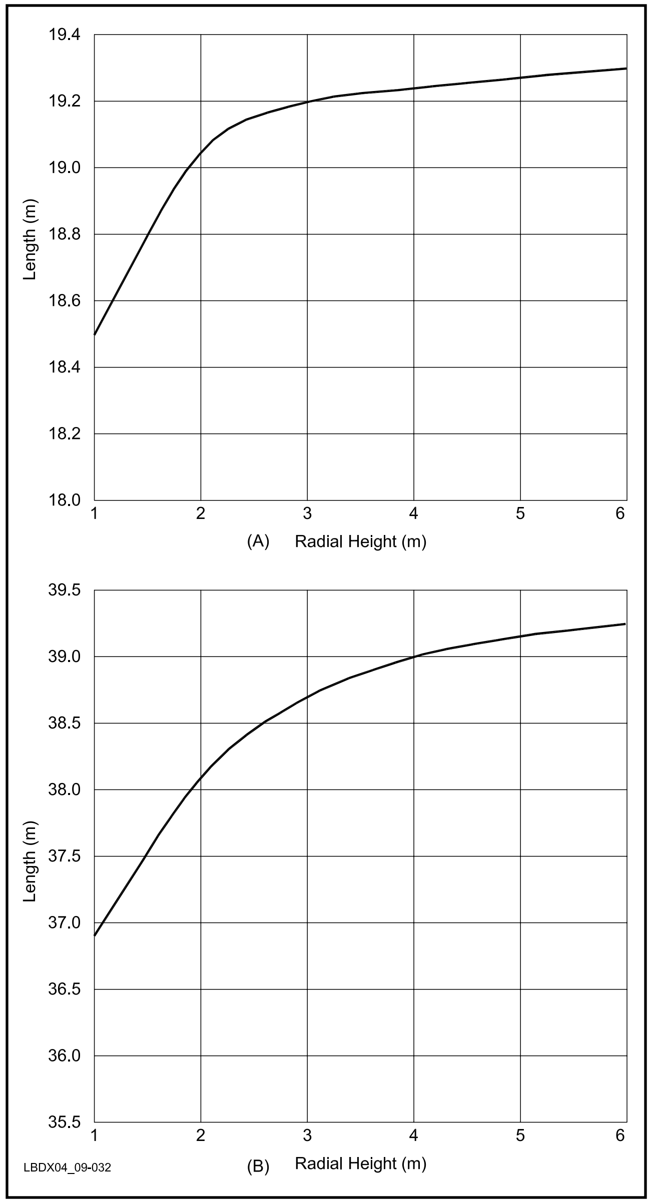


Fig 9-33 — Length of a λ/4 radial as a function of the height above ground. For 80 meters at A; for 160 meters at B.

2.2.10. The K5IU Solution to Unequal Radial Currents

D. Weber, K5IU, inspired by Moxon (Ref 693, pages 154-157 in the First Edition, pages 182-185 in the Second

Edition, and Ref 7833) installed radials shorter than λ/4 and tuned the radial assembly to resonance with a coil. It appears that slight changes in electrical length of these "short" radials have little influence on the current in the various radials (Ref 7822 and 7823).

Weber's modeling studies showed that radial lengths between 45° and 60° and between 115° and 135° resulted in minimum creation of high-angle radiation from unequal electrical radial lengths. When using radials longer than 90° the system can be tuned to resonance using a series capacitor, which is easier to adjust than a coil and which also has intrinsically less losses (see **Fig 9-34**). The purist may even use a motor-driven (vacuum) capacitor, which could be used to obtain an almost perfect SWR anywhere in the band.

I would suggest, however, not shortening the radials to less than approximately 60° to 70° if not really necessary. It is clear that we cannot indefinitely shorten radials, and expect to get the same results. If that were true we should all use two in-line loaded mobile whips on our 160-meter tower as a radial (current collecting) system. T. Rauch, W8JI, put it very clearly on the Top Band reflector: "The last thing in the world I'd want to do is concentrate the current and voltage in smaller areas. Resonant radials, or especially shortened resonant radials, concentrate the electric and magnetic fields in a small area. This increases loss greatly. The ideal case is where the ground system carries current that evenly, and slowly, disperses over a large physical area, and has no large concentrated electric fields from high voltage."

This is clearly another plea for the classic, multi-radial ground system. I did some modeling myself using *EZNEC* and found that:

- The fewer the radials, the greater the current imbalance due to length variations.
- The worse the ground quality the greater the impact of current imbalance on the radiation pattern.
- Starting with 16 radials, the effect of current imbalance is totally gone, even with 90° radials.

Conclusion

You can solve the problem of high-angle radiation by using a larger number of radials (for example, 16) or by improving the ground quality under the radials by installing a ground screen, at the same time yielding less near-field ground-absorption losses!

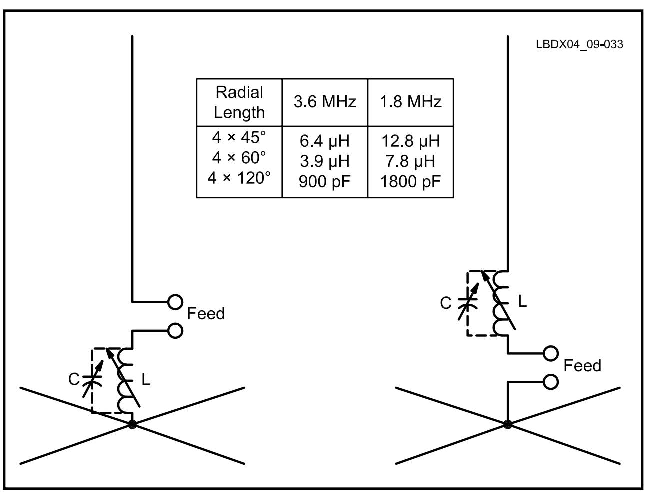
2.2.11. Should the Vertical be a Quarter-Wave?

From a radiation point of view, neither a vertical with a buried-radial ground system nor one with an elevated-radial system necessarily must be resonant. We usually make these resonant because it makes feeding the antenna easier.

A buried ground-radial system is a non-resonant, low-impedance system. Over such a ground system the vertical is usually made resonant (90° long electrically), to have a non-reactive feed-point resistance. Verticals somewhat longer than λ/4 (usually about 3λ/8) can be tuned to resonance using a series capacitor. Although most 3λ/8 verticals use ground-mounted radials, the same can be done with a 3λ/8 vertical using elevated radials.

Remember that with a small number of radials (up to about 10), the length of each of these radials is critical and the radial system has a resonant character that is more pronounced as the number of radials is reduced. This means that if you use only a few radials, you can adjust their length to change the resonant frequency of the vertical. With a large enough number of radials the system becomes non-resonant (like a ground screen) and changing radial lengths has no influence on the resonant frequency of the antenna system. See **Fig 9-35**.

Using this concept we can envision a 3λ/8 vertical to be used in conjunction with, say, λ/8 long radials. A



Radial Length	3.6 MHz	1.8 MHz
4 × 45°	6.4 μH	12.8 μH
4 × 60°	3.9 μH	7.8 μH
4 × 120°	900 pF	1800 pF

Fig 9-34 — When radials shorter than 90° are used, the system must be tuned to resonance using a coil. With radials longer than 90° the tuning element is a capacitor. Typical values for the tuning elements are also shown. The feed line can be connected in two different ways: Between the tuning element and the radiator or between the tuning element and the radials. The result is exactly the same. In both cases, a coaxial feed line connected to the feed point must be equipped with a current balun.

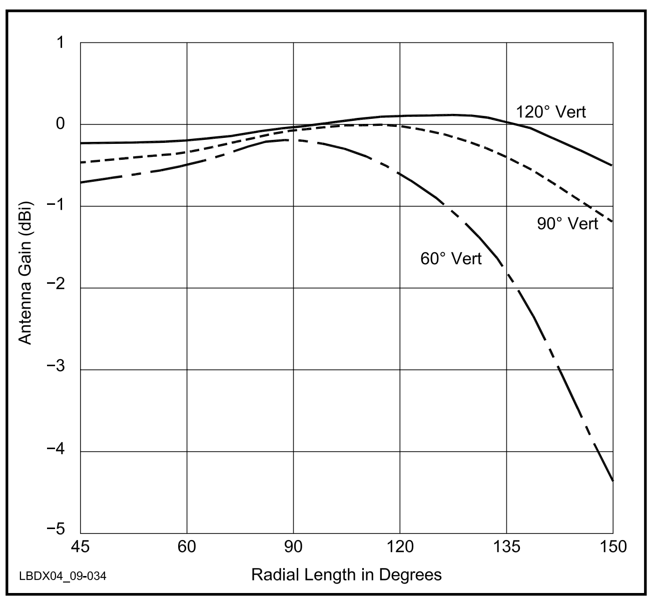


Fig 9-35 — Gain as a function of radial length for verticals measuring 60°, 90° and 120° over average ground (all using four elevated radials about 0.012-λ high) as calculated by K5IU.

3.75-MHz vertical designed according to these principles is shown in **Fig 9-36**. The combination of a 3λ/8-long radiator and λ/8-long radials does not require a coil to tune the antenna. The radiator length shown for a wire element whose diameter is 2 mm is 26.9 meters long. With four 10-meter long radials, the feed impedance is exactly 52 Ω, a excellent match for 50-Ω feed line.

The same vertical can be turned into an 80/160-meter vertical using 27-meter long radials (60° on 160 meters and 120° on 80 meters) as shown in **Fig 9-37**. The total system length on 160 meters is 60° + 60° = 120°, which is less than 180° (λ/2); hence a coil is required to resonate the antenna. On 80 meters, the total length is 120° + 120° = 240°, which is longer than λ/2; hence, a capacitor is required.

Here too, make sure you install common mode RF choke on the feed line!

2.2.12. Elevated Radials on Grounded Towers

2.2.12.1. The N4KG Antenna

T. Russell, N4KG, an eminent low-band DXer, described a method of shunt feeding grounded towers in conjunction with elevated radials (Ref 7813 and 7832). His tower uses a TH7DX triband Yagi as top loading to make it about 90° long with respect to the feed point (see **Fig 9-38**). It is important to find the attachment point of the radials on the tower whereby the part of the tower above the feed point becomes resonant in conjunction with the radials. Russell installed 10 λ/4 radials and moved the ring to which these radials were attached up and down the tower until he found the system in resonance. This point was 4.5 meters above ground.

Belrose, VE2CV, analyzed N4KG's setup using *NEC-4* (Ref 7821). He simulated the connection to earth of the tower (at the base) by using a 5-meter long ground rod (a decent dc ground). It is obvious that RF current is flowing through the tower section below the feed point. This current causes the gain of the antenna to be somewhat lower than that of a λ/4 base-fed tower. Belrose calculated the difference as 0.8 dB.

A typical configuration like the one described by N4KG will yield a 2:1 SWR bandwidth of 100 to 150 kHz. There are several approaches to broadband the design. Sam Leslie, W4PK, designed a system where he uses two sets of two radials, installed at right angles. One set is cut to resonate the system at the low end of 80 meters (CW band) and the other at the phone end. The SWR curve has two dips now, one on 3.5 and the other on 3.8 MHz.

Another approach is to design the antenna for resonance on 80-meter CW, and tune it to resonance in the SSB portion by inserting a capacitor between the feed line and the radials or the vertical conductor (tuning out the inductive reactance on 3.8 MHz).

Once more, make sure you have common-mode chokes on the feed line near to the feed point. In case of the two parallel 75-Ω feed lines, you need those on both!

2.2.12.2. Decoupling the Tower Base from the Real Ground

It is possible to minimize the loss by decoupling the base of the vertical from ground. Methods of doing so were described by Moxon (Ref 693 and 7833). **Fig 9-39** shows the layout of a so-called *linear trap* that turns the tower section between

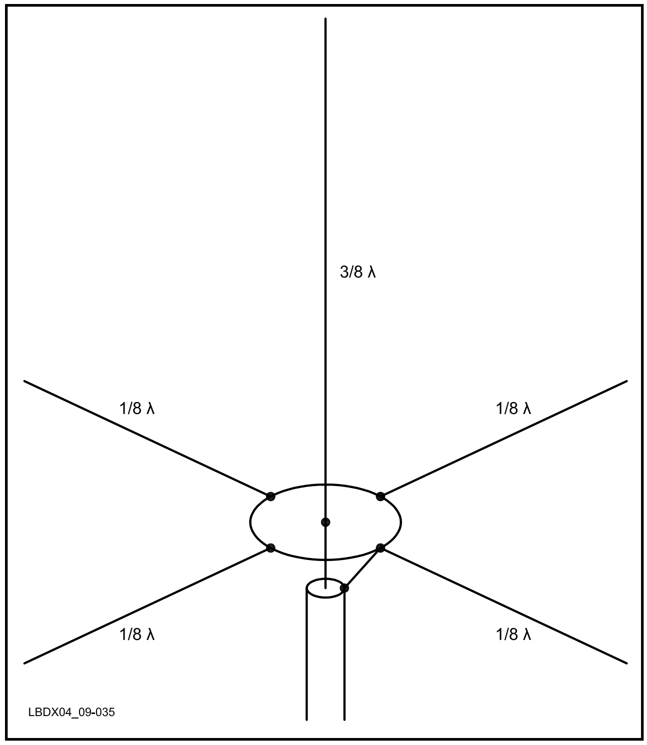


Fig 9-36 — A 3λ/8 vertical used in conjunction with 45° long radials does not require any series coil to tune the antenna, hence losses are minimized.

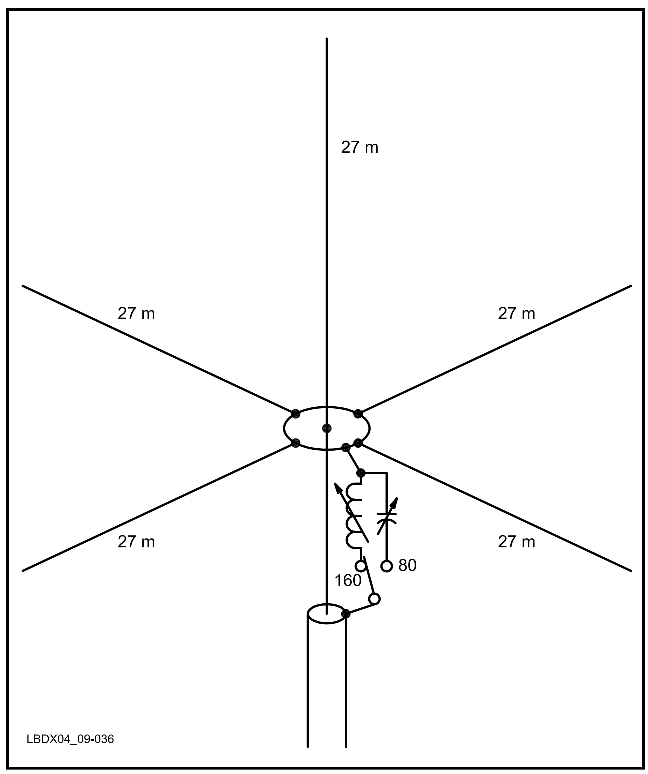


Fig 9-37 — A 27-meter long vertical with 27-meter long radial makes an excellent antenna for both 80 and 160. Band switching only requires the switching of the loading element from a coil (160 meters) to a capacitor (80 meters).

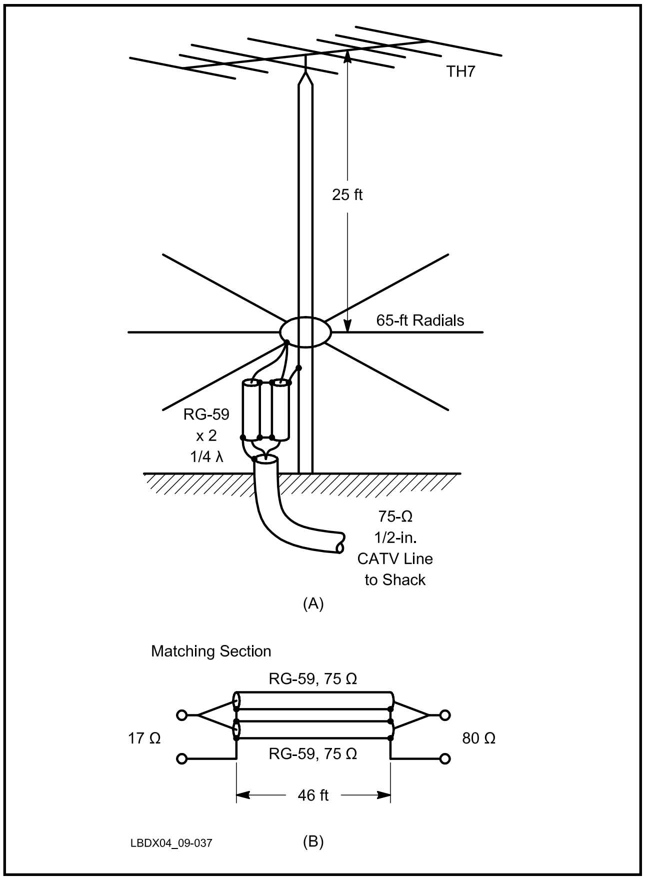


Fig 9-38 — N4KG grounded-tower feed system. The original N4KG system uses 90° long radials, which makes it necessary to adjust the vertical section of the antenna to be exactly 90° (including top loading).

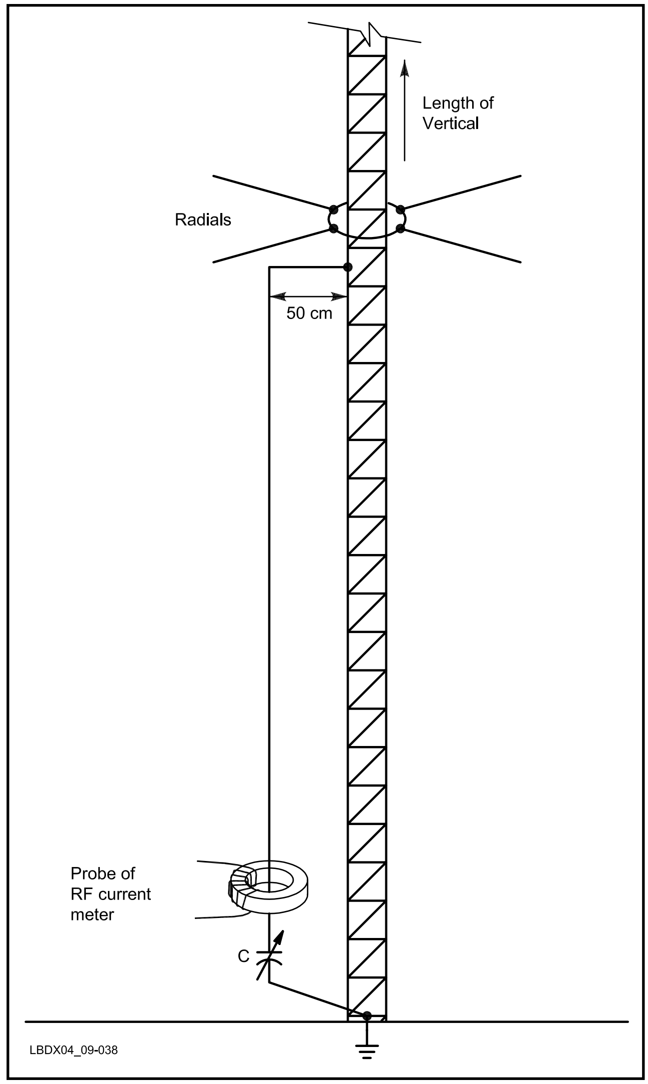


Fig 9-39 — The grounded-tower section below the antenna feed point can be made a resonant linear trap, which inserts a high impedance between the antenna feed point and the bottom of the tower. Tune capacitor for maximum in the loop.

the feed point and ground into a high impedance, effectively isolating the antenna feed point from the dc-ground rod. The trap is constructed as follows:

- Connect a shunt arm about 50 cm in length to the tower, just below the antenna feed point.
- Connect a drop wire, parallel with the tower, from the end of the arm to ground level and connect it back to the base of the tower. This forms a loop.
- Insert a variable capacitor in the drop wire (wherever convenient).
- Excite the vertical antenna (above the linear stub) with some RF.
- Use an RF current probe, such as the one described on W8JI's Web site (**www.w8ji.com/building_a_current_meter.htm**) and tune the capacitor for maximum current in the drop wire.
- You're done!

The loop tower plus drop wire plus capacitor now form a parallel-resonant circuit at the operating frequency. This ensures that no RF currents can flow through the bottom tower section to the lossy ground.

2.2.12.3. Summing Up

Using grounded towers with an elevated radial system can readily be done. The principles are simple:

- The vertical (top loaded or not) together with the radial system must be resonant
- Use the largest number of radials you can accommodate to obtain a ground-shielding effect.
- Provisions must be taken for minimum RF return current to flow in the ground. The section of the tower below the feed point should thus be decoupled.

2.2.12.4. The N4KG Reverse-Feed System

Russell feeds his design in Fig 9-38 in an unconventional way, with the center of the coax going to the radials, and the outer shield going to the vertical part. He claims this prevents arcing through from the braid of the coax to the tower. Tom coils up his parallel 75-Ω coax inside the tower leg, and that forms an RF choke. I would strongly suggest *not* to tape the coax (or the coiled coax) to the leg of the tower, especially when a linear trap (Section 2.2.12.2) is installed, since there may be a rather steep RF voltage gradient on that leg. I would keep the coax a few inches from all metal, and route it in the center *inside* the tower. In addition to the coiled coax I would certainly use a current balun made of a stack of ferrites, installed beyond the λ/4 transformer toward the transmitter. Whether or not the braid or the inner conductor goes to radials is irrelevant if a good current balun is used.

2.2.12.5. Practical Design Guidelines, Elevated Radials with Grounded Towers

If you have a grounded tower and you want to use it with an elevated radial system with four radials, you can proceed as follows:

1) Define the height where you want to have the radials. You might start at 6 meters. Convert to degrees (360° = $300/F_{MHz}$) and 6 meters = 13° on 160 meters. If you have enough physical tower height, put the radials as high as possible, since this helps reduce the near-field absorption losses from the ground.
2) Define the electrical length of the tower. Let us assume you have a 30-meter tower with a 5-element 20-meter Yagi on top. From Fig 9-81 later in this chapter we learn that this tower has an electrical length of about 123°.
3) The electrical length of the tower above the radial attaching point is 123° – 13° = 110°.
4) Cut four radials to identical electrical length as explained in Section 2.2.9.
5) Whether or not you will require a coil or a capacitor to tune the system to resonance depends on the *total* length of the antenna vertical part *plus* radials. If the length is greater than 180°, a capacitor will be required. An inductor will be required if the total length is less than 180°. Assume for this example that you use 120° long radials, so that the total antenna length is 110° + 120° = 230°. A series capacitor will be required to tune the system to resonance.
6) Measure the impedance at resonance using an antenna analyzer. If necessary use an unun or a quarter-wave transformer (or other suitable impedance matching system) to get an acceptable match to your feed-line impedance.
7) Install the linear trap on the tower section under the feed point and tune the loop to resonance by adjusting the loop variable capacitor (see procedure above).
8) You are all done!

Fig 9-40 shows the final configuration of the antenna we designed above. It is obvious that the tower must use non-conducting guys, or if steel guy wires are used they must be broken up in short lengths so that they do not interfere with the vertical antenna.

Finally, here's some perspective. Maybe it's not such a good idea after all to have elevated radials on your grounded tower because it makes things more complicated. You need a linear trap to decouple the bottom of the tower from the real ground and you need to have radials above ground. Maybe 10 or 20 radials on the ground would do the job just as well. The real reason I can see for elevated radials on a grounded tower is when that tower is electrically too long (for example, >140° rather than 90°). For this case you can shorten the tower electrically using an elevated radial system. Watch out, however, if the radial system is fairly high above ground, because the vertical radiation pattern becomes different from that of a ground-mounted vertical.

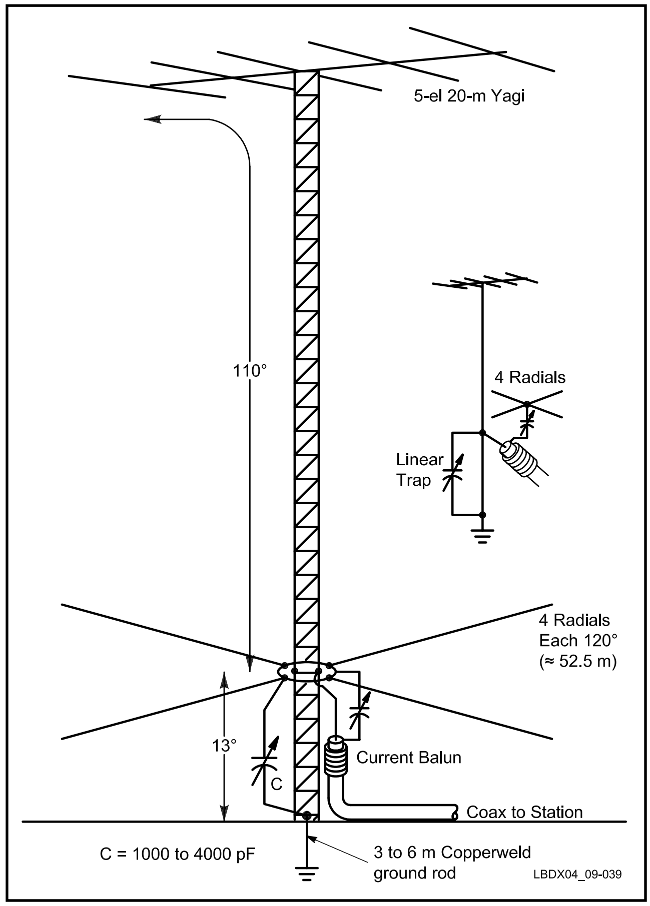


Fig 9-40 — Design example of a grounded vertical using an elevated-radial system (see text for details).

2.2.13. Elevated Radials Combined with Radial Screen on the Ground

All publications I have seen so far on the subject of elevated radials use either one of the modeling standard grounds (Average, Good, etc — see Table 5-2 in Chapter 5), or they have been done over whatever type of ground happened to be there where the tests were run.

The modeling I have done suggests that improving the ground right under the vertical and its elevated radials can increase the system gain, especially if only one to four elevated radials are used (see Section 2.2.3 and Fig 9-28). For the case of a single radial or when using radials approximately 90° long, improving the ground quality right under the antenna can greatly reduce horizontally polarized high-angle radiation and can increase the antenna gain. This can be accomplished by putting down radials or ground screens on the lossy ground.

It is important to understand that these on-the-ground radials (or screen in whatever shape) should *not* be galvanically connected to the elevated radials in any way. They should be

Fig 9-41 — The Titanex V160E antenna on the beach at 3B7RF (St Brandon Island). Note the two elevated radials about 2 meters above saltwater. The combination of one or two elevated radials with a perfect ground underneath is hard to beat.

connected to nothing, since we don't want any antenna return currents to flow in the ground.

If you have the space, and a potential 4 to 5 dB is worth the expense and effort to you, by all means provide a ground screen. In a situation where you do *not* want to use the screen for collecting antenna current, the screen does not have to have the shape of radial wires. A net of copper wires, with a mesh density measuring less than approximately 0.015 λ (1 meter on 80; 2 meters on 160), or even 0.03 λ if you are willing to sacrifice maybe 0.5 dB, is all that is needed to provide an effective near-field screen. Make sure that the crossing copper wires make good and permanent electrical connections at their joints (see Section 2.1.7).

If you use but one elevated radial, you may want to increase the ground net density in the area under that radial. In principle the screen should have a radius of λ/4 (for a λ/4 vertical), but a screen measuring only λ/8 in radius will typically be about 0.3 dB down from a λ/4 radius ground screen. Of course the saltwater environment shown in **Fig 9-41** makes for a virtually "perfect" ground screen, even though only two elevated radials were used!

For many years now, I have very successfully used λ/4 verticals in my Four Square array, each using a single λ/4 radial at about 5 meters in height. Judging an antenna's performance by the DX worked with it certainly makes no sense. But judging the same antenna's performance by the repetitive results obtained in world-class DX contests may be a good indication indeed about whether the antenna works well or not. Operated over ground that is literally swamped with copper wire, I have never scored less than a first or second place for Europe in the ARRL International DX Contest (single-band 80 meters), both CW and SSB and that is in 18 contests since 1994. In addition, I set a new European record with that antenna. Taking into account that my QTH is certainly not the best for working Ws (Normandy or the UK West Coast are better places), this means that such a vertical — even with a single elevated radial — can be a top performer.

2.2.14. Avoiding Return Currents Through the Soil

Fig 9-42 shows the vertical antenna return paths for different radial configurations. Fig 9-42A shows the case where a simple ground rod is used, where the antenna return currents have to travel entirely through the lossy soil. This reduces the radiation efficiency of the vertical to a very high degree because of the I^2R ground losses. Burying radials in the ground can greatly reduce the losses as the return currents can now travel, to a great extent (depending on the number and the length of the radials), through the low-loss radial conductors in the ground, as Fig 9-42B shows.

Fig 9-42C shows two radials elevated above ground. There are now two current return paths: the lossless path through the two radials and a lossy path through the soil.

We can minimize the currents in this parasitic path by:

- Raising the radials high above ground: Once the radials are a few meters above ground, the capacitance to the lossy soil is rather small.
- Using fewer radials: More radials means more capacitance, thus more current in the ground and hence more ground losses.
- Using more radials: More radials means a better screen. 100 radials, λ/4 long will perfectly screen the earth underneath the vertical. (This seems to contradict the previous item, but it doesn't — see Section 2.3.)
- Improving ground conductivity under the elevated radials by installing buried radials or a ground screen (not galvanically connected to the elevated radials, though!).

Another important issue is currents on the outside of the coaxial feed line. Fig 9-42D shows how unwanted currents can flow on the shield of the coaxial cable. In this situation, the coaxial feed line is just another conductor, a random-length radial. Return currents will flow in that conductor unless it is disconnected at the antenna's feed point. The question is now how can we disconnect the coaxial "radial" wire and not the coaxial feed line?

You must insert a current choke balun at the antenna feed point (see Fig 9-42E). The high impedance the current balun presents to any currents on the outside of the coax shield effectively suppresses common-mode currents on the cable. Several types of current baluns are described in Chapter 6, Section 7. If you are forced to use (for layout reasons) 3λ/4 feed lines in a Four Square array, you will wind up with a lot of surplus coax length. Wind it all up in a coil and mount it as close as possible to the antenna feed point. This makes an

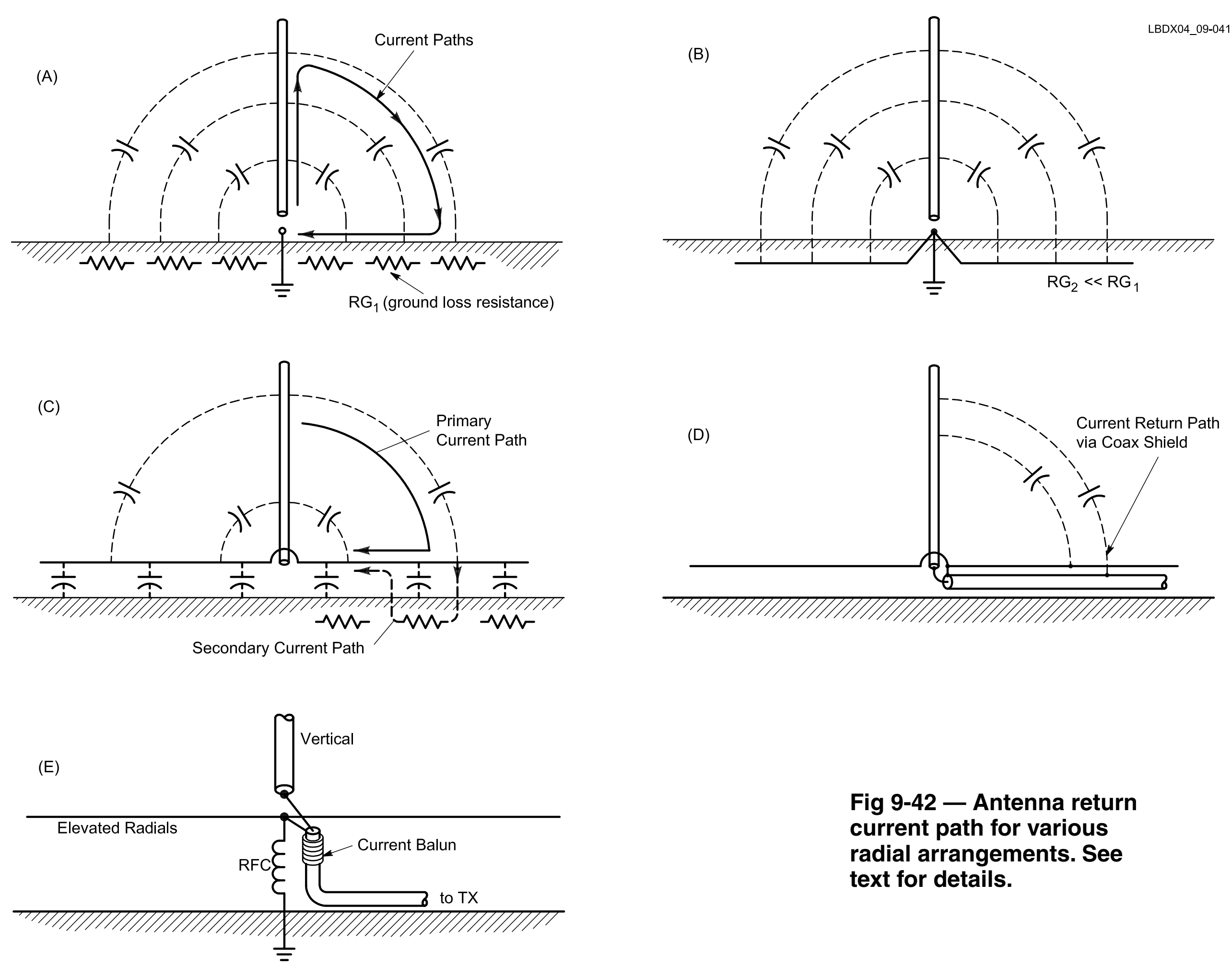


Fig 9-42 — Antenna return current path for various radial arrangements. See text for details.

excellent choke balun. It is always better to run the coax on or preferably in the ground, rather than supported on poles at a certain height, to prevent coupling and parasitic currents on the outer shield.

It also makes common sense to provide a dc ground for the common radial points. You can do this by connecting an RF choke (100 μH or more) between the radial common point and a safety ground rod below the antenna feed point, as shown in Fig 9-42E. In any case do not connect the radial common point of an elevated radial system to the ground (unless through an RF choke).

If you use only a few radials, each of them can radiate considerable near-field energy. They can induce currents on the feed line beyond where the choke balun has been inserted at the feed point. Burying the feed line can improve this situation. Feed lines supported off the ground are very sensitive to this kind of coupling. If you use only two radials, run the feed line at right angles to the two in-line radials. In other words, keep the feed line away from the near fields of the radials.

When using a number of elevated radials (eg >20), it is in theory unnecessary to use a current balun since the screening effect of the radials will be sufficient to prevent common-mode antenna-return currents of any significant magnitude to flow on the coax outer shield. But I would still use one…

2.2.15. Elevated Radials in Vertical Arrays

When a vertical is used as an element in an array, an additional parameter arises when choosing the ideal radial length, at least if you are concerned about reducing horizontally polarized high-angle radiation of the array to a minimum. Careful layout of the radials is very important. Never run radials belonging to two different array elements in parallel. Design your layout such that coupling is minimized.

Zero coupling is of course achieved by using buried radials, terminated in bus bars where radials of adjacent elements meet one another. (See Chapter 11, Section 8.) I should point out that if you use four 90° long radials on each element of an array, and have them laid out in such a manner that coupling does not exist between radials of adjacent elements, it may be just as good to use a single radial!

2.3. Buried or Elevated, Final Thoughts

It is clear, and it has been proven over and over in the real world, that an elevated radial system at a relatively low height is a valid alternative for a system of buried radials, *if there is a good reason you can't put down a decent radial system in or on the ground.* If you use only a small number of radials, perhaps 1 to 8, their task will be almost exclusively to efficiently collect the return currents of the vertical, and you will have to suffer substantial near-field losses in the ground, up to 5 dB. With a larger number the screening effect becomes important and near-field ground losses can be reduced by making use of the screening effect of a large number of radials. Elevated radials can have advantages such as:

- Providing the possibility of installing a decent ground system under very unfriendly circumstances, such as over rocky ground.
- Easy to install and very efficient for DXpeditions where used near saltwater.
- More flexibility in matching, since the real ground is not resonant. An elevated radial system using only a few radials — maximum of four — can be made inductive or capacitive, which may be an asset in designing a matching system.

For using elevated radials I would propose the following guidelines:

- Put the radials up as high as possible.
- Use as many radials as possible, since this makes the radial system non-resonant.
- If you use a small number (<16), install a ground screen.

If you have the space and if the ground is not too unfriendly, I would suggest you use buried radials however.

2.4. Evaluating the Radial System

Evaluating means measuring antenna field strength (FS), or measuring certain parameters for which we know the correlation with radiated FS. You cannot truly evaluate an antenna just by modeling it. You can develop, design and predict performance by modeling, but you cannot evaluate the actual performance of the antenna on a computer. However, there are some indirect measurements and checks that can and should be done.

2.4.1. Evaluating a Buried-Radial System

The classic way to evaluate the losses of a ground system is to measure the feed-point resistance of the vertical while steadily increasing the number of radials. The feed-point resistance will drop consistently and will approach a lower limit when a very good ground system has been installed. Be aware, however, that the intrinsic ground conductivity can vary greatly with time and weather, so it is recommended that you do such a test over a short time frame to minimize the effects of varying environmental factors on your tests (Ref 818, 819).

Peter Bobeck, DJ8WL, (now a Silent Key) performed such a test on his 23-meter-long top-loaded (T) antenna. He added 50-meter-long radials (on the ground) while measuring the feed-point impedance and found the following:

No. of radials	2	5	8	14	20	30	50
Impedance, Ω	122	66	48	39	35	32	29

Incidentally, eight radials look like a perfect match to 50-Ω coax, but the system efficiency for that case was below 50%!

Don't be surprised if the impedance gets lower than 36 Ω with a full-size λ/4 vertical. It first surprised me when I measured about 20 Ω for my 160-meter full-size λ/4 vertical made with a freestanding tower, but that was because of its very large effective diameter and tapering cross section.

For calculating antenna efficiency, you can use the values from Table 9-1 that lists the equivalent resistance of buried radial systems in good-quality ground. For poor ground, higher resistances can be expected, especially with only a few radials.

Measuring the impedance of a vertical and watching it decrease as you add radials tells us nothing about the near-field absorption ground losses. It only gives us an indication of the I^2R losses that determine return-current collecting efficiency.

Periodic visual inspections of the radial system for broken wires and loose or corroded connections will assure continued efficient operation. If you bury the radials, it is a good idea to make them accessible anyhow just where they connect to the bus bar. This way you can periodically check with a snap-on current meter if the radial still carries any current on transmit. If it doesn't, maybe the radial is broken at a short distance from the connection point.

2.4.2. Evaluating an Elevated-Radial System

Whether you have 1, 2 or 16 elevated radials, if these radials are the only antenna-current return paths (that is, the elevated radials are *not* connected to the lossy ground), the measured real part of the antenna impedance will not change. There is no gradual decrease of feed-point impedance as you increase the number of radials.

Measuring the antenna impedance does *not* give you any indication of near-field absorption ground losses. The only test you can perform on an elevated radial system is to measure the radial current, although this has little, if any, correlation with low-angle field strength. Nevertheless, when using only a few radials (2 to 8) it is a good idea to check the radial currents, and to make sure they are similar (within a few percent of one another).

Do regular inspections of your current balun. I would recommend to periodically measure its effectiveness by checking its impedance (R + *j*X). This should be measured at the operating frequency.

3. SHORT VERTICALS

We usually consider verticals as being *short* if they are physically shorter than λ/4. Short verticals have been described in abundance in the amateur literature (Ref 771, 794, 746, 7793 and 1314). Gerd Janzen published an excellent book on this subject, *Kurze Antennen* (in German). Unfortunately, this was completely based on antenna modeling, where in my opinion real-world measured results are greatly lacking (Ref 7818).

The radiation pattern of a short vertical is almost the same as that for a full-size λ/4 vertical. The small difference in peak gain (over a theoretical perfect ground at zero degree elevation) is also reflected in a small difference in radiation pattern shape. The area covered by the patterns are identical: all these antennas have the same average gain (averaged over all elevation angles). **Fig 9-43** shows the vertical radiation patterns of a range of short verticals over perfect ground, calculated using *EZNEC*.

If those short verticals over perfect ground are in essence almost as good as their full-size (λ/4) counterparts, why aren't we all using short verticals? A short monopole exhibits a feed-point impedance with a resistive component that is much smaller than 36.6 Ω and a reactive component that is highly capacitive. These two factors can make a short vertical more difficult to handle than a bigger one. To feed a short vertical with low losses using a coaxial feed line, you must first get rid of the reactive part and increase the real part of the feed

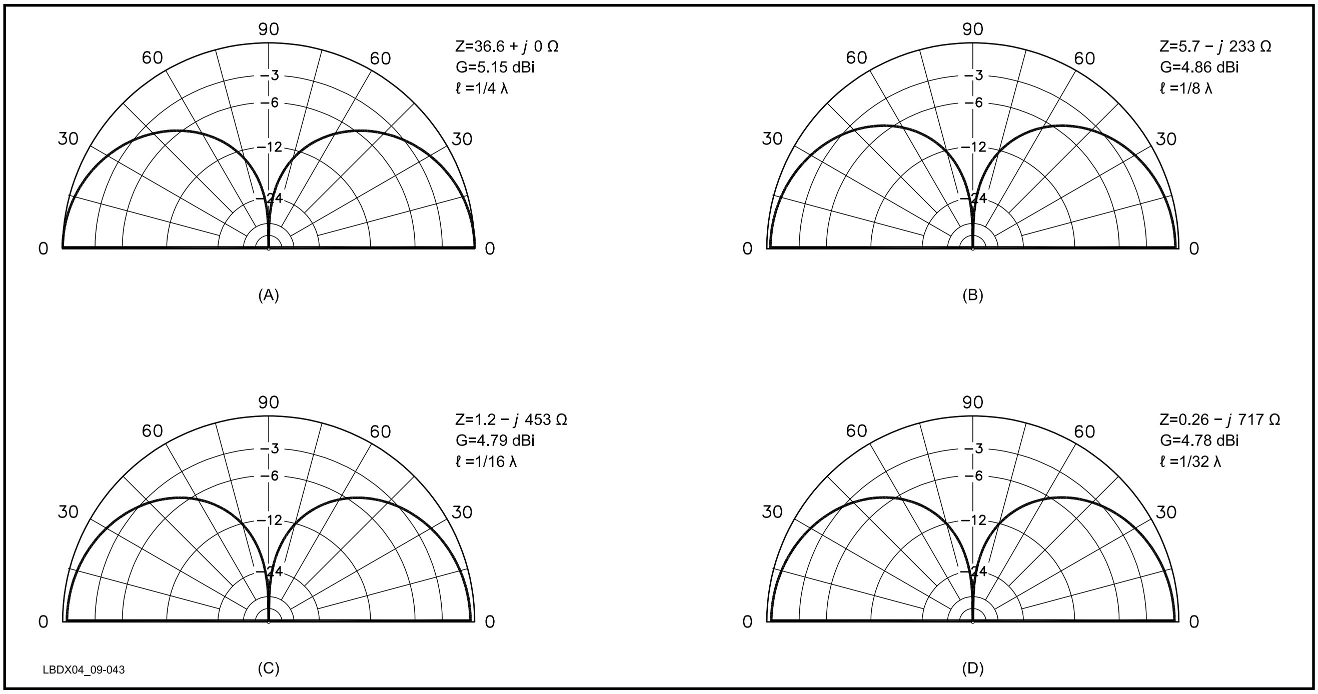


Fig 9-43 — Elevation-plane radiation patterns and gain in dBi of verticals with different heights. The 0-dB reference for all patterns is 5.2 dBi. Note that the gain as well as the shape of the radiation patterns remains practically unchanged with height differences. The patterns were calculated with *EZNEC* over perfect ground, using a modeling frequency of 3.5 MHz and a conductor diameter of 2 mm. At A, height = λ/4. At B, height = λ/8. At C, height = λ/16. At D, height = λ/32.

impedance up to 50 Ω. This requires *loading* and *matching* the vertical and these can greatly impact efficiency (see also Chapter 8, Section 2.3).

Short verticals can be loaded to be resonant at the desired operating frequency in different ways. Various loading methods will be covered in this section, and the radiation resistance for each type will be calculated. Design rules will be given, and practical designs are worked out for each type of loaded vertical. Different loading methods will be compared in terms of efficiency.

Loading a short vertical means canceling the reactive part of the impedance to bring the antenna to resonance. The simplest way is to add a coil at the base of the antenna, a coil with an inductive reactance equal to the capacitive reactance shown by the short vertical. This is the so-called *base-loading method*. **Fig 9-44** shows a number of classic loading schemes for short verticals, along with the current distribution along the antenna. Remember from Section 1.2 that the radiation resistance is a measure of the area under the current-distribution curve. Also remember from Section 1.3 that the radiation efficiency is given by:

$$\text{Eff} = \frac{R_{rad}}{R_{rad} + R_{loss}}$$

The real issues with short verticals are *efficiency* and *bandwidth*. Let us examine these issues in detail. With short verticals the numerator of the efficiency formula decreases in value (smaller R_{rad}), and the term R_{loss} in the denominator is likely to increase (losses of the loading devices such as coils). This means we have two terms, which tend to decrease the efficiency of loaded verticals. Therefore maximum attention must be paid to these terms by

- Keeping the radiation resistance as high as possible (which is *not* the same as keeping the feed-point impedance as high as possible).
- Keeping the losses of the loading devices as low as possible. Maximum radiation resistance occurs when current integrated over the vertical section is as high as possible, which means maximum current mid-height in the vertical section. With very short verticals the current distribution is almost constant and the exact position of the maximum becomes irrelevant.

3.1. Radiation Resistance

The procedure for calculating the radiation resistance was explained in Section 1.2, where we found that for a λ/4 vertical made with a very small size conductor is 36.6 Ω. (See Fig 9-44).

It is important to have the radiation resistance as high as possible, for three reasons:

1) To keep the radiation efficiency as high as possible: $\eta = R_{rad} / (R_{rad} + R_{loss})$.
2) To keep the efficiency and the bandwidth of a matching network (to the 50-Ω line) as high as possible.
3) To keep the overall (antenna plus matching network) SWR bandwidth as high as possible.

We will now analyze the following types of short verticals, all of which are about 30% of full-size quarter-wave (approximately 12 meters high on 160 meters) or 27.5° long:

1) Base loaded.
2) Top loaded.
3) Center loaded.

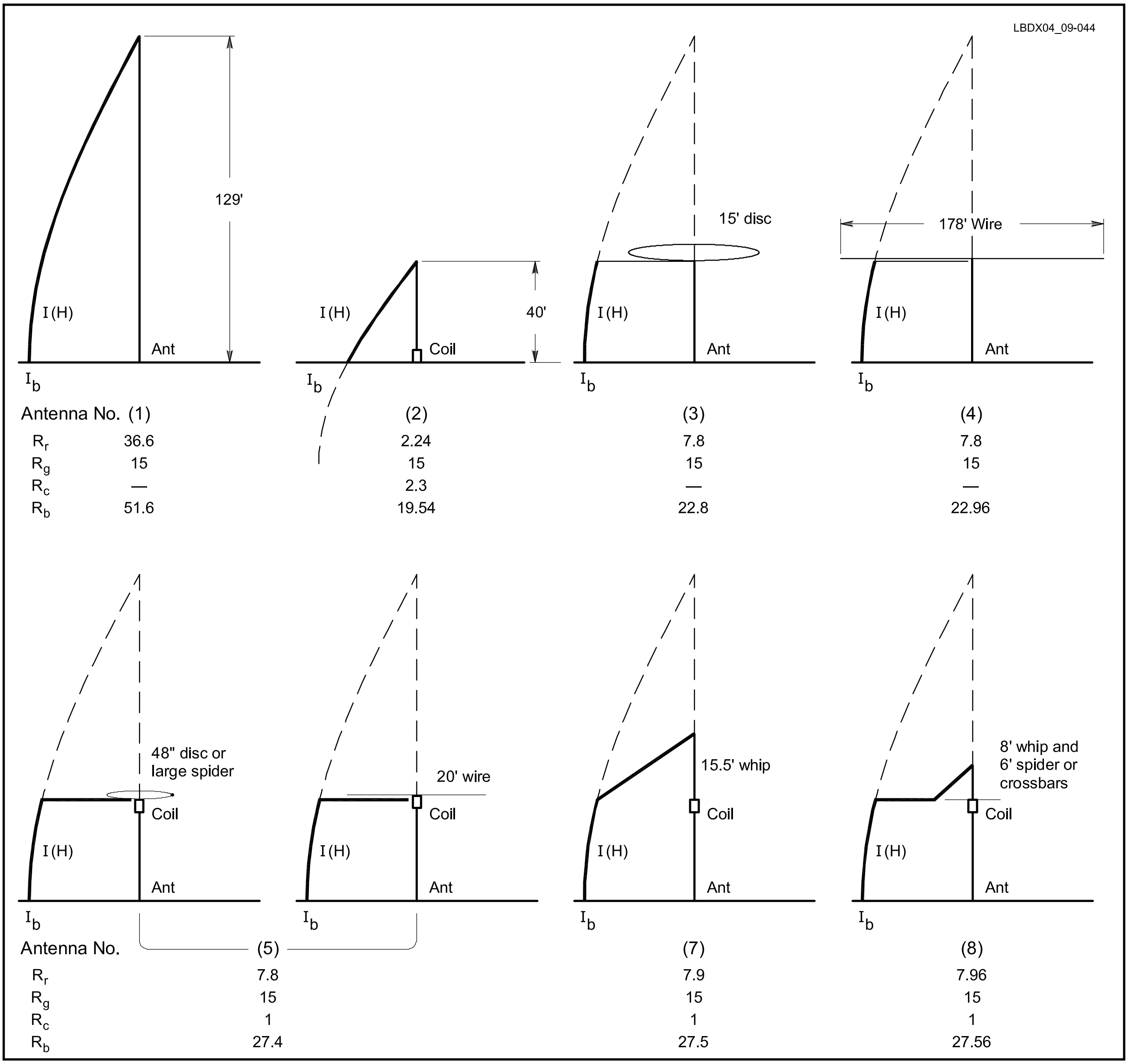


Fig 9-44 — The antennas described in the text are shown with their current distributions, radiation resistances R_r, assumed ground loss resistance R_g, coil loss R_c (if any), total base input resistance R_b, base current I_b for 1000-W input to the antenna, and finally radiating efficiency in % (Source: "Evaluation of the Short Top Loaded Vertical" by C. Michaels, W7XC, *QST* March 1990.)

4) Base plus top loaded.
5) Linear loaded.

3.1.1. Base loading

The radiation resistance can be calculated as defined in Section 1.2. A trigonometric expression that gives the same results, is given below (Ref 742).

$$R_{rad} = 36.6 \times \frac{(1 - \cos L)^2}{\sin^2 L} \qquad \text{(Eq 9-7)}$$

where L = the length of the monopole in degrees (1 λ= 360°).

Example: length of vertical: 12.5 meters, frequency: 3.8 MHz. On 3.8 MHz the antenna is approximately 60° long. So, R_{rad} = 12.2 Ω. This value is confirmed by the graph shown in Fig 9-8, and by modeling the antenna using *EZNEC*.

3.1.2. Top Loading

The patent for the top-loaded vertical was granted to Simon Eisenstein of Kiev, Russia, in 1909. **Fig 9-45** is a copy of the schematic which was part of the original patent application, where you can see a combined loading coil plus top-hat loading configuration. The resulting current distribution is also shown.

The tip of the vertical antenna is the place where there is no current, and maximum voltage. This is the place where *capacitive* loading is most effective, and *inductive* loading (loading coils) is least effective. In some cases, inductive loading is combined with capacitive top loading. Top loading is achieved by one of the following methods (see **Fig 9-46**).

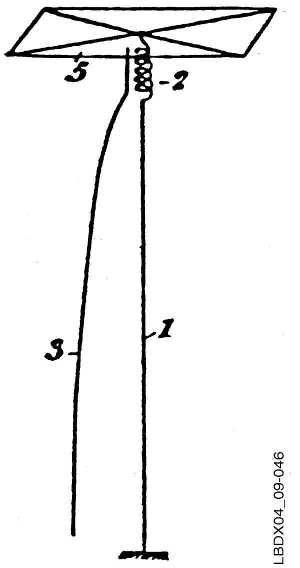


Fig 9-45 — Replica of the schematic patent application of August 10, 1909, showing the original drawing of the top-loaded vertical.

- **Capacitance top hat:** In the shape of a disk or the spokes of a wheel at the top of the shortened vertical. Details of how to design a vertical with a capacitance hat are given in Section 3.6.2. On the low bands the top hat is often made of a number of sloping wires (see Section 3.6.2.2) because this is easier to install than a large flat hat in the form of a wheel with spokes.
- **Flat-top wire loading (T loading):** The flat-top wire is symmetrical with respect to the vertical. Equal currents flowing outward in both flat-top halves essentially cancel the radiation from the flat-top wire. For design details see Section 3.6.2.3. Here too the wires are often installed in a sloping fashion.

For calculating the radiation resistance of the top-loaded vertical, it is irrelevant which of the above loading methods is used. In other words, For a given vertical height, all achieve the same radiation resistance. As can be seen in Fig 9-44, the various forms of top loading (3, 4 and 5) all result in the same "current area" below the top loading device.

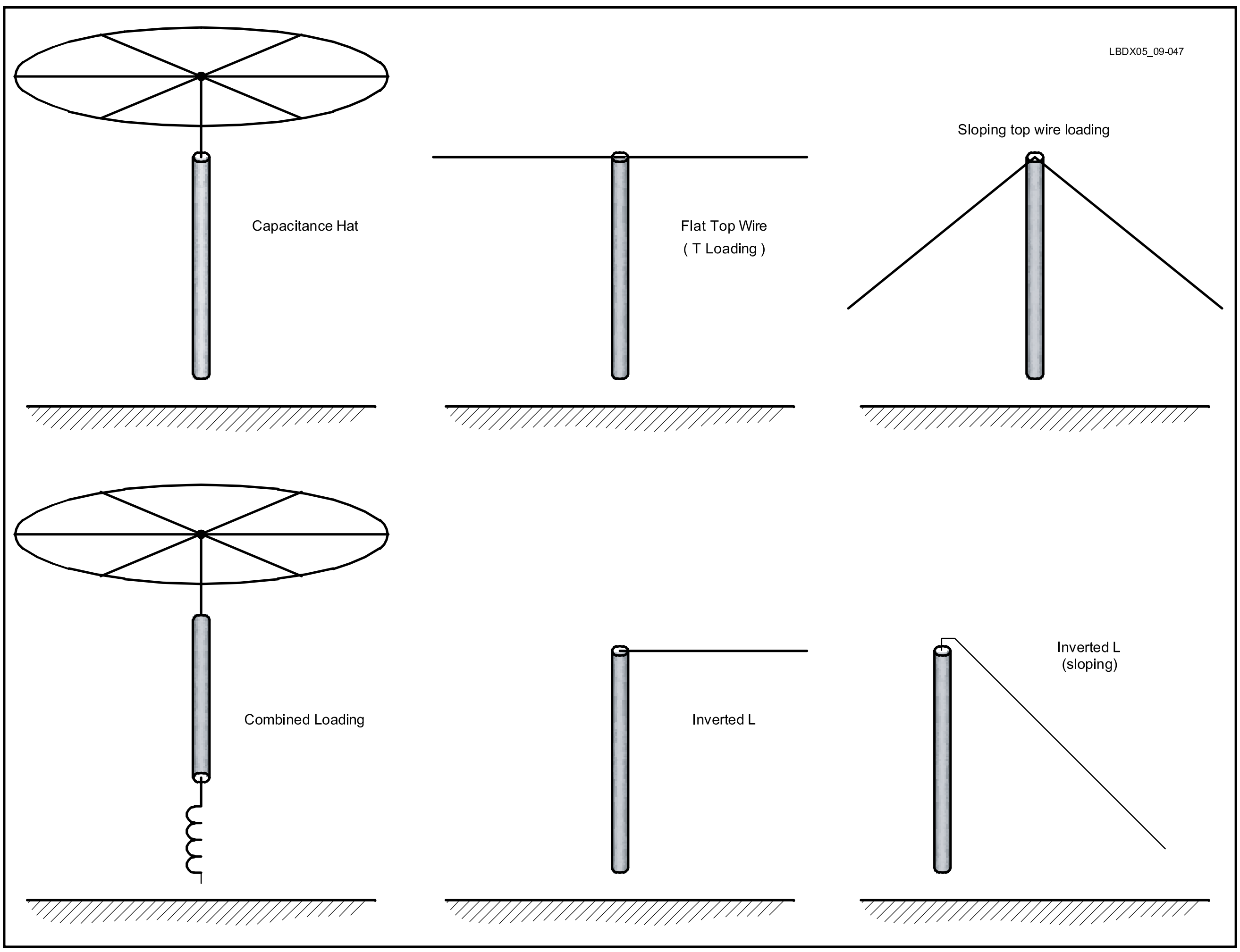


Fig 9-46 — Common types of top loading for short verticals. The inverted L and sloping inverted L are not true verticals, since their radiation patterns contain horizontal components.

However, when we deal with efficiency (where both R_{rad} and R_{loss} are involved) the different loading methods may behave differently because of different loss resistances. The radiation resistance can be calculated as defined in Section 1.2.

3.1.2.1. Top Loading by Capacitance Hat

Let us work out an example using *EZNEC*. Frequency = 3.8 MHz and length of vertical = 12.5 meters (60° on 3.8 MHz). The chart in Fig 9-8 gives R_{rad} = 27.6 Ω for a 60° long top loaded vertical. *EZNEC* tells us that a top hat with four spokes (diameter 25 mm) and a diameter of 7.65 meters resonates the 12.5-meter vertical on 3.8 MHz, under which conditions R_{rad} = 27.5 Ω.

Large surface capacitance hats in the form of disks are not often used because they are difficult to make and to keep up. Identical results can be obtained by using just two top loading wires, making the antenna to look like a T.

3.1.2.2. Top Loading by Flat Top Wire (T-Antenna) or Sloping T Wires

As with a capacitance hat, symmetric T-loading wires (equal length, in-line) do not produce any radiation in the far field (canceled because of the current distribution in the wires).

Genuine flat top loading wires (perfectly horizontal) are not often used, because they are difficult to build especially when a large degree of loading is required. Sloping loading wires are very popular as they can serve as guy wires for the vertical antenna at the same time.

Table 9-8 later in this chapter shows R_{rad} for two short verticals using top loading with four wires (from horizontal to steeply inclined).

3.1.2.3. Inverted L

This configuration is not really a top-loaded vertical, since the horizontal loading wire radiates along with the vertical mast to produce both vertical and horizontal polarization. Inverted-L antennas are covered separately in Section 7.

3.1.3. Center Loading

The radiation resistance can be calculated as defined in Section 1.2. A trigonometric expression that gives the same results is shown below (Ref 42 and 7993):

$$R_{rad} = 36.6 \times \left(1 - \sin^2 t2 + \sin^2 t1\right) \qquad \text{(Eq 9-10)}$$

Let's work out an example with the same data: Frequency = 3.8 MHz and length of vertical = 12.5 meters (60° on 3.8 MHz).

t1 = length of vertical below loading coil (30°)

t2 = 90° – length of vertical above loading coil (30°) = 60°

The formula gives us R_{rad} = 18.3 Ω. The chart from Fig 9-8 shows 17 Ω. Modeling with *EZNEC* (*NEC-2*) with a coil in the center of the vertical yields R_{rad} = 20.8 Ω. Although the values are not exactly the same, the trend is clear: base loading gives the lowest R_{rad}, top loading the highest and center loading a value in between.

3.1.4. Combined Top and Base Loading

Top and base loading are quite commonly used together, as shown in **Fig 9-47**. Top loading is often done with capacitance-hat loading, or even more frequently in the shape of two or more (possibly sloping) flat-top wires.

If a wide frequency excursion is required (eg, 3.5 to 3.8 MHz), you can load the vertical to resonate at 3.8 MHz using the top-loading technique. When operating on 3.5 MHz, a little base loading is added to establish resonance at the lower frequency.

Example: We have a vertical measuring 12.5 meters high (effective diameter 25 cm), using a capacitance hat comprising four spokes each 3 meters long.

Modeling with *EZNEC* tells us that this antenna resonates at approximately 4.2 MHz. To make it resonate on 3.8 MHz, we need a small base loading coil with a reactance of 38.7 Ω (L = 1.01 μH). R_{rad} is then 22 Ω. To resonate the same antenna on 3.5 MHz the loading coil becomes 3.1 μH (X_L = 68 Ω). In the CW band R_{rad} is 17.6 Ω.

As the value of the base loading coils is very small, the coil losses are very small as well. On 3.5 MHz a Q of 300 causes an equivalent loss resistance of 38.7/300 = 0.13 Ω.

Fig 9-48 shows the radiation resistance for monopoles with combined top and base loading. The physical length of the antenna (L) plus top loading (T) plus base loading (B) must

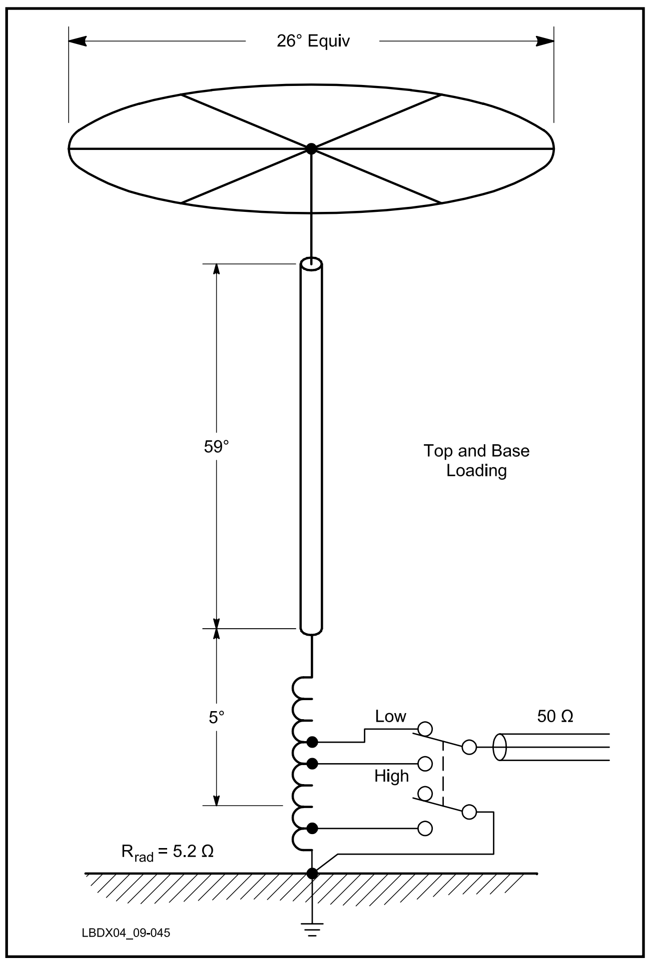


Fig 9-47 — Instead of series-feeding the antenna, we can look for a tap on the coil that gives 50 Ω. The coil serves two purposes: Some base loading and also impedance matching. Using a DPDT relay you could make provisions for a perfect 50-Ω match on two frequencies, for example one on CW and one on phone.

total 90°. In the above example L ~ 60°, T ~ 22° and B ~ 8° on 3.5 MHz. The value we read from the chart is 22 Ω, exactly the same as what we calculated with *EZNEC*.

When the antenna has a large capacitance hat compared to the distributed capacitance of the structure, there is no reason to put the required loading coil high on the structure. Current distribution will be essentially the same no matter where you put the coil, even when the antenna is *far* from self-resonance with just the hat. We can simply use a large hat and put a coil at the base, where it can do double-duty for impedance matching and loading, and we can reach it easily for adjustment, as shown in Fig 9-47.

3.1.5. Linear Loading

Linear loading is defined as replacing a loading coil at a given place in the vertical with a linear-loading section, which resembles a shorted stub, at the same place in the vertical. This places the two conductors of the loading device in parallel with the radiating element. Due to the current *not* being out-of-phase in the loading device, the device will radiate. The R_{rad} of the antenna will be slightly higher than if we were using a loading coil in the same place.

This linear-loading technique described above is used on the Hy-Gain 402BA shortened 40-meter beam, where linear loading is used at the center of the dipoles. It is also used successfully on the KLM 40 and 80-meter shortened Yagis and dipoles, where linear loading is applied at a certain distance from the center of the elements, but where the linear loading devices were *not* parallel to the elements, introducing some unwanted radiation. This reduced the directional characteristics of the antenna.

In recent years the better Yagi designs for 80 meters have employed optimized high-Q loading coils rather than linear-loading devices, with great success (see Chapter 13).

3.2. Keeping the Radiation Resistance High

As stated before, this is *not* the same as keeping the feed-point impedance high! Using any kind of transformers, such as folded elements, or any other type of matching systems do *not* change the radiation resistance. The rule for keeping the radiation resistance as high as possible is simple:

- Use as long a vertical as possible (up to 90°).
- Use top-capacitance loading rather than center or bottom loading. Fig 9-48 gives the radiation resistance for monopoles with combined base and top loading. The graphs clearly show the advantage of top loading.

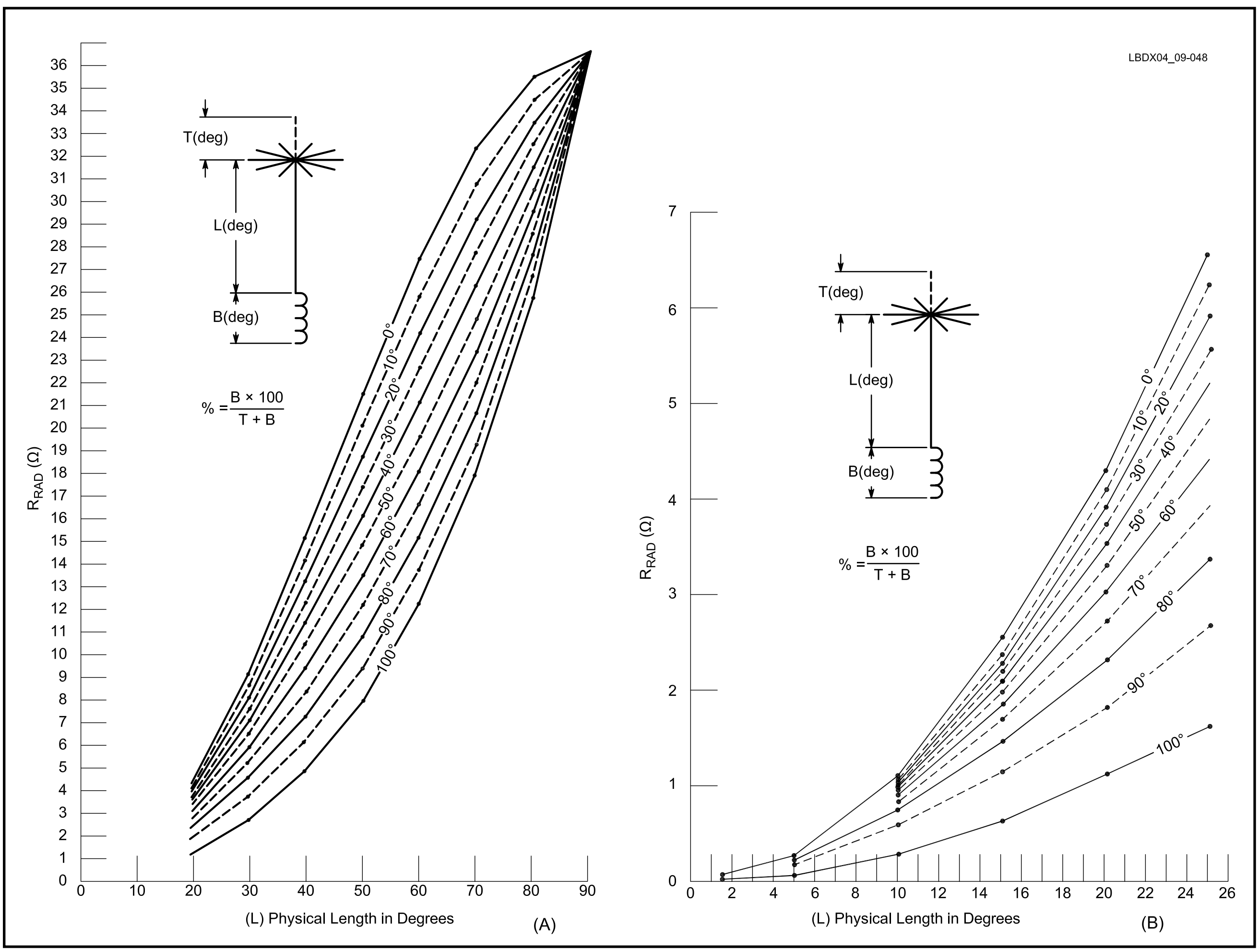


Fig 9-48 — Radiation resistances of a monopole with combined top and base loading. Use the chart at B for shorter monopoles to obtain better accuracy.

The values of R_{rad} given in these figures can be used for antennas with diameters ranging from 0.1° to 1° (360° = 1 λ). J. Sevick, W2FMI (Silent Key), (Ref 818) obtained very similar results experimentally, while the values in the figures mentioned above were derived mathematically.

For a given physical size, the way to maximize efficiency is to make current as large and uniform as possible over the maximum available vertical distance. The solution is to end-load the antenna with a large hat or some other form of termination that does *not* return to earth. The only thing fancy shunt tuning schemes or multiple drop wires do is to make the feed line see a new impedance.

Top loading with sloping wires is attractive from a mechanical point of view, but don't build a cage of sloping top hat wires around the vertical. A good rule of thumb that can be used is never to drop the ends of the top loading wires lower than half the height of the vertical.

3.3. Keeping Losses Associated with Loading Devices Low

Capacitance hat (disk): The losses associated with a capacitance hat are negligible. When applying top-capacitance loading, especially on 160 meters, the practical limitation is likely to be the size (diameter) of the top hat. Therefore, when designing a short vertical it is wise to start by dimensioning the top hat.

T-wire top loading: This method is lossless, as with the capacitance hat. It may not always be possible, however, to have a perfectly horizontal top wire. Slightly drooping of top-loading wires is just as effective, and when used in pairs (each wire of a pair being in-line with the second wire) the radiation from these loading wires is negligible.

If we use top loading wires that droop steeply, we build a cage around the vertical, which will reduce the R_{rad}. A perfect non-leaking case around it would yield a R_{rad} of 0 Ω. More on this subject in Section 3.6.2.2.

Loading coil: Even large loading coils are intrinsically lossy. The equivalent series loss resistance is given by:

$$R_{loss} = \frac{X_L}{Q} \qquad \text{(Eq 9-12)}$$

where

X_L = inductive reactance of the coil

Q = Q (quality) factor of the coil

Base loading requires a relatively small coil, so the Q losses will be relatively low, but the R_{rad} will be low as well. See Section 3.6 for practical design examples with real-life values. Top loading requires a large-inductance coil, with correspondingly larger losses, while in this case the R_{rad} is much higher.

As mentioned above, unloaded Q factors of 200 to 300 are easy to obtain without special measures. Well-designed and carefully built loading coils can yield Q factors of up to 800 (Ref 694 and 695). W8JI, wrote: "The most detailed and accurate loading inductor text readily available to amateurs appears in the chapter 'Reactive Elements and Impedance Limits' in Kuecken's book *Antennas and Transmission Lines* (Ref 696). I've measured hundreds of inductors. A typical B&W Miniductor or Airdux coil of #12 wire operated far from self-resonance with a form factor of 2:1 L/D has a Q in the 300 range. Optimum Q almost always occurs with bare wire space wound one turn apart, but optimum L/D can range from 0.5 to 2 or more depending on how far below self-resonance you operate the inductor and what is around the inductor and how big the conductors in the coil are. Large optimal edge-wound or copper tubing coils can get into the Q ~800 range. I've never in my life seen an inductor of reasonable reactance above that Q, and very few make it that high."

Linear-loading: W8JI measured the Q of typical linear loading devices and found an amazing low figure of between 50 and 100, while loading coils of moderate quality easily reach an unloaded Q of 200 and well-designed and optimized coils may reach a Q of well over 400. Tom, W8JI remarks: "For example, the Q of a 400 ohm reactance with a #14 folded wire stub is much less than 100. I can easily obtain a Q of 300 with the same size wire in a conventional coil."

3.4. Short-Vertical Design Guidelines

From the above considerations we can conclude the following:

- Make a short vertical physically as long as possible.
- Make use of top loading (capacitance hat. horizontal T wires or slightly sloping T wires) to achieve the highest radiation resistance possible.
- Use the best possible radial system.
- Design and build your own loading coils with great care (high Q). Airdux TL coils from B&W (**www.bwantennas.com/coils/coilcat.htm**) can be used but are relatively expensive.
- Take extremely good care of electrical contacts — contacts between antenna sections and between the antenna and the loading elements. This becomes increasingly important when the radiation resistance is low.

Though you may be able to build small verticals with low intrinsic losses, it may not always be possible to improve the losses in the ground-return circuit (radials and ground) to a point where a small loaded vertical achieves good efficiency. Small loaded verticals will often be imposed by area restrictions, which may also mean that an extensive and efficient ground (radial) system may excluded. Keep in mind that with short loaded verticals, the ground system is even more important than with a full-size vertical.

It is a widespread misconception that vertical antennas don't require much space. Nothing is farther from the truth. Verticals take a lot of space! A good ground system for a short vertical takes much more space than a dipole, unless you live right at the coast, over saltwater, where you might get away with a simple ground system.

3.4.1. Verticals with Folded Elements

Another common misconception is that folded elements increase the radiation resistance of an antenna, and thus increase the system efficiency. However, the radiation resistance of a folded element is not the same as its feed-point resistance.

A folded monopole with two equal-diameter legs will show a feed-point impedance with the resistive part equal to $4 \times R_{rad}$. The higher feed-point impedance does not reduce the losses due to low radiation resistance, however, since with the folded element the lower feed current now flows in one more conductor, totaling the same loss. In a folded monopole, the same current ends up flowing through the lossy ground system,

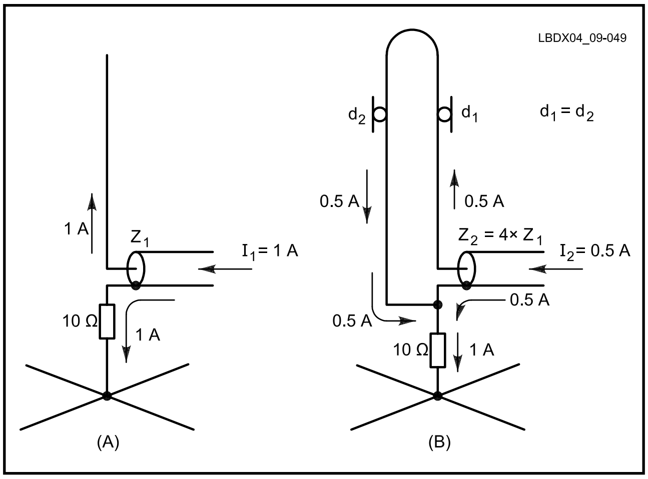


Fig 9-49 — The same net current flows in the ground system, whether an open or a folded element is used. This is clearly illustrated for both cases. See text for details.

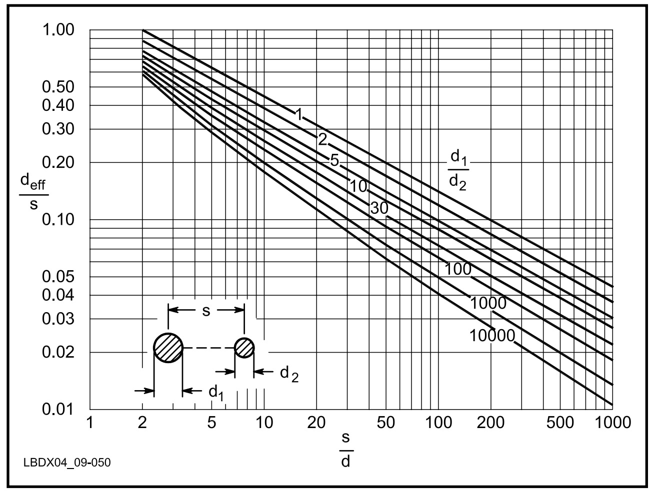


Fig 9-50 — Normalized effective antenna diameters of a folded dipole using two conductors of unequal diameter, as a function of the individual conductor diameters d1 and d2, as well as the spacing between the two conductors (S). (After Gerd Janzen, *Kurze Antennen*.)

resulting in the same loss whether a folded element is used or not.

This is illustrated in **Fig 9-49**. In the non-folded situation in Fig 9-49A it is clear that the total 1 A current flows through the 10-Ω equivalent ground-loss resistance. The ground loss is $I^2 \times R = 10$ W.

Figure 9-49B shows the folded-element situation. In this example equal-diameter conductors are assumed; hence the feed impedance is four times the impedance of the single-conductor-equivalent vertical, and the current is half the value of the same antenna with a single conductor. Thus, 0.5 A flows in the folded-element wire and from the feed point down to the 10-Ω resistor. There is another 0.5 A coming down the folded wire and also going to the top of the 10-Ω resistor. In the ground system through the 10-Ω ground loss resistor, we have a total current of 1 A flowing, the same as with the unfolded vertical. The loss is again $I^2 \times R = 10$ W.

In other words, the impedance transformation of the folded monopole also transforms the ground loss part of the equation in the same way as it does for the radiation resistance, and there is no net improvement. It is just another form of transformer and is no different than adding a toroidal step-up transformer at the base of a regular monopole.

Although the folded monopole does not gain anything in efficiency due to the impedance transformation, it does have some advantages. The impedance transformation will result in a higher impedance that might be more easily matched by a more efficient network than would be required by a plain monopole. The folded monopole has some advantages in lightning protection due to the possibility of dc grounding the structure. And the folded monopole may have a wider bandwidth due to the larger effective diameter of the two conductors (see also Chapter 8, Section 1.4.1).

Fig 9-50 shows the effective normalized diameter of two parallel conductors, as a function of the conductor diameters and spacing (from *Kurze Antennen*, by Gerd Janzen, ISBN 3-440-05469-1). A folded element consisting of a 5-cm OD tube and a 2-mm OD wire (d1/d2 = 25), spaced 25 cm has an effective round conductor diameter of 0.6 × 25 = 15 cm.

3.5. The SWR Bandwidth of Short Verticals

When hams talk about the bandwidth of an antenna they usually refer to the SWR bandwidth. A practical way of characterizing performance is to model the antenna at different frequencies, using software such as *MININEC* or *EZNEC*. The Q of the vertical is a clear indicator of bandwidth. Antenna Q and SWR bandwidth are discussed in Chapter 5, Section 2.14

Table 9-5 shows the results obtained by modeling full-size quarter-wave verticals of various conductor diameters. Both the perfect and the real-ground case are calculated. The vertical with a folded element clearly exhibits a larger SWR bandwidth than the single-wire vertical. Note that with a tower-size vertical (25-cm diameter), both the CW and the phone DX portions of the 80-meter band are well covered. If a wire vertical is planned (eg, suspended from trees), the folded version is to be preferred. Matching can easily be done with an L network.

It is evident that loaded verticals exhibit a much narrower bandwidth than their full-size λ/4 counterparts. For a given antenna height, the loading system that results in the highest R_{rad} will also give the greatest SWR bandwidth. With short verticals, the quality of the ground system (the equivalent loss resistance) plays a very important role in the bandwidth of the antenna. **Table 9-6** shows the calculated impedances and SWR values for short top-loaded verticals. The same equivalent ground resistance of 10 Ω used in Table 9-5 has a very drastic influence on the bandwidth of a very short vertical. Note the drastic drop in Q and the increase in bandwidth with the 10-Ω ground resistance.

Two factors definitely influence the SWR bandwidth of a vertical of a given length: the conductor diameter and the total loss resistance. We only want to increase the conductor diameter to increase the bandwidth where possible. If you want to use the loss resistance to increase the bandwidth, you might as well use a dummy load for an antenna. After all, a dummy load has a large SWR bandwidth and the worst possible radiating efficiency!

Table 9-5
Quarter-Wave Verticals on 80 Meters

Z_t, SWR_t and Q_t indicate the theoretical figures assuming zero ground loss. Z_g, SWR_g and Q_g values include an equivalent ground resistance of 10 Ω.

	Diameter Vertical	*2 mm (0.08")*	*40 mm (1.6")*	*250 mm (10")*
3.5 MHz	Z_t =	31.6 – *j* 31.4	31.4 – *j* 23.5	31.1 – *j* 16.7
	Z_g =	41.6 – *j* 35.9	41.4 – *j* 23.5	41.1 – *j* 16.7
	SWR_t =	2.8:1	2.0:1	1.7:1
	SWR_g =	2.2:1	1.7:1	1.5:1
3.65 MHz	Z_t =	35.9	35.9	35.9
	Z_g =	45.9	45.9	45.9
	SWR_t =	1:1	1:1	1:1
	SWR_g =	1:1	1:1	1:1
3.8 MHz	Z_t =	40.0 + *j* 35.5	40.9 + *j* 24.5	41.1 + *j* 16.6
	Z_g =	50.0 + *j* 35.5	40.9 + *j* 24.5	51.1 + *j* 16.6
	SWR_t =	2.5:1	1.9:1	1.6:1
	SWR_g =	2.1:1	1.7:1	1.4:1
All	Q_t =	12.1	8.1	5.6
	Q_g =	9.5	6.4	4.4

Table 9-6
Verticals with 40-mm OD for 80 Meters

Z_t, SWR_t and Q_t are the values for a 0-Ω ground resistance. Z_g, SWR_g and Q_g relate to an equivalent ground resistance of 10 Ω.

Frequency		*λ/8 Long (9.9 m) (28.4 ft)*	*3λ/16 Long (12.6 m) (41.3 ft)*
3.5 MHz	Z_t =	5.37 – *j* 340	9.3 – *j* 237
	Z_g =	15.37 – *j* 340	19.3 – *j* 237
	SWR_t =	15.7:1	6.0:1
	SWR_g =	3.6:1	2.7:1
3.65 MHz	Z_t =	5.9 – *j* 319	10.3 – *j* 217
	Z_g =	10.5 – *j* 319	20.3 – *j* 217
	SWR_t =	1:1	1:1
	SWR_g =	1:1	1:1
3.8 MHz	Z_t =	6.47 – *j* 299	11.4 – *j* 198
	Z_g =	16.47 – *j* 299	21.4 – *j* 198
	SWR_t =	12.3:1	4.9:1
	SWR_g =	3.3:1	2.4:1
All	Q_t =	42	23
	Q_g =	15	12

If you use a coil for loading a vertical (center or top loading), you can see that for a given antenna diameter, the bandwidth will decrease as the antenna is shortened and the missing part is partly or totally replaced by a loading coil. Then with more shortening, the bandwidth will begin to increase again as the influence of the equivalent resistive loss in the coil begins to affect the bandwidth of the antenna.

If you measure an unusually broad bandwidth for a given vertical design, you should suspect a poor-quality loading coil or some other lossy element in the system. (Or did you forget a ground system?)

3.6. Designing Short Loaded Verticals

In the previous editions of this book, I went through the mathematics needed to calculate various forms of loaded short verticals. Over the years, modeling programs have become increasingly popular and accurate, and more and more hams who want to build their own antenna use an antenna modeling program.

Therefore, for this edition I decided to simply use such a modeling program to calculate (model) a few examples of various forms of short loaded verticals. The examples worked out below all use 60 λ/4 radials over good ground. The estimated equivalent loss resistance is about 10 Ω.

3.6.1. Base Coil Loading

Assume a 24-meter high vertical with an effective diameter of 25 cm, which you can use as a 3λ/8 vertical on 80 meters. You can also resonate it on 160 meters using a base-mounted loading coil (**Fig 9-51**).

I plugged these data into *EZNEC* (*NEC-2* engine), and selected *Real/Mininec ground* as the ground type (average ground, dielectric constant ε = 13, conductivity = 0.005 S/m). We will soon find out that the quality of the ground system (its equivalent loss resistance, see Table 9-1) determines the antenna efficiency much more than the loading device.

Without any loading the vertical resonates around 3 MHz. Adding a coil with an impedance of 182 Ω at the bottom, the antenna now resonates on 1.83 MHz.

Further data are:

R_{rad} = 9.52 Ω
X_L (base loading) = 182 Ω works out to 15.8 μH
(at 1.83 MHz)

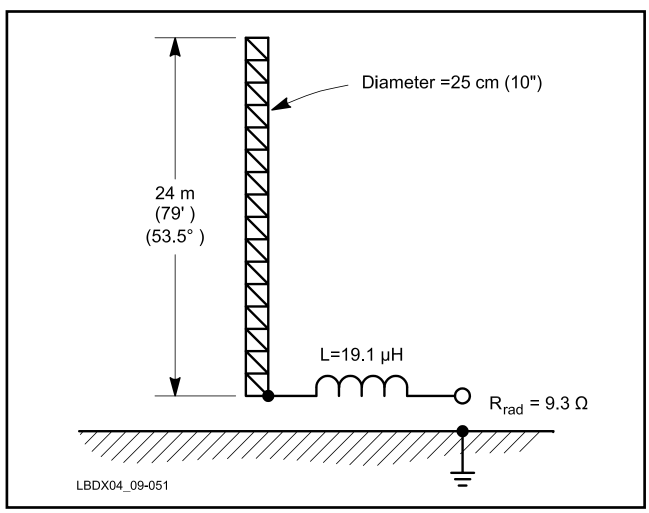


Fig 9-51 — Base-loaded tower for 160 meters. See text for details on how to calculate the radiation resistance as well as the value of the loading coil. The loss resistance is effectively in series with the radiation resistance. With 60 λ/8 radials over good ground, the feed-point impedance will be approximately 20 Ω and the radiation efficiency about 50%.

If Q = 300 then R_{loss}-coil = 182/300 = 0.6 Ω
Z_{feed} = 20.25 Ω
Gain = –1.73 dBi

Radiation efficiency = 9.5/20.2 = 49%. Only 5% of the losses are coil losses — 95% are ground losses. If we had a perfect ground (saltwater) the efficiency would be 9.5/10.1 = 94 %

3.6.2. Capacitance-Hat Loading

As explained before (Section 3.1.2.1.) capacitance top (end) loading is the most efficient way to load a short vertical (or dipole). Using *EZNEC*, I modeled a capacitance hat in the form of a cross (4 spokes) that would be large enough to resonate the 24-meter vertical on 1.83 MHz.

3.6.2.1. Start from Maximum Capacitance Hat Dimension

In Section 3 we said that to feed a short vertical with low losses using a coaxial feed line, we first get rid of the reactive part and then increase the real part of the feed impedance up to 50 Ω. So far, in all the examples we have given, we have really only bothered with loading the short vertical (making it resonant). What if we can do the loading and the matching simultaneously?

Assume we can only put up 12.5 meters of small section tower, but can construct a top hat with four 6-meter-long tubes (say, a couple of reflector elements from a full size Yagi for 20 meters, placed at right angles). See **Fig 9-52**.

EZNEC says that on 3.65 MHz (halfway between the phone and the CW DX bands) the antenna has an impedance of 43.6 + *j* 47 Ω. If we simply use a series capacitor (X_C = –*j* 55 Ω, C = 793 pF on 3.65 MHz), this becomes a very nice broad antenna with an SWR on 1.5:1 on 3.5 MHz, 1.15:1 on 3.65 MHz and 1.5:1 on 3.8 MHz.

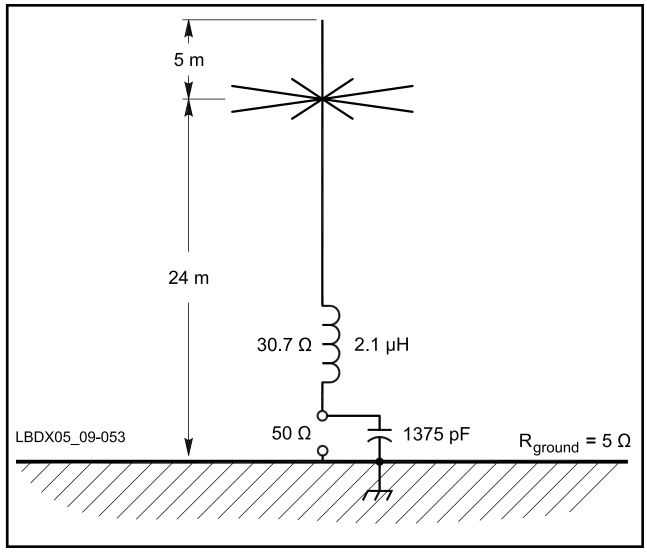


Fig 9-53 — A 24-meter tower with an 8-spoke capacitance hat (spokes 6 meters long) results in a high performance 160-meter vertical with at least 300 kHz 2:1 SWR bandwidth.

A similar capacitance hat with eight 6-meter-long tubes suffices to bring a 24-meter tall tower to resonance on 1.85 MHz. Note there is a 5-meter-long mast extending above the capacitance hat. This mast can be used to connect truss wires (if any) that help support the loading tubes.

The same tower height (24 meters) that yields R_{rad} of 9.5 Ω when base loaded (see Fig 9-51) now shows a radiation resistance of 25 Ω with top loading.

The performance data in **Table 9-7** were calculated assuming an equivalent ground loss of 10 Ω (obtained with 50 λ/4 radials over average ground).

If we have a better ground system with only 5 Ω equivalent losses, we will need a small L-network to match the 30.7 Ω impedance (at 1.835 MHz) to 50 Ω (see **Fig 9-53**). The 2:1 SWR bandwidth is 220 kHz, the 1.5:1 bandwidth 100 kHz.

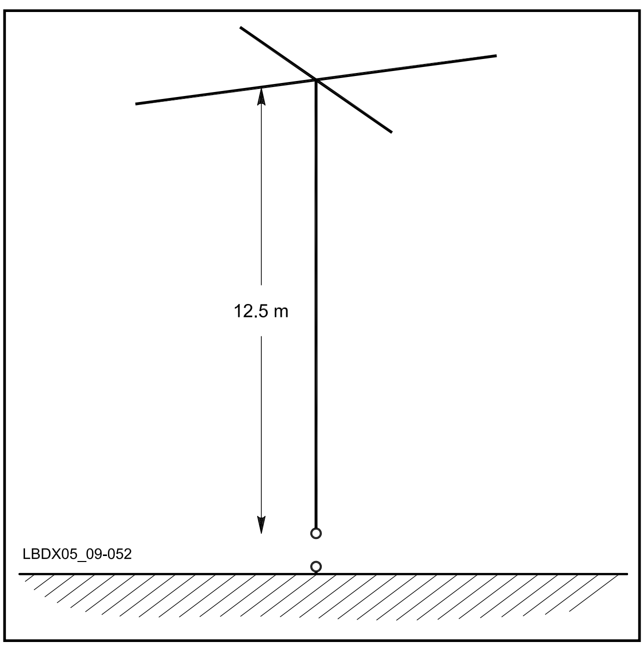


Fig 9-52 — A 12.5-meter tall vertical loaded with two top hat crossed elements (having the size of a 20-meter Yagi reflector) makes a perfect broadband 80-meter antenna covering from 3.5 to 3.8 MHz with an SWR of less than 1.5:1 (assuming 10 Ω equivalent ground losses), requiring no other components than a simple series capacitor to match the antenna to a 50-Ω feed line.

3.6.2.2. Using Wires as Top Loading Devices

Very often wire loading elements are used, and in most cases they slope down from the top of the vertical (sometimes they are part of the vertical guying system). While from a performance point of view horizontal capacitive loading elements are best, we can live with "slightly drooping" loading wires.

What is "slightly drooping"? If we use top loading wires that droop steeply, we build a cage (coaxial structure) around the vertical which will reduce the R_{rad}. A perfect non leaking case around it would yield a R_{rad} of 0 Ω. **Fig 9-54** shows such a vertical (12.5 meters high) with eight sloping loading wires coming down to 2 meters above ground. The R_{rad} is 2.7 Ω to be compared with 9 Ω radiation resistance in case of perfect flat top capacitance hat.

I modeled two 160-meter short verticals; one measures 12.5 meters (33°) the other one 20 meters (53°). As a reference I modeled these verticals using four perfectly horizontal top loading wires. Then I gradually drooped the loading wires, changing their length to keep the antenna resonant at 1.83 MHz.

Table 9-7
Impedance of 160 Meter Top Loaded Vertical

Frequency	1.75	1.8	1.83	1.85	1.9	1.95
Z (Ω)	32.3 - j20.6	34.2 - j8.7	35.5 - j1.5	36.4 + j3.2	38.7 + j15	41.2 + j27
SWR (50 Ω)	1.9	1.5	1.4	1.4	1.5	1.85

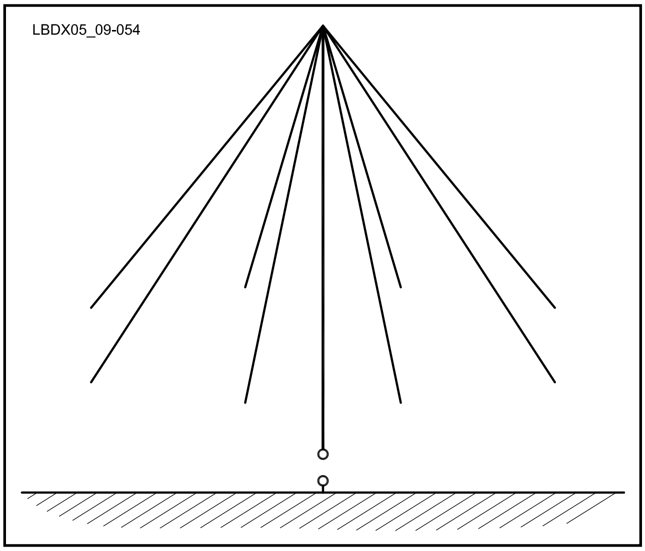


Fig 9-54 — If the drooping top loading wires encircle the antenna and come almost as low as the ground, the R_{rad} becomes very small. It is as if we "encased" the antenna in a shield.

Table 9-8 shows the results. The R_{rad} drops to less than 50% of the original (flat top) value, and the gain, assuming an equivalent ground loss resistance of 10 Ω (obtained, for example, with 50 λ/4 radials over average ground) drops by almost 3 dB for the shorter vertical and by 1.6 dB for the taller one. By the way, with a lossless ground (such as saltwater), there is no reduction in efficiency.

From the table we learn that the higher the R_{rad}, the higher the gain and also the higher the SWR bandwidth. The bandwidth was calculated including the 10 Ω equivalent ground loss, and using the normalized impedance values (50 Ω at resonance). Note that for both examples the bandwidth is rather similar; this is of course due to the low R_{rad} and the relatively high content of ground loss resistance.

Conclusion: Keep the top loading wires as horizontal as possible, and certainly do not slope them lower than half the height of the vertical. Also, don't forget that the smaller the value of R_{rad}, the more important it is to have a good radial system. The most important part of a shortened vertical antenna is "in" the ground.

3.6.2.3. T-wire Loading (T Antenna)

If the vertical is attached at the center of the top-loading wire, the horizontal (high-angle) radiation from this top wire will be effectively canceled in the far field. **Fig 9-55** shows a typical configuration of a T antenna. Two existing supports, such as trees, are used to hold the flattop wire.

Fig 9-56 shows a design chart derived using the *EZNEC* modeling program. The dimensions can easily be extrapolated to other design frequencies. In practice, the T-shaped loading wires will often be slightly upward sloping loading wires. In this case the radiation resistance will be slightly higher (the effective lengths of the vertical is greater due do the vertical component from the upward sloping loading wires).

3.6.3. Center Loading

Center loading is rarely used on the low bands. It is not easy to insert a loading coil halfway up in a vertical structure (tower, mast), and you are confronted with an inherently lossy element, the loading coil.

3.6.4. Combined Loading Methods

Often top loading is complemented with some degree of base loading. That might be needed because it is impossible to achieve sufficient top loading. Another reason is that various small amounts of base loading can be used to "tune" the vertical for different parts of the band. (See Section 3.1.4 and Fig 9-47.)

It you want to design your own vertical, start modeling it and see where you can get with top loading. If you cannot completely tune out the capacitive reactance with the top hat, you will need to add a small coil to do that.

Using a center loading coil to do that is not meaningful (nor practical in most cases), because once you are using a sizeable amount of top (capacitive) loading, the current in the vertical is almost constant. It will

Table 9-8
Effects of Top Loading Wire Tip Height (1.83 MHz)

Height (m)	*Tip height (m)*	*R_{rad} Ω*	*Relative Gain for 10 Ω ground loss*	*2:1 SWR BW (10 Ω ground loss)*
12.5	4	3.7	–2.3 dB	90 kHz
12.5	6	4.7	–1.6 dB	100 kHz
12.5	8	5.8	–1.0 dB	110 kHz
12.5	10	7.1	–0.5 dB	120 kHz
12.5	12.5	8.7	0	140 kHz
20	5	8.3	–1.6 dB	100 kHz
20	7.5	10.1	–1.2 dB	100 kHz
20	10	11.9	–0.8 dB	120 kHz
20	12.5	13.7	–0.6 dB	125 kHz
20	15	15.5	–0.4 dB	130 kHz
20	17.5	17.5	–0.2 dB	140 kHz
20	20	19.3	0	150 kHz

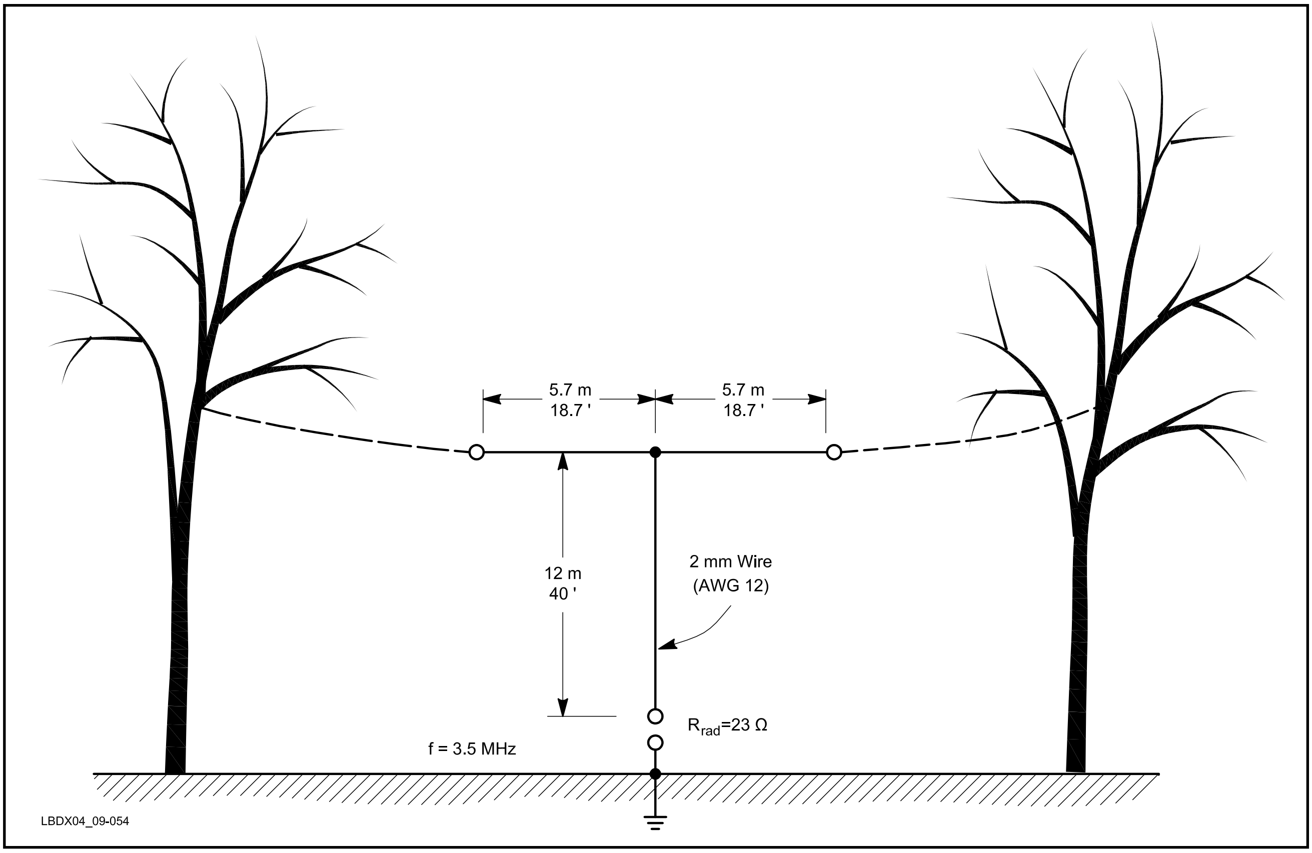


Fig 9-55 — Typical setup of a current-fed T antenna for the low bands. Good-quality insulators should be used at both ends of the horizontal wire, as high voltages are present.

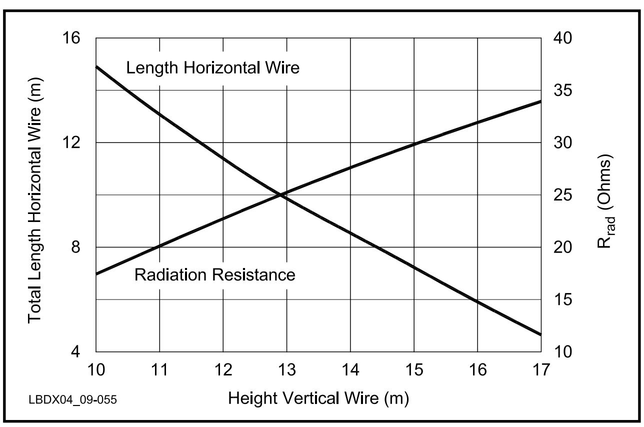


Fig 9-56 — Design chart for a wire-type λ/4 current-fed T antenna made of 2-mm OD wire (AWG #12) for a design frequency of 3.5 MHz. For 160 meters the dimension should be multiplied by a factor of 1.9.

make very little difference where in the vertical section you put the loading coil.

Example

A vertical, 24 meters long (25 cm diameter), using two sloping top hat wires, each 15.3 meters long with tips 16 meters above the ground. Equivalent ground resistance: 10 Ω. We modeled this antenna with a coil at the bottom, halfway up and at the top of the vertical section. Using coils with Q = 300, the results are:

Base loading: X_L = 15.2 Ω, Z = 29.45 Ω
Center loading: X_L = 19 Ω, Z = 20.41 Ω
Top loading: X_L = 30 Ω, Z = 30.45 Ω

This example makes clear that it does not pay off to put the coil in the center or even at the top of the vertical section; all three cases yield approximately the same feed point impedance, which means the same equivalent loss resistance and efficiency.

3.7. Loading Coils

How much power is dissipated in a loading coil? Let's take the base loading example given above. First, we should model the antenna again and look at the current along the vertical section. *EZNEC* always calculates with a feed point current of 1 A. For 1500 W (and a 10 Ω equivalent ground loss resistance, R_{ground}) the current is:

$$I_{base} = \sqrt{\frac{1500}{29.5}} = 7.13 \text{ A}$$

The base loading coil has an impedance of 15.2 Ω, for a Q of 300 the loss resistance is 15.2/300 = 0.05 Ω. $P = I^2R = 7.13^2 \times 0.05 = 2.54$ W. This is a low value (0.007 dB loss)

Let's do the same calculation for the top loaded case:

EZNEC tells us that the current at the top of the vertical section is 71.9% of the value at the bottom.

$$I_{top} = \sqrt{\frac{1500}{30.45}} = 7.02 \text{ A}$$

The loss resistance for Q = 300: R = 30/300 = 0.1 Ω

The dissipated power in the coil is P = $(7.02 \times 0.719)^2 \times 0.1 = 2.55$ W

Note that this is exactly the same value as calculated for the base loading coil. This proves once more that for a short vertical antenna with lots of top loading, where the current in the verticals section is fairly constant, the losses are much the same whether you put the coil at the bottom, halfway or at the top of the antenna.

3.7.1. A Large Loading Coil with Little Capacitive Top Loading

Let's take a really short 160-meter vertical, measuring only 8 meters tall, using a small capacitance top hat consisting of four 5-meter-long spokes. Let us see what loading coil we need at the top (just under the hat) and at the bottom, and what the power losses are.

Top loading

Modeling with *EZNEC* tells us we need a coil with $X_L = 495\ \Omega$ to resonate the antenna on 1.83 MHz (L = 43 µH). That is a pretty big coil — diameter 75 mm; length 330 mm; wire diameter 3 mm; winding pitch, turns: 55. Let's assume we make a coil with a poor Q (150). The coil resistive losses are 495 /150 = 3.3 Ω.

$$I_{top} = \sqrt{\frac{1500}{17.66}} = 9.22\ \text{A}$$

EZNEC calculates the current at the top of the vertical section as 108% of the value at the bottom. The dissipated power in the coil is P = $(9.22 \times 1.08)^2 \times 3.3 = 327$ W. That is a lot of power, and chances are the coil will be destroyed.

Eff = 4.4/17.7 = 25%, which means that 373 W will be radiated, 327 W heats the coil and 800 W heats the soil!

Base loading

For base loading we require $X_L = 343\ \Omega$ (L = 30 µH). A similar coil with a relatively poor Q of 150 means an equivalent coil series resistance of 2.3 Ω.

$$I_{base} = \sqrt{\frac{1500}{15.37}} = 9.87\ \text{A}$$

The dissipated power in the coil is P = $9.87^2 \times 2.3 = 224$ W. It is a little better than in the top loading case but chances are still there that the coil will be destroyed.

On the other hand the total efficiency is lower (because of slightly lower R_{rad}): Eff = 2.6/15.4 = 17%, which means that 253 W will be radiated, 224 W heats the coil and 1023 W heats the soil!

Top loading with better ground system (R_{ground} equivalent = 5 Ω)

In this case the total impedance drops by 5 Ω:

$$I_{top} = \sqrt{\frac{1500}{12.66}} = 10.9\ \text{A}$$

The dissipated power in the coil is P = $(10.9 \times 1.08)^2 \times 3.3 = 457$ W.

It is clear that, the better the ground system, the more power will be dissipated in the loading coil.

3.7.2. Making or Buying High-Q Loading Coils

C. J. Michaels, W7XC (SK), investigated the construction and the behavior of loading coils for 160 meters (Ref 797). In the above examples we assumed a (poor) Q factor of 150. (See also Ref 694 and 695.) How can we build loading coils having the highest possible unloaded Q? Michaels came to the following conclusions:

- For coils with air dielectric, the L/D (length/diameter) ratio should not exceed 2:1.
- For coils wound on a coil form, this L/D ratio should be 1:1.
- Long, small-diameter coils are not good.
- The highest Q that can be achieved for a 150-µH loading coil for 160 meters is approximately 800. This can be achieved with an air-wound coil (15 cm long by 15 cm diameter), using 35 turns of AWG #7 (3.7-mm OD) wire, or with an air-wound coil (30 cm long by 15 cm diameter, wound with 55 turns of AWG #4 (5.1-mm OD) wire.
- Coil diameters of 10 cm wound with AWG #10 (2.5 mm OD) to #14 (1.6 mm OD) wire can yield Q factors of 600, while coil diameters of 5 cm wound with BSWG #20 (0.9 mm OD) to #22 (0.7 mm OD) will not yield Q factors higher than approximately 250. These smaller wire gauges should not be used for high-power applications.

You can use some common sense and simple test methods for selecting an acceptable plastic coil-form material:

- High-temperature strength: Boil a sample in water for ½ hour, and check its rigidity immediately after boiling while still hot.
- Check the loss of the material by inserting a piece inside an air-wound coil, for which the Q is being measured. There should be little or no change in Q.
- Check water absorption of the material: Soak the sample for 24 hours in water and repeat the above test. There should be no change in Q.
- Dissipation factor: Put a sample of the material in a microwave oven, together with a cup of water to load the oven. Run the oven until the water boils. The sample should not get appreciably warm.

Another guru in quality coil construction, W8JI, says that a high-Q coil for use as an antenna loading coil can usually be made using a turn-to-turn spacing that equals the conductor diameter, and an L/D ratio between 2 and 4. This is quite different from more common applications of coils such as in amplifier networks where optimum Q can usually be obtained with L/D ratios close to 1. Always use air-wound coils and stay away from coil forms or dielectric coatings. For more details visit **www.w8ji.com/mobile_and_loaded_antenna.htm** and **www.w8ji.com/loading_inductors.htm**.

Commercial high-quality inductors are available from Barker & Williamson (**www.bwantennas.com/coils/coilcat.htm**). The Airdux TL stock coils can be used to make high quality loading coils.

3.8. Comparing Different Loading Methods

Capacitance Top Loading

We have learned that only top loading is okay, unless you really need just a little bit of loading. In that case a small coil at the base of the vertical (base loading) will do an equally good job.

Table 9-9
Comparison of 15-Meter Tall Verticals for 160 Meters

Length of vertical = 15 m (= 35°); diameter = 25 cm effective; F = 1.83 MHz

	----Ground R = 15 Ω ----		*---- Ground R = 10 Ω ----*		*---- Ground R = 5 Ω ----*	
	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*
¼ λ full size (39 m)	36 / 51	71 / –1.5	36 / 46	78 / –1.1	36 / 41	88 / –0.6
Flat top hat (T ant)	12.4 / 27.4	45 / –3.4	12.4 / 22.4	55 / –2.6	12.4 / 17.4	72 / –1.5
2 sloping wires[1]	8.3 / 23.3	36 / –4.5	8.3 / 18.3	45 / –3.4	8.3 / 13.3	65 / –1.9
2 sloping wires[2]	5.6 / 20.6	27 / –5.6	5.6 / 15.6	36 / –4.4	5.6 / 10.6	53 / –2.8

[1]top loading, sloping T, end loading wires 8 m above ground
[2]top loading, sloping T, end loading wires 2 m above ground

Table 9-10
Comparison of 20-Meter Tall Verticals for 160 Meters

Length of vertical = 20 m (= 46°); diameter = 25 cm effective; F = 1.83 MHz

	----Ground R = 15 Ω ----		*---- Ground R = 10 Ω ----*		*---- Ground R = 5 Ω ----*	
	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*
¼ λ full size (39 m)	36 / 51	71 / –1.5	36 / 46	78 / –1.1	36 / 41	88 / –0.6
Flat top hat (T ant)	19.7 / 34.7	57 / –2.5	19.7 / 29.7	66 / –1.8	19.7 / 24.7	80 / –1.0
2 sloping wires[1]	13.2 / 24.9	47 / –3.3	13.2 / 23.2	57 / –2.4	13.2 / 18.2	73 / –1.4
2 sloping wires[2]	9.9 / 24.9	37 / –4.3	9.9 / 19.9	50 / –3.0	9.9 / 14.9	66 / –1.8

[1]top loading, sloping T, end loading wires 8 m above ground
[2]top loading, sloping T, end loading wires 2 m above ground

Table 9-11
Comparison of 25-Meter Tall Verticals for 160 Meters

Length of vertical = 25 m (= 58°); diameter = 25 cm effective; F = 1.83 MHz

	----Ground R = 15 Ω ----		*---- Ground R = 10 Ω ----*		*---- Ground R = 5 Ω ----*	
	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*
¼ λ full size (39 m)	36 / 51	71 / –1.5	36 / 46	78 / –1.1	36 / 41	88 / –0.6
Flat top hat (T ant)	26.5 / 41.5	64 / –1.9	26.5 / 36.5	73 / –1.4	26.5 / 31.5	85 / –0.7
2 sloping wires[1]	22.4 / 37.4	60 / –2.3	22.4 / 32.4	69 / –1.6	22.5 / 27.4	82 / –0.9
2 sloping wires[2]	19.4 / 34.4	56 / –2.5	19.4 / 29.4	66 / –1.8	19.4 / 24.4	80 / –1.0

[1]top loading, sloping T, end loading wires 8 m above ground
[2]top loading, sloping T, end loading wires 2 m above ground

Table 9-12
Comparison of 30-Meter Tall Verticals for 160 Meters

Length of vertical = 30 m (= 69°); diameter = 25 cm effective; F = 1.83 MHz

	----Ground R = 15 Ω ----		*---- Ground R = 10 Ω ----*		*---- Ground R = 5 Ω ----*	
	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*
¼ λ full size (39 m)	36 / 51	71 / –1.5	36 / 46	78 / –1.1	36 / 41	88 / –0.6
Flat top hat (T ant)	31.8 / 46.8	68 / –1.7	31.8 / 41.8	76 / –1.2	31.8 / 36.8	87 / –0.6
2 sloping wires[1]	29.8 / 44.8	66 / –1.8	29.8 / 39.8	75 / –1.3	29.8 / 34.8	86 / –0.7
2 sloping wires[2]	28.0 / 43.0	65 / –1.9	28.0 / 38.0	74 / –1.3	28.0 / 33.0	85 / –0.7

[1]top loading, sloping T, end loading wires 8 m above ground
[2]top loading, sloping T, end loading wires 2 m above ground

Table 9-13
Characteristics of 25-Meter Tall Verticals for 160 Meters with Base Loading
Length of vertical = 25 m; diameter = 25 cm effective; F = 1.83 MHz; XL = 169 Ω; L = 14.7 µH

	----Ground R = 15 Ω ----		*---- Ground R = 10 Ω ----*		*---- Ground R = 5 Ω ----*	
	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*
¼ λ full size (39 m)	36 / 51	71 / –1.5	36 / 46	78 / –1.1	36 / 41	88 / –0.6
Q coil = 100	10.5 / 27.2	39 / –4.1	10.5 / 22.2	47 / –3.2	10.5 / 17.2	61 / –2.1
Q coil = 200	10.5 / 26.4	40 / –3.9	10.5 / 21.4	49 / –3.1	10.5 / 16.4	64 / –1.9
Q coil = 400	10.5 / 25.9	41 / –3.9	10.5 / 20.9	50 / –3.0	10.5 / 15.9	66 / –1.8

Table 9-14
Characteristics of 30-Meter Tall Verticals for 160 Meters with Base Loading
Length of vertical = 30 m; diameter = 25 cm effective; F = 1.83 MHz; XL = 106 Ω; L = 9.2 µH

	----Ground R = 15 Ω ----		*---- Ground R = 10 Ω ----*		*---- Ground R = 5 Ω ----*	
	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*	R_{rad} / R_{feed}	*Eff (%) / –dB*
¼ λ full size (39 m)	36 / 51	71 / –1.5	36 / 46	78 / –1.1	36 / 41	88 / –0.6
Q coil = 200	16.8 / 32.3	52 / –2.8	16.8 / 27.3	62 / –2.1	16.8 / 22.3	75 / –1.2
Q coil = 400	16.8 / 32.1	52 / –2.8	16.8 / 27.1	62 / –2.1	16.8 / 22.1	76 / –1.2
Q coil = 600	16.8 / 31.8	53 / –2.8	16.8 / 26.8	63 / –2.0	16.8 / 21.8	77 / –1.1

Let's compare "short" verticals, ranging from 15 to 30 meters in length (on 160 meters), using flat top T loading and sloping T loading wires. I used *EZNEC* (*NEC-2* engine) and to simulate the loss of the radial system just plugged in various values of equivalent radial system loss (15, 10 and 5 Ω). The results of the analysis are reported in **Table 9-9** through **Table 9-12**. **Table 9-13** and **Table 9-14** show characteristics of 160 meter verticals with a little bit of base loading.

Al Christman, K3LC, did a similar study (Ref 7809). He used an excellent radial system consisting of 120 quarter-wave radials (over average ground). I'd estimate the equivalent loss resistance for that system to be around 5 Ω. The radiation resistances he reports are the same as shown above. However, he claims that the peak gain of all these verticals, from short (35°) to long (89°) all show the same. (They range from 0.65 to 0.73, which I call "the same.") That seems unlikely and is not what I would expect nor what I have found.

Given the wide range of radiation resistances (as low as 5 Ω to as high as 35 Ω), and using an identical ground system (120 λ/4 radials), and the constant gain he reports, it would mean that the equivalent R_{ground} of the radial systems varies greatly with the length of the vertical radiator:

To make it simple, let's say that the gain is constant. If that is true it means that the equivalent resistive loss of the radial system changes with the physical length of the antenna, at the same rate as the radiation resistance. This means that R_{ground} of the radial system is seven times higher with a full-size quarter wave than with a 35° long vertical (the same ratio as with the radiation resistances). Extrapolating a little further, it would mean that we can use a very short loaded vertical with a radial system using a great number of very short radials. If this were the case, why don't we all use such short loaded verticals (and how about the broadcast stations)?

Conclusions

From all of this it is clear that base loading is *not* the preferred way of loading a short vertical even if the short vertical is "relatively" long (25 or 30 meters vs 39 meters for a quarter-wave). Let's compare the 25 meter vertical over a ground with 10 Ω loss. If we load with gently sloping T wires, the efficiency is 69% (–1.6 dB). If we use base loading even with a coil with Q = 400, the efficiency is 50% (–3 dB), which makes 1.4 dB difference — on 160 that can be a lot!

The reason for this is *not* coil losses (changing from Q = 100 to Q = 400 in Table 9-13 makes 0.2 dB in difference). The difference is due to the fact that R_{rad} with base loading is much lower.

3.9. "Over"-Loading (Short) Verticals

The radiation resistance of a vertical is a measure of the area under the current distribution curve along the vertical (see Section 1.2). With a very short vertical and a large degree of top loading the current remains almost constant along the vertical radiator.

With not-so-short vertical radiators we can sometimes increase the area under the current distribution curve by making the antenna electrically longer than a quarter wave, in such a way that the current maximum is about halfway up the vertical radiator.

This will raise the R_{rad}, which may also make it "easier" to match the antenna to the 50-Ω feed line.

Example 1

We have a 24-meter high tower that we want to turn into a 160-meter top loaded vertical. The equivalent ground loss is estimated at 10 Ω.

In **Fig 9-57A** we see the situation for capacitance top

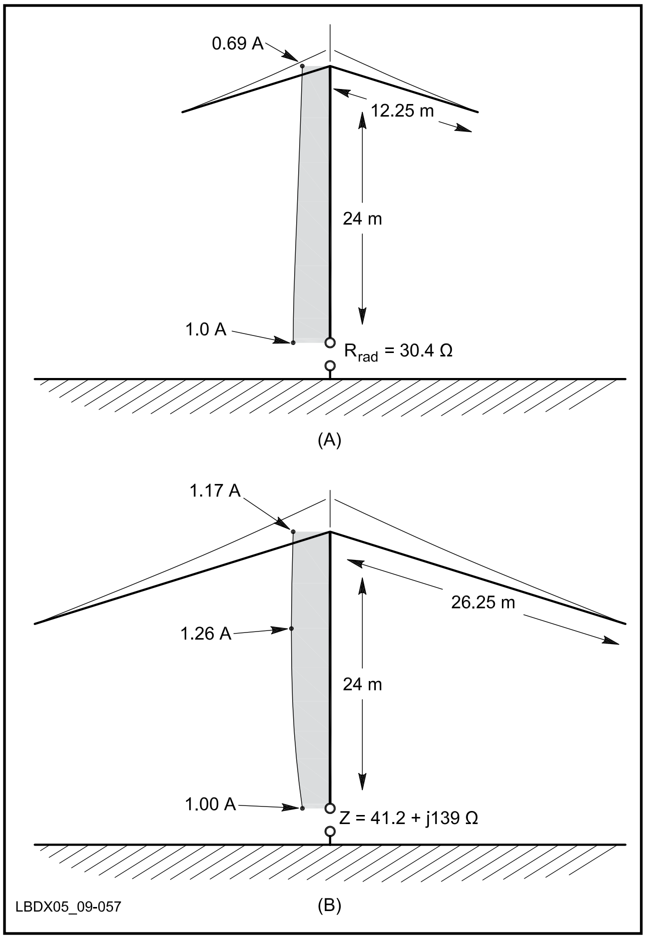


Fig 9-57 — By applying extra top loading we can increase the signal by 0.3 dB and achieve a super simple matching system.

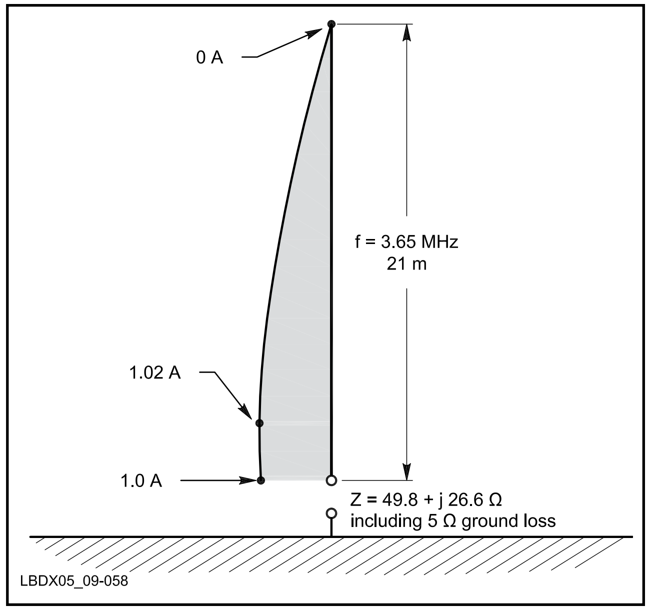


Fig 9-58 — Slightly lengthened 80-meter vertical that requires only a series capacitor to match to 50-Ω coax and covers 3.5 to 3.8 MHz with less than 1.5:1 SWR.

loading (two loading wires, 15.25 meters long, tips 16 meters above ground) that makes the tower resonant on 1.83 MHz. In this case R_{rad} is 30.4 Ω.

If we add a little more top loading (Fig 9-57B) we can obtain more area under the current distribution line along the vertical. More area means higher R_{rad}. The loading wires are now 26.25 meters long, tips 16 meters above ground). On 1.83 MHz the feed impedance is now 41.2 + *j* 139 Ω assuming zero ground loss.

If we bring the ground loss resistance (10 Ω) into the picture, the feed impedance of case (A) is 40.4 Ω, for case B it is 51.2 + *j* 139 Ω. Connecting a capacitor with $X_C = -139$ Ω (at 1.83 MHz) in series with the feed point now matches the feed line impedance, Z_{feed} becomes 51.2 Ω, a perfect match for the 50 Ω feed line. Using *EZNEC* we can easily calculate the bandwidth of this antenna which is approximately 170 kHz (2:1 SWR). The efficiency of this antenna is 41.2/51.2 = 80.5% or –0.9 dB. For case A this is 75% or –1.2 dB. We win 0.3 dB in signal strength and at the same time have a super simple matching system, just a series capacitor.

Example 2

Let's design an 80-meter vertical (**Fig 9-58**) that is a little longer than a quarter-wave in the center of the band, so that the resistive part of the feed impedance at that frequency is about 50 Ω. Assume a pretty good ground system with an equivalent resistive loss of 5 Ω. *EZNEC* tells us that a vertical measuring 21 meters (250 mm effective diameter) exhibits an impedance of 44.75 + *j* 26.5 Ω. Adding the equivalent ground loss resistance we end up with 49.75 + *j* 26.6 Ω. We can tune out the positive reactance with a series capacitor with a value of approximately 1634 pF.

If we model this antenna plus (negative) loading system on 3.5 MHz and 3.8 MHz we find that the SWR on these band edges is less than 1.5:1! This makes a very simple broadband 80-meter antenna with the simplest possible (negative) loading plus matching system (a series capacitor).

4. TALL VERTICALS

In this section we'll examine verticals that are substantially longer than λ/4, especially their behavior over different types of ground. Is a very low elevation angle computed over ideal ground ever realized in practice?

First of all, you need to ask whether you really need very low elevation angles on the low bands. A very low incident angle grazes the ionosphere for a long distance, increasing loss. More hops with less loss from a sharper angle can actually decrease propagation loss. We saw in Chapter 1 that relatively high launch angles are actually a prerequisite to allow a "duct" to work on 160 meters, typically at sunrise. On 160 meters, we can state that the antenna with the most gain at the lowest elevation angle under almost all circumstances will produce the strongest signal.

In this section I will dispel a myth that voltage-fed antennas do not require an elaborate ground system. In fact, long verticals require an even better radial system and an even better ground quality in the Fresnel zone to achieve their low-angle and gain potential compared to a λ/4 vertical.

In earlier sections of this chapter, I dealt with short verticals in detail, mostly for 160 meters. On higher frequencies, electrically taller verticals are quite feasible. A full-size λ/4

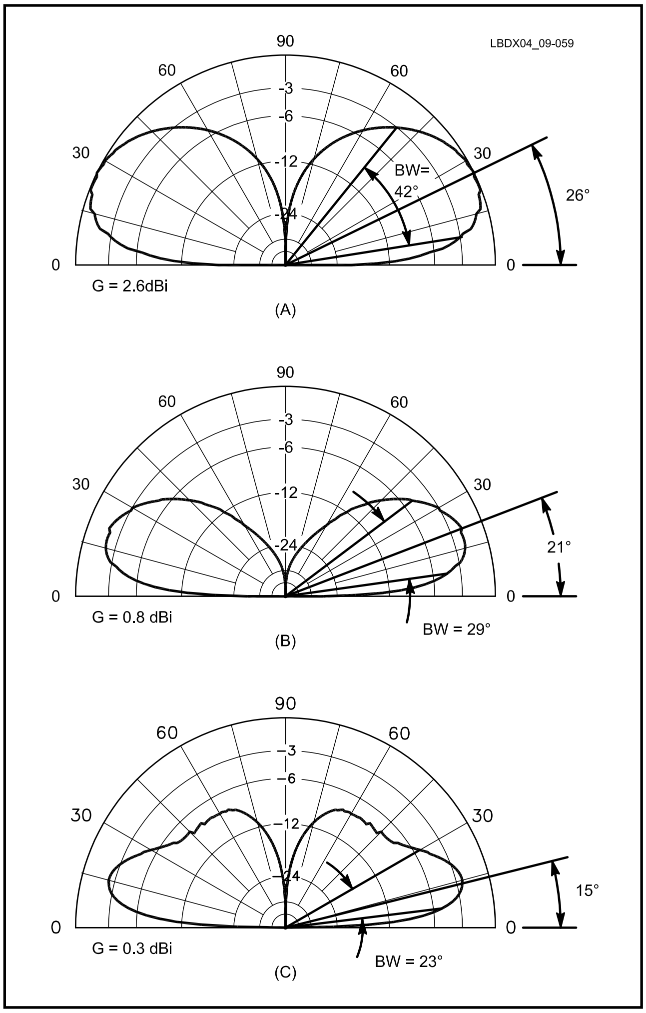


Fig 9-59 — Vertical radiation patterns of different-length verticals over average ground, using 60 λ/4 radials. The 0-dB reference for all patterns is 2.6 dBi. At A, λ/4 vertical. At B, λ/2 and at C, 5λ/8.

radiator on 80 meters is approximately 19.5 meters in height. Long verticals are considered to be λ/2 to 5λ/8 in length. Verticals that are slightly longer than a quarter-wave (up to 0.35 λ) do not fall in the *long vertical* category.

4.1. Vertical Radiation Angle

Fig 9-59 shows the vertical radiation patterns of two long verticals of different lengths. These are analyzed over an identical ground system consisting of average earth with 60 λ/4 radials. A λ/4 vertical is included for comparison.

Note that going from a λ/4 vertical to a λ/2 vertical drops the maximum-elevation angle from 26° to 21°. More important, however, is that the 3-dB vertical beamwidth drops from 42° to 29°. Going to a 5λ/8 vertical drops the elevation angle to 15° with a 3-dB beamwidth of only 23°. But notice the high-angle lobe showing up with the 5λ/8 vertical. If we make the vertical still longer, the low-angle lobe will disappear and be replaced by a higher-angle lobe. A 3λ/4 vertical has a radiation angle of 45°.

Whatever the quality of the ground, the 5λ/8 vertical will always produce a lower angle of radiation and also a narrower vertical beamwidth. The story gets more complicated, though, when you compare the efficiency of the antennas.

4.2. Gain of a Tall Vertical

I have modeled both a λ/4 and a 5λ/8 vertical over different types of ground, in each case using a realistic number of

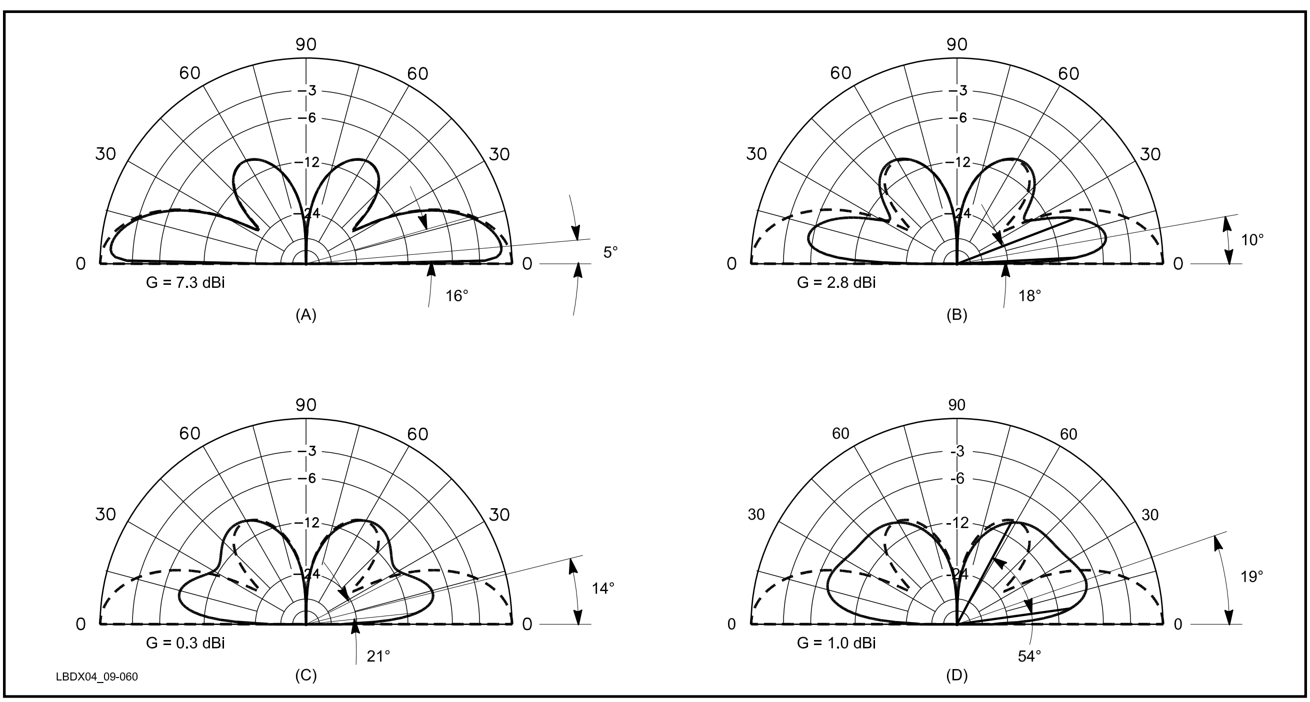


Fig 9-60 — Vertical radiation pattern of the 5λ/8 vertical over different types of ground. In all cases, 60 λ/4 radials were used. The theoretical perfect-ground pattern is shown in each case as a reference (broken line, with a gain of 8.1 dBi). Compare with the patterns and gains of the λ/4 vertical, modeled under identical circumstances (Fig 9-5). At A, over saltwater. At B, over very good ground. At C, over average ground. At D, over very poor ground.

60 λ/4 radials. Fig 9-5 earlier in this chapter shows the patterns and the gains in dBi for the quarter-wave vertical, and **Fig 9-60** shows the results for the 5λ/8 antenna.

Over perfect ground, the 5λ/8 vertical has 3.0 dB more gain than the λ/4 vertical at a 0° elevation angle. Note the very narrow lobe width and the minor high-angle lobe (broken-line patterns in Fig 9-60).

Over saltwater the 5λ/8 has lost 0.8 dB of its gain already; the λ/4 has lost only 0.4 dB. The 5λ/8 vertical has an extremely low elevation angle of 5° and a vertical beamwidth of only 17°. The λ/4 has an 8° take-off angle, but a 40° vertical beamwidth.

Over very good ground, the 5λ/8 vertical has now lost 5.0 dB; the λ/4 has lost only 1.9 dB. The actual gain of the λ/4 in other words equals the gain of the 5λ/8! Note also that the high-angle lobe of the 5λ/8 becomes more predominant as the quality of the ground decreases.

Over average ground the situation becomes really poor for the 5λ/8 vertical. The gain has dropped 7.3 dB, and the secondary high-angle lobe is only 4 dB down from the low-angle lobe. The λ/4 vertical has lost 2.6 dB versus ideal ground, and now shows 2.0 dB more gain than the 5λ/8 vertical!

Over very poor ground the 5λ/8 vertical has lost 6.6 dB from the perfect-ground situation, while the λ/4 vertical has lost only 3.0 dB. Note that the 5λ/8 vertical seems to pick up some gain compared to the situation over average ground. From Fig 9-60 you can see this is because the radiation at lower angles is now attenuated so much that the radiation from the high-angle lobe at 60° becomes dominant. Note also that the level of the high-angle lobe hardly changes from the perfect-ground situation to the situation over very poor ground. This is because the reflection for this very high angle takes place right under the antenna, where the ground quality has been improved by the 60 λ/4 radials.

This must come as a surprise to most. How can we explain this? An antenna that intrinsically produces a very low angle (at least in the perfect-ground model) relies on reflection at great distances from the antenna to produce its low-angle radiation. At these distances, radials of limited length do not play any role in improving the ground. With poor ground, a great deal of the power that is sent out at a very low angle to the ground-reflection point is being absorbed in the ground, refracted, scattered or absorbed on "obstacles" rather than being reflected (see also Section 1.1.2). For Fresnel-zone reflections the long vertical requires a better ground than the λ/4 vertical to realize its full potential as a low-angle radiator.

4.3. The Radial System for a Half-Wave Vertical

Here comes another surprise. A terrible misconception about voltage-fed verticals is that they do not require either a good ground or an extensive radial system.

4.3.1. The Near Field

If you measure the current going into the ground at the base of a λ/2 vertical, the current will be very low (theoretically zero). With λ/4 and shorter verticals, the current in the radials increases in value as you get closer to the base of the vertical. That's why, for a given amount of radial wire, it is better to use many short radials than just a few long ones.

With voltage-fed antennas, however, the earth current will increase as you move away from the vertical. Brown (Ref 7997) calculated that the highest current density exists at approximately 0.35 λ from the base of the voltage-fed λ/2 vertical. Therefore it is even more important to have a good radial system with a voltage-fed antenna such as the voltage-fed T or a λ/2 vertical. These verticals require longer radials to do their job efficiently compared to current-fed verticals.

4.3.2. The Far Field

In the far field, the requirement for a good ground with a long vertical is much more important than for a λ/4 vertical. I have modeled the influence of the ground quality on the gain of a vertical by the following experiment.

- I compared three antennas: a λ/4 vertical, a voltage-fed λ/4 T (also called an inverted ground plane) and a λ/2 vertical.
- I modeled all three antennas over average ground.
- I put them in the center of a disk of perfectly conducting material and changed the diameter of the disk to determine the extent of the Fresnel zone for the three antennas.

The results of the experiment are shown in **Fig 9-61**. Let us analyze those results.

- With a conducting disk λ/4 in radius (equal to a large number of λ/4 radials) the λ/4 current-fed vertical is almost 2 dB better than the voltage-fed λ/4 and the λ/2 vertical.
- The λ/4 vertical remains better than the other antennas up to a disk size of 1.5-λ diameter. This means that over good ground you must be able to put out radials at least 2-λ long with a λ/2 vertical before it shows any gain over the λ/4 current-fed vertical.
- The voltage-fed λ/4 vertical (voltage-fed T) equals the current-fed λ/4 for a disk size of at least 2 λ in diameter. This is because the current maximum is at the top of the antenna, which means that for a given elevation angle, the Fresnel zone (where the main wave hits the ground to be reflected) is much farther away from the base of the vertical than is the case with a λ/4 current-fed vertical. In other words, there is no advantage in using such a voltage-fed λ/4 antenna.

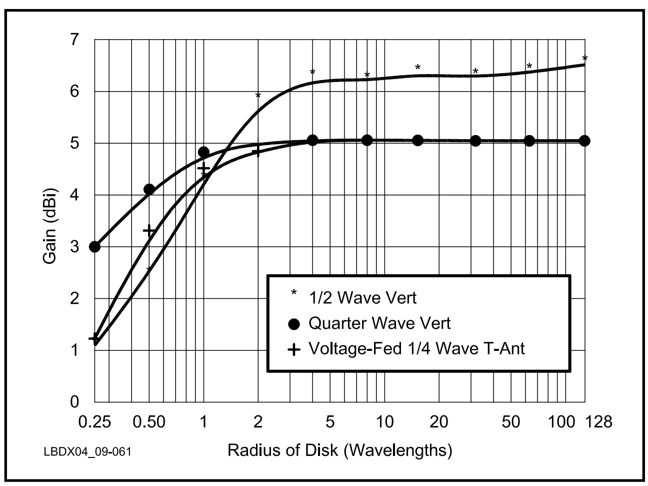


Fig 9-61 — Gain of three types of verticals over a perfectly conducting disk of varying radius. The ground beyond the disk is of good quality. This means that the λ/2 vertical requires 0.6-λ radials to perform as well as the λ/4 vertical with λ/4 radials. Be aware that the radiation angle of the λ/2 vertical will be much lower, however.

- For both the voltage and the current-fed λ/4 vertical, the Fresnel zone is situated up to 4 λ away from the vertical. For the λ/2 vertical, the Fresnel zone stretches out to some 100 λ!

4.4. In Practice

On 40 meters, a height more than λ/4 (10 meters) should be easy to install in most places. In many cases it will be the same vertical that is used as a λ/4 vertical on 80 meters.

I have been using a 5λ/8 vertical for 40 meters for more than 20 years with good success. With Beverage receiving antennas it has always been a relatively good performer. After switching to a 3-element Yagi at 30 meters some 15 years so, I found out that the vertical solution was far from ideal.

Earl Cunningham, K6SE (SK) experienced similar results: "I used a grounded 1/2-λ vertical in the Houston/Gulf Coast area where the soil conductivity is abnormally high. It was a super performer. The same vertical here in the desert (Palmdale, CA) was a ho-hum performer, even with a much more extensive ground radial system."

A similar testimony comes from Tom Rauch, W8JI, who wrote: "…I had the same results using BC arrays on 160 meters. The 250-ft to 300-ft verticals stunk; my ¼-λ vertical would beat them. I find the same effect on 80 meters."

Figs 9-9 and 9-12 earlier in this chapter give the base resistance, $R_{Rad(B)}$, and feed-point reactance for monopoles as a function of the conductor diameter in degrees, and in Figs 9-10 and 9-13 as a function of the antenna length-to-diameter ratio. The graphs are accurate only for structures with rather large diameters (not for single-wire structures) and that have *uniform* diameters. A conductor diameter of 1° equals 833/f (MHz) in mm.

5. MODELING VERTICAL ANTENNAS

ELNEC as well as other versions of *MININEC* are well suited to do your own vertical antenna modeling, as is the *NEC-2*-based *EZNEC* program. Be aware, however, that all *MININEC*-based antenna modeling programs assume a perfect ground under the antenna base for computing the impedance of the antenna. You cannot use these programs to assess the efficiency of the vertical, where I have defined efficiency as:

$$\text{Eff} = \frac{R_{rad}}{R_{rad} + R_{loss}}$$

MININEC will show the influence of the reflecting ground in the far field that creates the low-angle radiation pattern of the vertical antenna. If you want to include the losses of the ground, you can insert a resistance at the feed point, having a value equivalent to the assumed loss resistance of the ground (see Table 9-1).

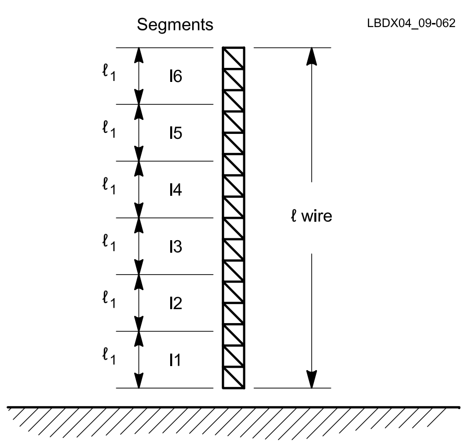
5.1. Wires and Segments

In modeling terminology, a *wire* is a straight conductor and is part of the antenna. A *segment* is a part of a wire. Each wire can be broken up into a number of segments, usually all with the same length. Each segment has a different current. The more segments a wire has, the closer the current (pulse) distribution will come to the actual current distribution. There are limits, however.

- Many segments take a lot of computing time.
- Each segment should be at least 2.5 times the wire diameter (according to *MININEC* documentation).

There is no general rule about the minimum number of segments that should be used on a wire. There is only the cut-and-try rule, where you gradually increase the number of segments and look for the point where no further significant changes in the results are observed. This is commonly called *convergence testing* (see also Chapter 4).

Fig 9-62 shows an example of a straight vertical for 80 meters (19 meters long). This antenna consists of a single wire. To evaluate the effect of the segment length, I varied the number of segments in the wire from five to 150. Gain and pattern are very close to modeling with only five segments



Number of Segments	*Segment Length (mm)*	*2 mm OD Wire (AWG No. 12)*	*250 mm Mast Diam*
5	3800	43.1 – *j*10.9	36.0 + *j*3.0
10	1900	34.4 – *j*11.6	37.0 + *j*4.1
20	950	34.6 – *j*11.2	37.5 + *j*5.2
30	630	34.6 – *j*11.0	37.7 + *j*5.8
50	380	34.7 – *j*10.8	38.0 + *j*6.2
70	240	34.7 – *j*10.8	38.0 + *j*6.3
100	190	34.7 – *j*11.1	34.5 – *j*13.0
150	130		34.0 – *j*14.0

Fig 9-62 — *MININEC* analysis of a straight 19-meter long vertical antenna shown in the drawing. The analysis frequency is 3.8 MHz. *MININEC* impedance results are shown as a function of the number of segments in the table. Note that for reliability with a "fat" (200-mm) vertical, the maximum number of segments (in this case segments = pulses) is 70. The *MININEC* documentation states that the segment length should be greater than 2.5 times the wire diameter (2.5 × 200 mm = 500 mm). In this particular case errors occur when the segment length is smaller than the wire diameter.

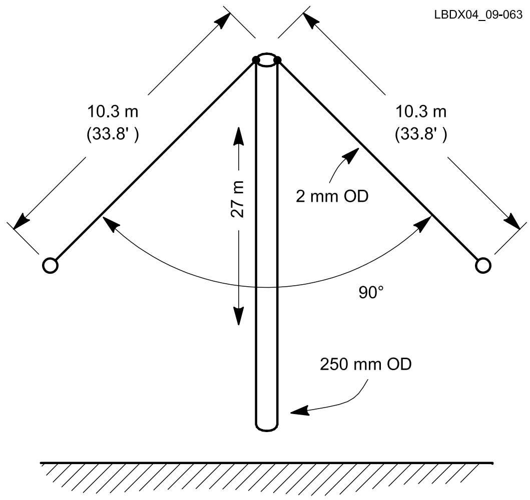


Fig 9-63 — Impedances calculated by *MININEC* for a top-loaded 1.8-MHz vertical, using a 250-mm OD mast and two 2-mm OD slant loading wires. The segment lengths are stated in mm. A large number of segments on all wires always gives more reliable results, provided the segment length is not very different. Judicious choice of segment length on the different wires can also yield very accurate results with a smaller number of total segments. To obtain accurate impedance results using *MININEC*, the wire sections near the acute-angle wire junctions must be short.

	Vertical Mast				*Slant Wires*		
No. of Segments	*Segment Length (min.)*	*Segment Length (max.)*	*No. of Segments*	*Segment Length (min.)*	*Segment Length (max.)*	*Total Pulses*	*Impedance*
3	9000	9000	3	3765	3765	9	9.0 – *j*184.0
5	5400	5400	5	2260	2260	1	14.2 – *j*107.0
10	2700	2700	10	1130	1130	3	16.2 – *j*83.5
20	1350	1350	20	565	565	6	16.6 – *j*87.1
30	900	900	30	437	437	9	16.7 – *j*77.0
40	625	625	40	282	282	120	16.6 – *j*76.5
50	540	540	50	226	226	150	16.8 – *j*76.3
10	2700	2700	5	2260	2260	15	16.8 – *j*78.6
6	4500	4500	2	5650	5650	10	16.8 – *j*80.4
5	5400	5400	2	5650	5650	9	16.7 – *j*80.7
35	770	770	2	5650	5650	39	14.9 – *j*103.0

f = 1.8 MHz
length of vert mast = 27 m
slant wire = 10.3 m

compared to 150. For impedance calculations, at least 20 sections are required for a reasonably accurate result. The table in Fig 9-62 also shows an example of too many segments for a vertical measuring 250 mm in diameter. As the segment length becomes very short in comparison to the wire diameter, the results become erroneous.

5.2. Modeling Antennas with Wire Connections

When the antenna consists of several straight conductors, things become more complicated. **Fig 9-63** shows the example of a 27-meter high vertical tower (250-mm OD), loaded with two sloping top-hat wires, measuring 2-mm OD (AWG #12).

The standard approach is to use three wires, one for each of the three antenna parts, and divide the three conductors into a number of segments (which are equal length inside each wire).

To obtain reliable results, you must make sure that the lengths of the segments near the junctions of wires are similar. The table in Fig 9-63 shows the impedance obtained for the top-loaded vertical with different numbers of segments. A large number of segments on the vertical mast (eg, 35 segments, which results in a segment length of 770 mm), together with a small number of segments on the sloping wires, give an unreliable result, while a good result is obtained with a total of just nine segments if the lengths are carefully matched. The segment *tapering* technique, described in Chapter 14 on Yagis and quads, can also be used to minimize the number of segments and improve the accuracy of the results.

5.3. Modeling Verticals Including Radial Systems

MININEC does not analyze antenna systems with horizontal wires close to the ground. Therefore, modeling ground systems as part of the antenna requires the *NEC* software. *NEC-2*, or software such as *EZNEC*, which uses the *NEC-2* engine, can model radials over ground. There seem, however, to be documented cases (models verified against real-world measurements) of *NEC-2* giving very optimistic results (sometimes up to nearly 6 dB too high gain).

NEC-4 can model buried radials (see Chapter 4, Section 1.4), but apparently still shows optimistic gain values for wires very close to ground. The results of modeling buried radial systems (see Sections 2.1.2 and 2.1.3) are largely confirmed by N7CL's experimental work. While we can't be absolutely sure real gain figures match modeled numbers within fractions of a dB, nonetheless the trends are certainly correct.

5.4. Radiation at Very Low Angles

Most modeling programs amateurs use show zero radiation at zero elevation and very little at low angles, unless over saltwater. How can we hear ground-wave signals even over average ground? We should not, according to what the model tells us. Experiments show that in real life the very low-angle performance of vertical antennas is better than these modeling programs tell us.

5.5. Measurements, Verifications, Real Life

It is beyond the reach of almost all amateurs to do real-life experiments with low band antennas. The reasons are many. Verifications of modeling results can only be done by few, because of lack of test equipment and most of all the necessary acres.

On the other hand, modeling involves mathematics, and a computer can show us results expressed in fractions of a dB. Some models (the great majority) were never verified, and I suspect that the error could be many dB.

I always get nervous when I read articles that show results of gain, impedance and other parameters with two or even three digits after the decimal point. Antenna modeling is a mathematical exercise where we *try* to predict the behavior of the antenna. The behavior is a matter of physics. Good models can describe the physical properties quite well, but if after a lengthy modeling session one concludes that model A is better than model B because the math shows a gain advantage of 0.1 dB, this is when I get nervous.

Modeling most often does seem to make sense if you compare one model to another model (modeled with the same program and using the same methodology), but you should not automatically conclude that the number results apply directly to the real world!

L.B. Cebik, W4RNL (SK) once said that modeling is only a comparison of one model to another model, and the models do not in any form represent real conditions. W4RNL was an authority in the "science" of antenna modeling. Coming from him, this statement is a serious warning!

6. PRACTICAL VERTICAL ANTENNAS

A number of practical designs of verticals for 40, 80 and 160 meters are covered in this section, as well as dual and tri-band systems. A number of practical matching cases are solved, and the component ratings for the elements are discussed. All the L networks have been calculated using L-Network Design module from the *New Low Band Software.*

6.1. Single-Band Quarter-Wave Vertical

Fairly large-diameter conductors are used for various reasons, such as increasing the bandwidth (by increasing the D/L ratio) or simply for mechanical reasons. The effective diameter of wire cages and flat multi-wire configurations is covered in Chapter 8 in Section 1.4.

Often, triangular tower sections (such as Rohn 25 or equivalent) are used to make vertical antennas. The effective equivalent diameter of a tower section is shown in **Fig 9-64**. A tower section measuring 25-cm wide, with vertical tubes measuring 2.5-cm diameter, has an equivalent diameter of 0.7 × 25 = 17.5 cm.

The length of a resonant full-size quarter-wave vertical

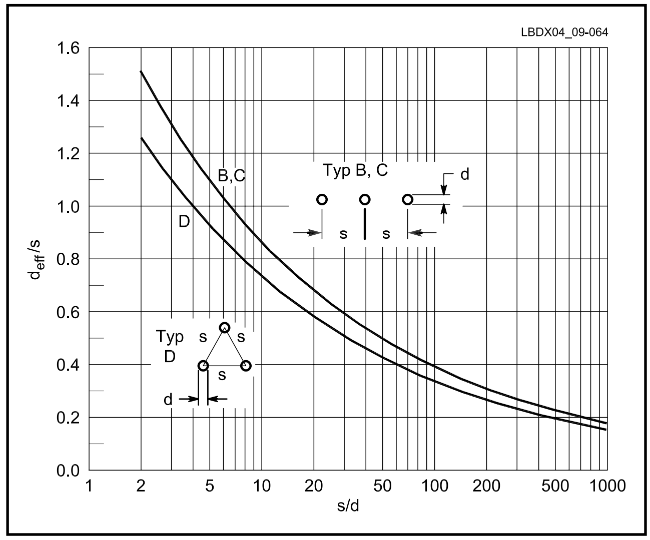


Fig 9-64 — Normalized (round solid conductor) effective diameter of a triangular tower section as a function of the vertical tube diameter (d) and the tower width (S). The graph for three parallel conductors is also given (curve BC). (After Gerd Janzen, *Kurze Antennen*.)

depends on its physical diameter. See **Table 9-15**, which shows the physical shortening factor of a λ/4 resonant antenna as a function of the ratio of antenna length to antenna diameter. The required physical length is given by:

$$L = \frac{74.95 \times SF}{f_{MHz}}$$

where

L = length (height) of the vertical in m
SF = correction factor (from Table 9-15)
f_{MHz} = design frequency in MHz.

Quarter-wave verticals are easy to match to 50-Ω coaxial feed lines. The radiation resistance plus the usual earth losses will produce a feed-point resistance close to 50 Ω.

If you don't mind using a matching network at the antenna base, and if you can manage a few more meters of antenna height, the extra height will give you increased radiation resistance and higher efficiency (see Section 3.8). The feed-point impedance can be found in the charts of Figs 9-9, 9-10, 9-12 and 9-13. More accurate results can be obtained through modeling.

Consider the following examples:

Table 9-15
λ/4 Resonance for Vertical as Function of Length/Diameter

Length/Diameter Ratio	*Shortening Factor (%)*
5000	97.3
2500	97.1
1000	96.8
500	96.2
250	95.7
100	94.6
50	93.4

Example 1

Tower height = 27 meters
Tower diameter = 25 cm
Design frequency = 3.8 MHz

Modeling with *EZNEC* (*NEC-2*) we find:

R = 190 Ω
X = + *j* 218 Ω

Let's assume we have a very good ground radial system, with an equivalent ground resistance of 5 Ω. We calculate the matching L network with the following values:

$Z_{in} = 195 + j\,218\ \Omega$
$Z_{out} = 50\ \Omega$

The values of the matching network were calculated for 3.8 MHz (**Fig 9-65**). The low-pass filter network gives a little additional harmonic suppression, while the high-pass assures a direct dc ground for the antenna and some rejection of medium-wave broadcast signals.

Example 2

Another interesting design was already explained in Section 3.9, example 2. A single series capacitor is all that's needed to match the vertical and obtain a very attractive bandwidth. What type of capacitor should you use? The current requirement can be calculated as follows:

$$I = \sqrt{P / R}$$

For 1.5 kW,

$$I = \sqrt{1500 / 50} = 5.48 \text{ A}$$

Voltage across the capacitor (having an impedance of 26.6 Ω):

$$E = \sqrt{P \times R} = \sqrt{1500 \times 26.6} = 200\ V_{RMS}$$

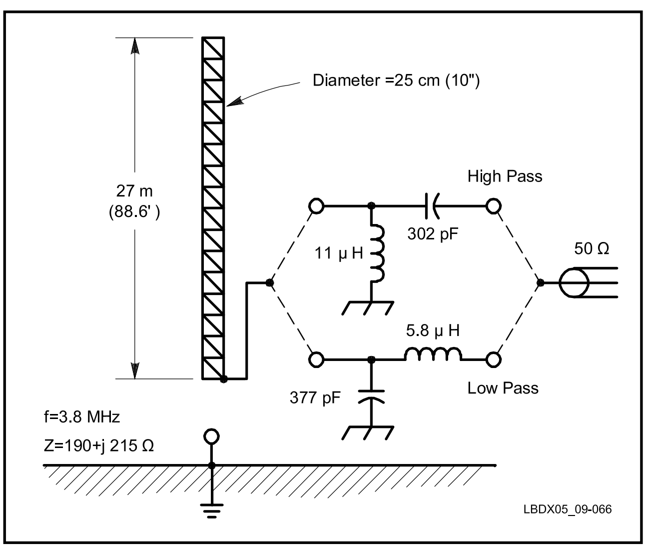


Fig 9-65 — Making the vertical 40% longer than λ/4 wave, raises the R_{rad} substantially, which is advantageous as far as the efficiency is concerned. An L network (either high- or low-pass) can be used to match to the 50-Ω feed line.

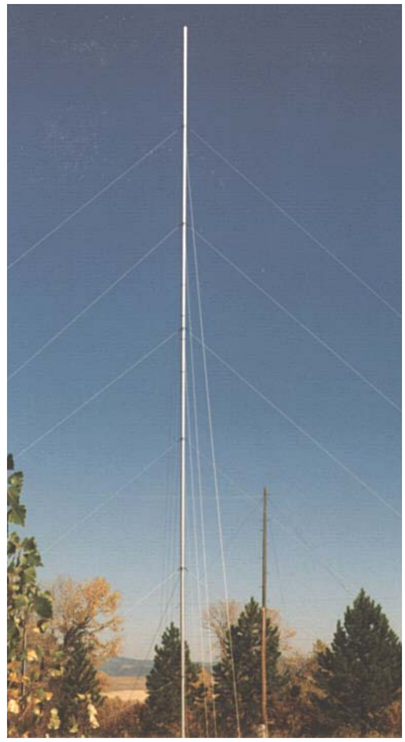


Fig 9-66 — Ninety-foot (27-meter) irrigation-pipe vertical at W7LR in Montana, which is used for both 80 and 160 meters.

For voltage rating you need to take peak values, plus a certain safety margin. A 1 kV rating is what I would use.

6.1.1. Mechanical Design

I don't want to give many detailed mechanical designs that list materials, tubing diameter and other details since availability is different in every country. **Fig 9-66** shows an example of a vertical made from irrigation pipe, a popular material for tall verticals.

Guy Hamblen, AA7ZQ/2, described an attractive 80/75-meter design that uses 12-foot long aluminum tubing sections ranging from 1.5-inch OD to 0.875-inch OD. He also describes the installation details (Ref 7819).

If you consider making a vertical with a rather long unguyed top section, you can use the Element Strength Module of the *ON4UN Yagi Design Software*. Using the software you can design a Yagi element with a length equal to twice the length you need for the non-guyed top section of the vertical. Because this top section, unlike the Yagi half-element, will not be loaded by its own weight (causing the sag in a Yagi element), the vertical section will have an added safety factor.

Fig 9-67 shows the design for an 80-meter vertical using 4-inch and 3-inch aluminum irrigation tubing, as designed by Steve Kelly, K7EM. The verticals are mounted on 6×6-inch pressure treated lumber. The total length of each post is 12 feet, of which 4 feet is in the ground. The arms that hold the verticals in place are made from 2×6-inch lumber. Steve used ⅜-inch threaded rod to bind the arms to the posts and a ½-inch threaded rod goes through the base of each vertical (see **Fig 9-68**). The ½-inch rod acts as a hinge for raising and lowering and is insulated from the vertical with PVC tubing. Steve recently replaced the 2×6-inch lumber with ½-inch thick Plexiglas sheet. The bottom ends of the 4-inch tubing is insulated by 4-inch (inside diameter) PVC pipe. Steve mentions splitting

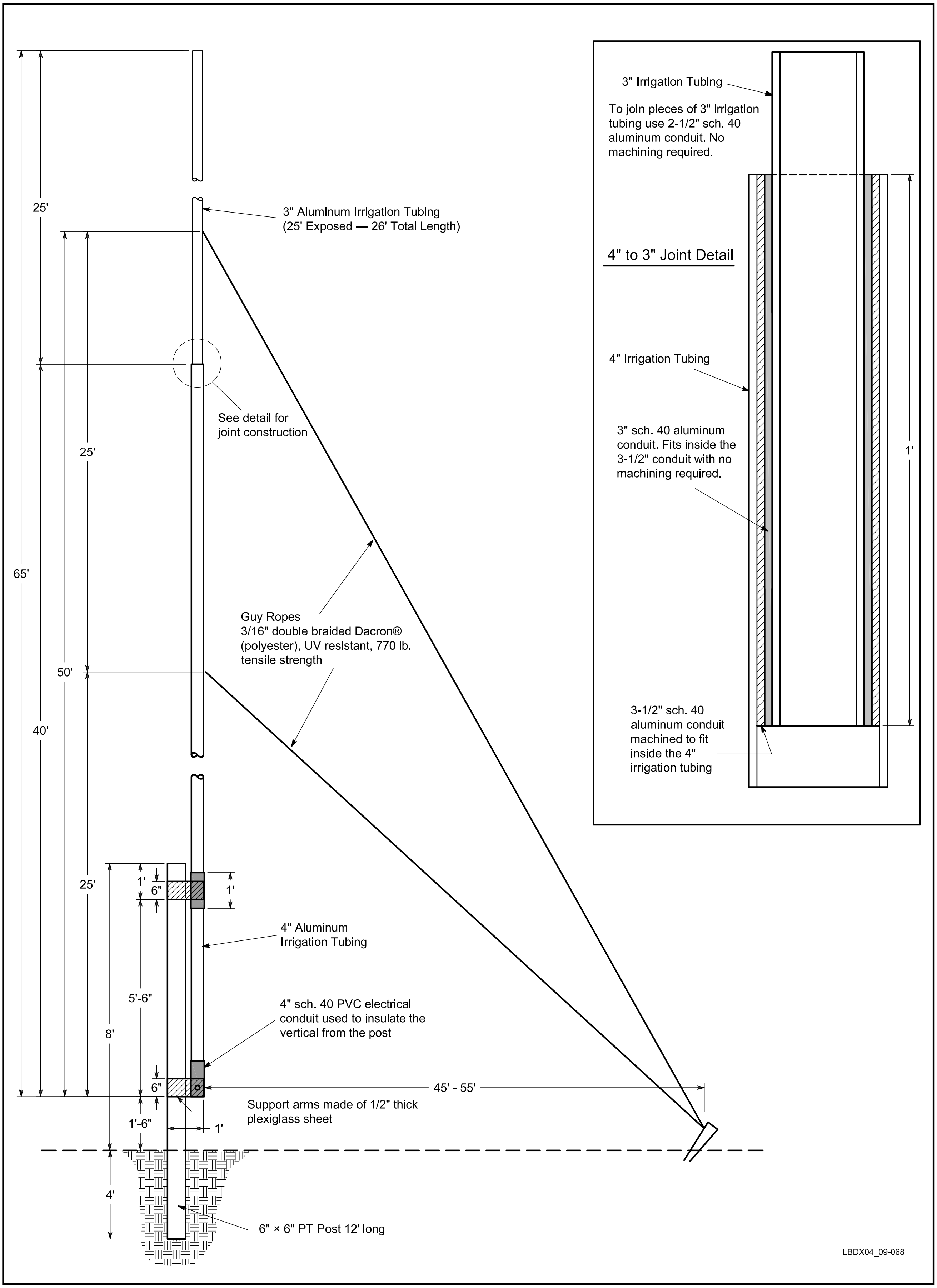


Fig 9-67 — Construction details for 80-meter λ/4 elements made of 4-inch and 3-inch irrigation tubing designed by K7EM.

Fig 9-68 — Base of one of the irrigation-pipe 80-meter verticals constructed by K7EM.

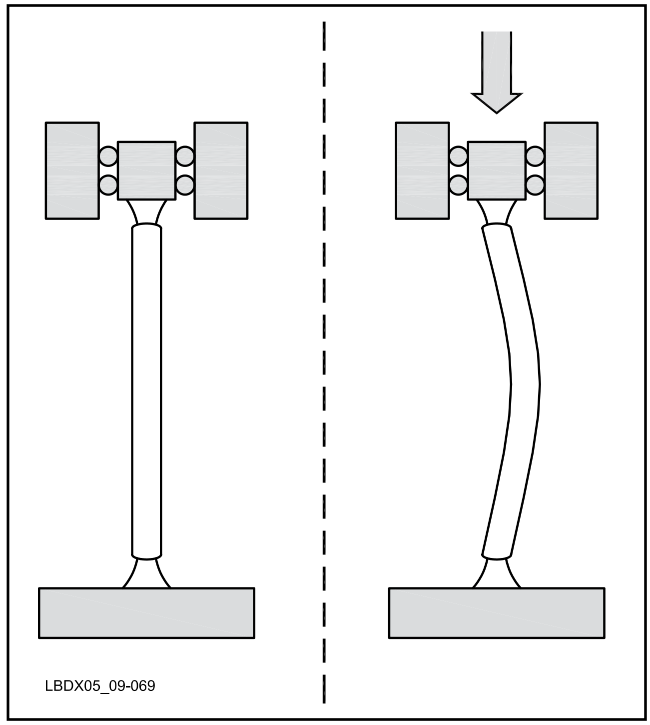


Fig 9-69 — A column under an axial load will buckle if the material is flexible enough and the load high enough.

this pipe lengthwise, heating it with a special PVC bending blanket and then sliding it over the 4-inch irrigation tubing.

Element Buckling (Bending)

When you use small tower sections for building an 80-meter or 160-meter vertical there will be no need for a lot of guying (provided you are not using a capacitance hat). Two sets of guy wires should be enough.

If you use aluminum tubing, especially to build a 160-meter vertical, make sure that the distances between guying heights is small enough to prevent buckling of the section between the guying points by either excessive dead weight or in combination with wind loading. Ice loading on the vertical and the guy wires can add a lot of weight and the slightest wind can make a section between two guying points bend and break.

In construction, buckling is a failure mode characterized by a sudden failure of a structural member subjected to high compressive stresses, where the actual compressive stresses at failure are smaller than the ultimate compressive stresses that the material is capable of withstanding (see **Fig 9-69**). This mode of failure is also linked to the elasticity of the material. Aluminum is much more elastic than steel, which means that the compressed column may more easily deform due to external influences (wind loading) or internal causes (homogeneity of the material). Another reason can be eccentric loads on the vertical tube (eg due to improper guying or guy attachment). Once a guyed element starts buckling, the guy wires, when submitted to extra loads (wind, ice etc) will pull the buckled section down and it will collapse — most likely failing in its center.

The phenomena of buckling is the reason why we have horizontal bars at regular distances in our tower sections. The oblique bars are there to prevent torsion of the tower sections.

6.2. Top-Loaded Vertical

The design of loaded verticals has been covered in great detail in Section 3.6. Capacitive top loading using wires (usually slightly sloping) are quite easily constructed from a mechanical point of view. It is more difficult to insert a husky loading coil in a vertical antenna. In addition, because of their intrinsic losses, loading coils are always a second choice when it comes to loading a vertical.

A top hat wire-loaded vertical for 160 meters is described in Section 6.4 as part of an 80/160-meter duoband system. Inverted-L antennas, which are a specific form of top-loaded verticals, are the subject of Section 7.

6.3. Duo-Bander for 80/160 Meters

Full-size, $\lambda/4$ verticals (40 meters tall on 160 meters) are out of reach for most amateurs. Often an 80/160-meter duoband vertical will be limited to a height of around 30 meters. This represents an electrical length of 140° at 3.65 MHz and 70° on 160 meters. We can determine R and X from Figs 9-9 and 9-11 or through modeling:

80 meters: $Z = 280 + j\,278\ \Omega$

160 meters: $Z = 17 - j\,102\ \Omega$

We use L-networks to match these impedances to our feed line. It's a good idea to use your antenna analyzer to check the impedances on 80 and 160. They should be close to those mentioned above. Remember that the magnitude of the reactive part depends on the effective diameter of the antenna (large diameter antennas exhibit less reactance). Use the L-networks section of the *New Low Band Software* to calculate the component values. **Fig 9-70** shows the antenna configuration together with the switchable matching system.

6.4. 80/160 Top-Loaded Vertical with Trap

Traps are frequency-selective devices incorporated in radiating elements to adapt the electrical length of the element depending on the frequency at which the element is being used.

Commercial multiband antennas make frequent use of traps. Home-made antennas use the technique less often. There are two types of commonly used traps:

- Isolating traps
- Shortening/lengthening traps

6.4.1. Isolating Traps

An isolating trap is a parallel-tuned circuit that presents a high impedance at the design frequency, effectively decoupling the "outer" section of the radiator from the "inner" section. A good isolating trap meets the following specifications:

- It represents a high impedance on the design frequency.
- It represents as high a Q as possible, together with the high impedance.
- It represents as low a series inductance as possible on the frequencies where the trap is not resonant (minimize inductive loading), unless you want to use the trap off resonance as a loading device, which could shorten or lengthen the element. The LC circuit off resonance acts as an L or a C, depending on which side of resonance you are.

Traps have been described in literature is several configurations:

- Regular LC parallel-tuned circuits.
- Resonant circuits with the coil created by a so-called linear loading device (for example, see KT-34 Yagis).
- Traps made with coaxial cable, where the capacitance of the cable is used as capacitor, and where the cable shield acts as the coil in the resonant circuit.

Losses are the main issue with traps. An ideal trap will have an infinite parallel (shunt) loss resistance (R_p). Tom, W8JI investigated different traps and found the following results (**www.w8ji.com/traps.htm**).

- Copper tubing and vacuum cap: R_p = 300,000 Ω
- 60-pF doorknob and #10 Airdux coil: R_p = 250,000 Ω
- 100-pF doorknob and #12 Airdux coil: R_p = 99,850 Ω
- Mosley TA33: R_p = 79,000 Ω
- Cushcraft A3: R_p = 76,270 Ω
- Coax RG-58/U: R_p = 17,800 Ω
- Teflon-insulated semi-rigid copper tubing type coax: R_p = 45,000 Ω

W8JI adds: "Stubs, linear-loading, and coaxial-cable

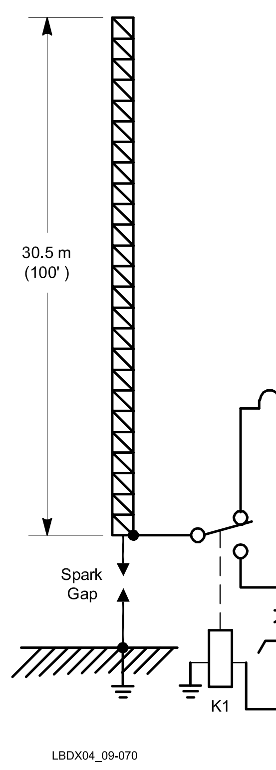


Fig 9-70 — Two-band (80- and 160-meter) vertical system using L-networks to match the 50-Ω feed line. The values of the L-networks were calculated assuming an equivalent ground-loss resistance of 10 and 5 Ω respectively on 160 and 80 meters.

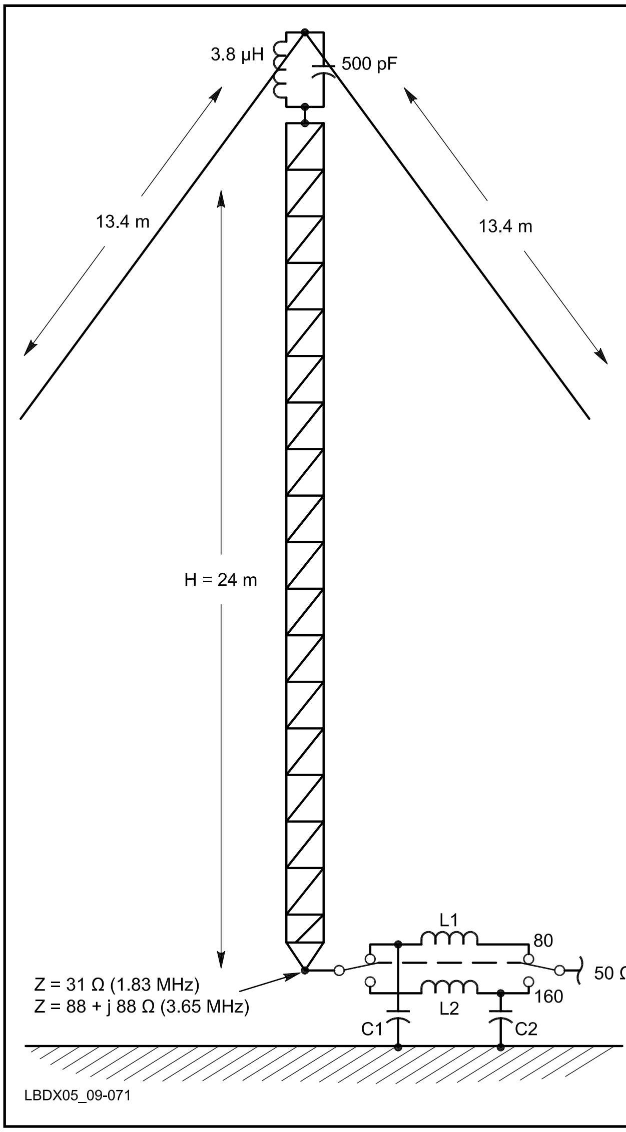


Fig 9-71 — A vertical, somewhat longer than a quarter wave on 80 meters, has an 80-meter isolating trap at its top. The trap serves to decouple the slanting loading wires for 160 meters.

capacitors generally have very low Q compared to other systems. Expect about 1.5 dB or so loss for a coaxial trap in an inverted L or vertical at the trapped frequency. Loss using a small Airdux coil and a doorknob capacitor would be less than 0.25 dB. Loss on other than the exact trapping frequency are insignificant with all types of traps."

The exception would be traps using very low C and high inductive reactance, which may have significant loss at the pass frequency. This type of trap would be the high-inductance coils used to isolate antenna sections while, substantially reducing length on the non-trap frequency (as sometimes used in small Yagis). Those traps can seriously degrade the pass-frequency performance.

Fig 9-71 shows a typical 80/160-meter vertical antenna arrangement where an isolating trap is installed on top of an 80-meter vertical, effectively isolating the sloping top loading wires when operating on 80 meters. In this example the vertical is 24 meters long. Two L-networks will match the impedances on 160 and 80 meters to 50 Ω. An alternative without a trap is described in Section 6.5.

Traps are not widely used in antennas for the low bands, and there is a good reason for that. On the higher bands, a loss of a couple of dB in a trap is not "lethal" for the signals produced. In most cases it will not make all that much difference whether you are S9 or S9 + 3 dB. On 80 and more so on 160, a couple of dB often make the difference between being heard or not being heard. There is not much room for compromising, and especially not for compromise upon compromise upon compromise: 1 + 1 + 1 + 1 + 1 make a lot of dB!

The Battle Creek Special is probably the best known trap vertical for the low bands. Up until 2005 this was the "standard" antenna for DXpeditions. Nowadays DXpeditions of any size operate different low bands (two or three) simultaneously all through the night. That means they need separate antennas for each band, and in addition antennas that are not resonant on the same frequencies (to avoid excessive signal coupling). More on the Battle Creek Special in Section 6.7.

6.4.2. Shortening/Lengthening Traps

If the isolating trap principle were to be used on a triband antenna, it would require two isolating traps. Three-band trap Yagis of the early sixties indeed used two traps on each element half, the inner one being resonant on the highest band, the outer one on the middle band. Modern trap-design Yagis only use a single trap in each element half to achieve the same purpose. Y. Beers, WØJF (SK), wrote an excellent article covering the design of these traps (Ref 680). In this design, the trap is not resonant on the high-band frequency, but somewhere in between the low and the high band. In the balanced design described by Y. Beers, the frequency at which the trap is resonant is the geometrical mean of the two operating frequencies (equal to the square root of the product of the two operating frequencies). For an 80/160-meter vertical, the trap would be resonant at:

$$f = \sqrt{1.83 \times 3.85} = 2.65 \text{ MHz}$$

On a frequency below the trap resonant frequency the trap will show a positive reactance (it acts like an inductor), while above the resonant frequency the trap acts as a capacitor. A single parallel-tuned circuit can be designed that inserts the necessary positive reactance at the lowest frequency and negative reactance at the highest frequency. In the balanced design the absolute value of the reactances is identical for the two bands; only the sign is different. There are five variables involved in the design of such a trap system: the two operating frequencies, the trap resonant frequency, the total length and the L/C ratio used in the trap parallel circuit. The design procedure and the mathematics are covered in detail in the above-mentioned article.

It's clear that the "isolating" trap we designed in Section 6.4.1 was really a lengthening trap as we designed it to be a trap slightly lower than 80 meters.

6.5. The Alternative: Parallel Antennas

Fig 9-72 shows the better alternative for the antenna shown in Fig 9-71, this time without a trap. Instead of a trap you need a little more antenna wire and a DPDT relay to switch from 80 to 160 meters.

In this example we use a 24-meter high small tower

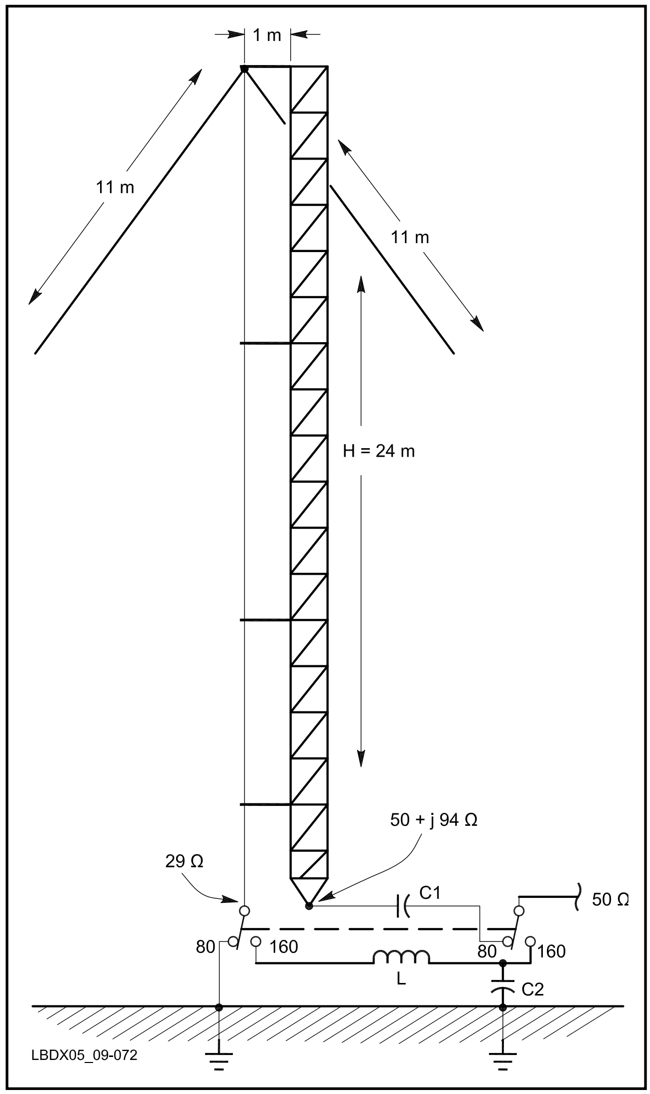


Fig 9-72 — In this duoband 160 and 80 meter vertical antenna system, the 80-meter vertical is used as a mechanical support for a 160-meter wire vertical with sloping top hat wires.

(eg Rohn 25) as a vertical for 80 meters. It is longer than λ/4, and if we take into account a earth loss resistance of 5 Ω, the resistive part of the impedance is close to 50 Ω (actually 50 + *j* 95 Ω as modeled with *EZNEC*). A simple series capacitor with X_C = –95 Ω (~460 pF) is all what we need to match the antenna to 50 Ω.

We now run a wire parallel to the tower (at approximately 1-meter distance) and from the height of 24 meters run two sloping top hat wires (length approximately 11 meters). Cut the exact lengths to tune the antenna to 1.83 MHz. The feed point impedance of the 160-meter loaded vertical is approximately 25 Ω (15 Ω R_{rad} and 10 Ω ground losses). We can best match this with a simple L network (L-C2). The values can be calculated with The L-Network section of the *New Low Band Software* (L = 2.175 μH, C = 1480 pF).

When operating on 80 meters, it is very important that the 160-meter antenna be grounded, otherwise it will act as an 80-meter half-wave antenna and upset the whole antenna system. The DPDT relay takes care of the bandswitching and the grounding.

Note that when operating on 160 meters, the 80-meter vertical needs to be disconnected from ground, which is the case with the switching arrangement as shown.

6.6. The Self-Supporting Full-Size 160-Meter Vertical at ON4UN

A full-size λ/4 vertical antenna for 160 meters is just about the best transmitting antenna you can have on that band, with the exception of an array made of full-size or top-loaded verticals. I use a 32-meter tall triangular section self-supporting tower, measuring 1.8 meters across at the base, and tapering to

Fig 9-73 — Giving out new countries and chasing new countries on 160 meters are not the only hobbies for Rudi, DK7PE, (left) and ON4UN, who are ready to go on a bike trip. In the background is the base of ON4UN's 160-meter vertical showing the cabinet that houses the matching circuitry for the 160-meter vertical and the Four Square 80-meter system.

Fig 9-74 — Self-supporting 39.5-meter λ/4 vertical for 160 meters at ON4UN. The base is 1.8 meters wide and the tower tapers to just a few inches at the top. The tower is shunt fed with a gamma match and also serves as a support for an 80-meter Four Square array made of λ/4 verticals, supported from sloping catenary lines running from the 160-meter tower.

20 cm (8 inches) at the top. I knew that the taper would make the tower electrically shorter than if it had a constant diameter, so that had to be accounted for. On top of the tower I mounted a 7-meter long mast. It is steel at the bottom and aluminum at the top, tapering from 50 mm OD to 12 mm at the top.

To make up for the shortening due to the tower taper, I installed a small capacitance hat at the top of the tower. The highest point I could do this was at 32.5 meters, where the 7 meter mast starts. I decided to try a disk with a diameter of 6 meters, because I had 6-meter-long aluminum tubing available. Two aluminum tubes were mounted at right angles, the ends being connected by bronze wire to make a square. In **Fig 9-73** you can see the base of the vertical in the background and **Fig 9-74** shows the top 20 meters of my antenna.

I hoped I would come close to an electrical λ/4 on 160 meters, and fortunately the antenna resonated on exactly 1830 kHz. In the beginning I had the tower insulated at the base, and was able to measure its impedance, approximately 20 Ω, while I expected 36 Ω with a 0-Ω earth-system resistance. Such a low radiation resistance has been reported in the literature, and must be due to the large tower cross-section. Originally I suspected mutual coupling with one of two other towers (or both), but decoupling or detuning those towers did not change anything.

Fig 9-75 shows the radiation resistance of a λ/4 vertical over a radial system consisting of 60 λ/4 radials, measured as a function of the diameter of the vertical. You can see that for a height/diameter ratio of 44 (eg, a self-supporting tower with a diameter of 1 meter operating at 1.83 MHz) the radiation resistance should be about 20 Ω. The classic 36-Ω figure applies for a very thin conductor!

After a series of unsuccessful attempts to use the vertical on 80 meters, I grounded the tower and shunt fed it using a gamma match for 160 meters and dropped the idea of using this "much-too-long" tower on 80 meters. A tap at a height of 8 meters and a 500-pF series capacitor provided a 1:1 SWR on Top Band, and a 2:1 SWR bandwidth of 175 kHz. The gamma wire is approximately 1.5 meters from the tower. This vertical really plays extremely well. I use quite an extensive radial system, consisting of approximately 200 radials ranging from 18 to 40 meters in length. The tower now also supports a Four Square sloping λ/4 vertical array as described in the chapter on vertical arrays.

Fig 9-76 shows the vertical's base and the cabinet housing the series capacitor for the 160-meter gamma match (as well as the hybrid coupler for the 80-meter Four Square array). I obtained detailed feed-point information for my 160-meter vertical with the assistance of ON6WU and his professional-grade test equipment (HP network analyzer). See Fig 9-76.

6.7. The Battle Creek Special Antenna

Everyone familiar with DX operating on 160 meters has heard about the Battle Creek Special and its predecessor, the Minooka Special. These antennas are transportable verticals for operating on the low bands. The Minooka Special (Ref 761) was designed by B. Boothe, W9UCW, for B. Walsh, WA8MOA, to take on his trips to Mellish Reef and Heard Island many years ago.

Basically the antennas were designed to complement a triband Yagi on DXpeditions to provide excellent six-band coverage for the serious DXpeditioner. The original Minooka was a 40 through 160-meter antenna, using an L network for matching and an impressively long 160-meter loading coil near the top. WØCD built a very rugged and easily transportable version of the Minooka Special, but soon found out that the slender loading coil simply melted when the antenna was taking high power for longer than a few seconds. No wonder! It was more than 100-cm long with a diameter of only 27 mm. Michaels, W7XC (SK), later calculated the Q factor of the coil

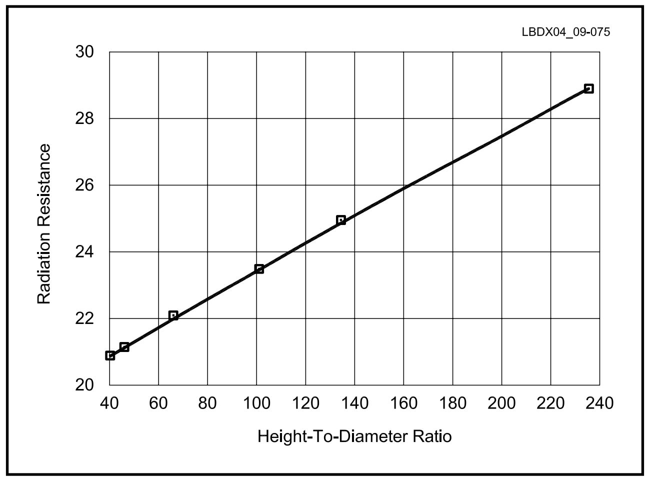


Fig 9-75 — The feed-point resistance of a resonant λ/4 vertical over 60 λ/4 radials, as a function of the conductor diameter. Verticals made of a large-diameter conductor, such as a tower, exhibit much lower feed-point resistances than encountered with wire verticals.

Fig 9-76 — Roger Vermet, ON6WU, with his professional antenna measuring setup, tuning the new vertical at ON4UN. An HP network analyzer is used which directly produces a Smith Chart. Although such sophisticated test equipment is not necessary to tune a vertical, it is very instructive to know the impedance (admittance) curve.

to be around 20! That's an equivalent loss resistance of 100 Ω!

WØCD improved the antenna both mechanically and electrically. Instead of developing a better loading coil, he simply did away with the delicate part, and replaced the loading coil with a loading wire. His design uses two sloping wires, one for 80 meters and one for 160, which now makes it really an inverted L, but nothing would prevent you from using a T-shaped loading wire as described in Section 3.6.2.3.

The new design, named the Battle Creek Special, takes 1.5 kW of RF on SSB or CW without any problem for several minutes. For continuous-duty digital modes the RF output should not exceed 600 W. An 80-meter trap isolates the loading wires for 80 and 160.

The section below the 40-meter trap is 9.75 meters long, which makes it a full-size quarter-wave on that band. The SWR bandwidth is less than 2:1 from 7 to 7.3 MHz.

On 80 meters the 15-meter length of tubing below the 80-meter trap, together with the loading wire, make it an inverted L. The antenna will cover 3.5 to 3.6 MHz with an SWR of less than 2:1. On 3.8 MHz the antenna is "too long," but a simple series capacitor of 200 to 250 pF will reduce the SWR to a very acceptable level (typically 1.3:1).

On 160 meters the entire vertical antenna plus the top-loading wire make it a λ/4 L antenna. The SWR is typically 2:1 over 20 kHz, indicating a feed-point impedance of approximately 25 Ω (depending to a large extent on the quality of the radial system).

There are several ways to obtain a better match to the feed line. WØCD uses an unun with a 2:1 impedance ratio (see also Chapter 6 on Feed Lines and Matching). The unun is switched in the circuit on 160 meters, and out of the circuit on 80 and 40 meters. Under certain circumstances it can be even advantageous to use the unun on the higher bands as well. The unun is an unbalanced-to-unbalanced wideband toroidal transformer (Ref 1521 and 1522). WØCD actually built a 9:4 (2.25:1) balun, and removed the top turn to get an exact 2:1 ratio.

Another alternative is to use an L network. A simple tunable L network that has been especially designed for matching "short" 160-meter loaded verticals is shown in **Fig 9-77** and **Fig 9-78**. The L network was made by ON7TK and has been traveling around the world on various DXpeditions (A61, 9K2, FOØC, etc).

The Battle Creek Special uses high strength aluminum tubing, 6061-T6 alloy, in sizes ranging from 2 inches to 1 inch (5 to 2.5 cm). The guy lines are 2.4-mm Dacron double-braided rope with a rating of 118 kg breaking strength. Wind survival rating is 160 km/hr assuming proper guy-rope anchors. It is guyed four ways at three levels so the side guy ropes act as a hinge allowing it to be "walked up" by one person.

The original traps were coaxial-cable traps using RG-58, but they ran too hot with power levels over 800 W. Instead of changing to Teflon coax the designers decided to switch to regular L/C traps with the inductor made of #10 wire and the capacitor made from some lengths of RG-213 with 100 pF/meter. The coaxial capacitors fit inside the aluminum mast sections. A single open-ended coax stub of about 90-cm length (90-pF) is used for the 40-meter trap and two parallel-connected pieces of coax of approximately 120-cm length each (240 pF total) are used for the 80-meter trap.

WØCD recommends using at least 30 radials, each of 20-meters length. I consider this a bare minimum. The Battle

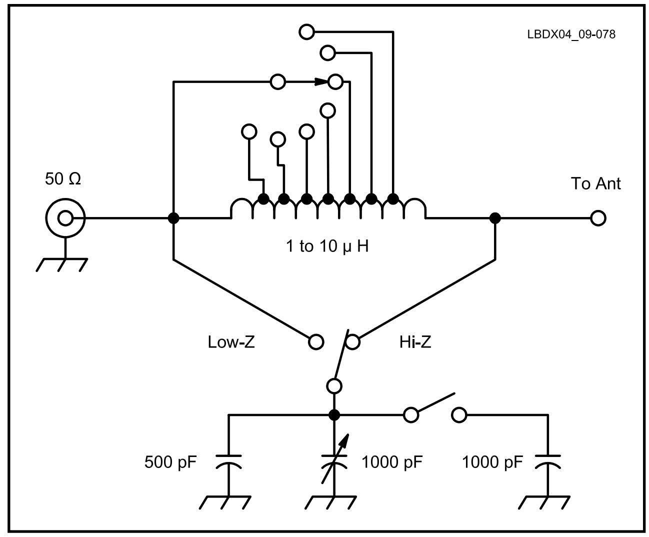


Fig 9-77 — L network to be used with inverted-L antennas and other loaded 160-meter verticals. With the component values shown, impedances in the range 20 – j 100 to 100 + j 100 Ω can easily be matched on 160 meters.

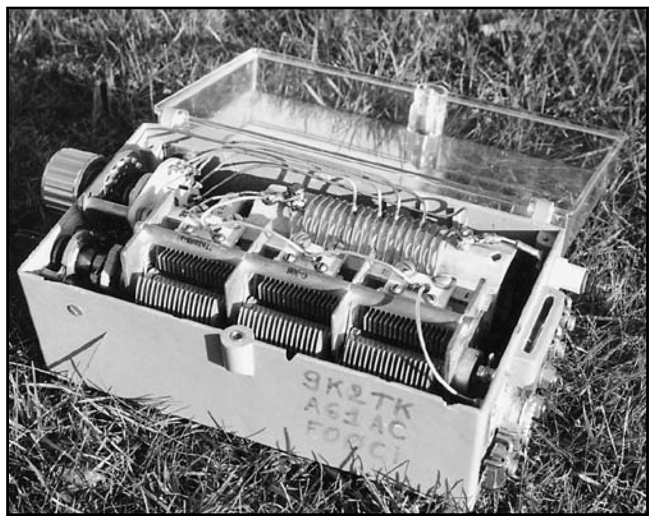


Fig 9-78 — The L network of Fig 9-77 is contained in a small plastic housing. This particular unit was built by ON7TK and used on several DXpeditions (A61, 9K2, FOØC).

Creek Special is not for sale, but is available for loan to DXpeditions to rare countries. Interested and qualified DXpeditioners should contact W8UVZ for further details. The antenna was used at Bouvet on 80 and 160 meters in 1989/90, and during the DXpeditions to ZSØZ, 7P8EN, 7P8BH, G4FAM/3DA, 3Y5X, 5X4F, ZS8IR, XRØY, VP8SGP, YKØA, 8Q7AJ, VKØIR, ZS9Z, V51Z, P4ØGG, CY9AA, ZS6EX, ZS6NW, AHØ/AC8W, AL7EL/KH9, XF4DX, AH1A, 3YØPI, 9MØC, J37XT, K5VT/JT, VK9LX, ZK1XXP, 3B7RF and many other locations with great success.

The entire antenna, with its base, guy-wires and radials is packed in a strong wooden case for safe transport to the remotest DXpedition spot. The package weighs 30 kg (66 lb). For

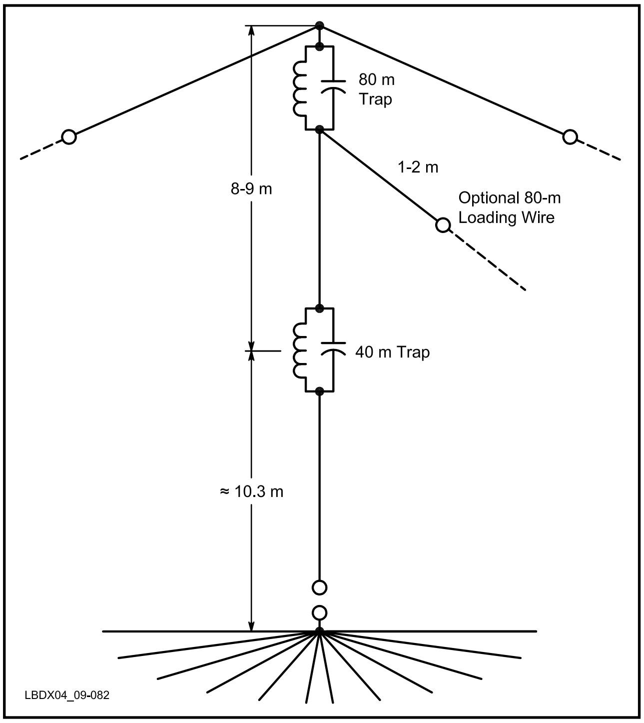


Fig 9-79 — This version of the famous Battle Creek Special vertical is about 18 to 19 meters tall. If resonance on 80 cannot be achieved where wanted (if the support is not high enough), an 80-meter loading wire (typically 1 to 2 meters long) can be added as shown. Slope this wire at right angle to the 40 and 160-meter loading wires if possible. The sloping angle of the 160-meter loading wire has a pronounced influence of the impedance of the antenna. You can use this to tune the antenna on 160. Keep the wire as horizontal as possible, however. It's better is to trim the length of the wire to tune it.

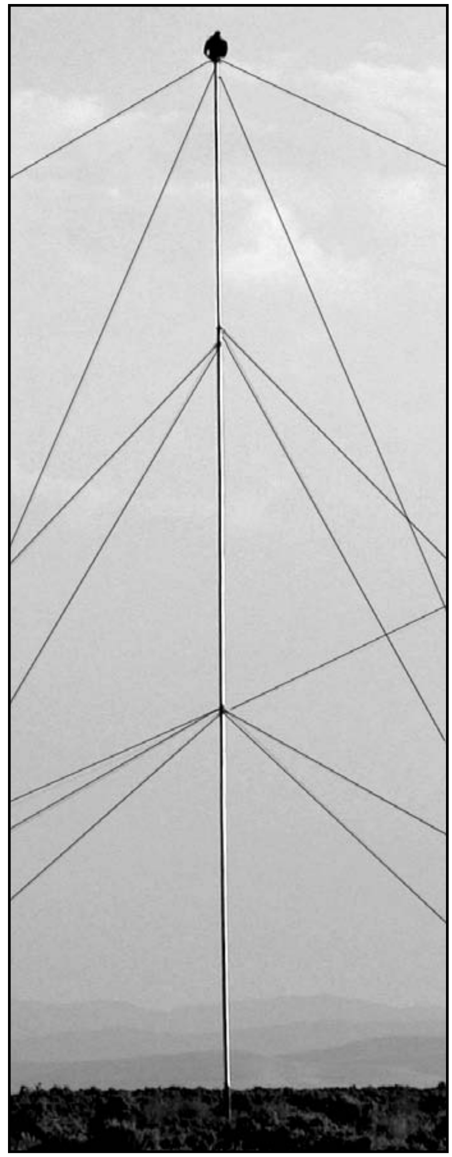


Fig 9-80 — This 20-meter tall vertical with an eagle sitting on top is located in the desert at N7JW's remote-antenna setup in southern Utah.

construction details, visit: **www.ok1rr.com** (search for Battle Creek Special). A wire-type Battle Creek special vertical is shown in **Fig 9-79**.

6.8. A Very Attractive 160-Meter Vertical

Remember that a short vertical is as good as a full-size vertical if the losses in the system are zero. That means that your loading system has no losses (which means top loading). It also means that your ground losses are zero, which you can come close to if you use 100 λ/4 long radials or if you operate over saltwater. K7CA and N7JW developed such a vertical, which they both use in their 160-meter arrays (see Chapter 11), but which is very attractive as a single vertical as well. See **Fig 9-80**. The design set out to achieve R_{rad} = 12.5 Ω, so that a simple λ/4 coax transformer can be used to match a 50-Ω feed line.

The vertical is 20 meters tall, and is made of aluminum tubing, top loaded with two in-line-sloping top hat wires, each approximately 18 meters long and sloping at an angle of about 55°. The tips end up at a height of approximately 10 meters. This yields a resonant frequency of 1830 kHz and the desired feed-point impedance of 12.5 Ω (see similar design in Section 3.8 and Table 9-10). Note that in N7JW's case, this low impedance is reached with 120 quarter-wave radials under the antenna. All you need to do is use a 4:1 broadband transformer (2:1 turns ratio on an appropriate ferrite core) to have a perfect 50-Ω match. This short antenna has a 2:1 SWR bandwidth of 60 kHz and a 1.5:1 SWR bandwidth of 40 kHz, more than decent!

If you want play in the top league of the 160-meter DXers, make no compromises in your ground system! This is also the ultimate performer above saltwater, where just two elevated radials in a gull-wing configuration would do the job. This would be quite an attractive design for DXpeditions where the antenna can be set up over saltwater.

6.9. Using the Beam/Tower as a Low-Band Vertical

The tower supporting the HF antennas can often make a very good loaded vertical for 160 meters. A 24 meter tall tower with a triband or monoband Yagi, or a stack of Yagis, will exhibit an electrical length between 90° and 150° on 160 meters. These are lengths that are very attractive for low-angle work on 160.

6.9.1. The Electrical Length of a Loaded Tower

You can use **Fig 9-81** to determine the electrical length of a tower loaded with a Yagi antenna. The chart shows the situation for a tower with an effective diameter of 30 cm, loaded with five different types of Yagis, ranging from a 3-element, 20-meter Yagi to a 3-element 40-meter full-size Yagi. These figures are for Yagis that have their elements electrically connected to the boom, using so-called "plumber's delight" construction. An antenna like a KT-34 will show little capacitance loading, because all elements are insulated from the boom. More on this below.

A 24-meter tall tower, loaded with a 5-element, 20-meter Yagi, will have an electrical length of 103° on 1.825 MHz. The effect of capacitance top loading depends to a great extent on the diameter of the tower under the capacitance hat. The capacitance hat (the Yagis) will have a greater influence with "slim" towers than with large-diameter towers. If you increase the tower diameter to 60 cm, this will shorten the electrical length between 4° and 7° (4° for the tower loaded with the 3-element, 20-meter Yagi and 7° for the tower loaded with the 40-meter, 3-element full-size Yagi). W. J. Schultz, K3OQF, published the mathematical derivation of the shunt-fed top-loaded vertical (Ref 7995).

There are neither data nor formulas available for calculating the exact electrical length of a tower loaded with multiple Yagis. The best way to find out is to attach a drop wire to the very top of the tower (turn it into a folded element) and grid dip the entire structure, as shown in **Fig 9-82** (see also Section 6.9.3). You could, of course, also model the entire structure, but that seems a rather tedious task.

If your tower, top loaded with a Yagi, is still a little short, you may want to add some extra wires from the top of the tower sloping down to increase the top loading. You might use part of the top set of guy wires, for example.

Towers with stacked Yagis are more difficult to assess. Basically it's the bottom Yagi that determines the capacitance, as this Yagi *hides* the Yagis above it, especially if they are nearby and smaller.

6.9.2. Yagis with Elements Insulated from the Boom

There are two problems associated to having Yagis with insulated elements on a tower for use on the lower bands:

- The insulated elements will only add little top loading.
- Possible arcing of the Yagi insulating parts and destruction of baluns.

Some Yagis use fiberglass or PVC for insulating their elements from the boom. While these are good enough for the Yagi, where the voltage between the center of the floating elements and the boom are low, voltages in case of a top-loaded tower may be very high in these same places. The highest RF voltage always occurs at the farthest end of an antenna from the feed point. For a shunt-fed tower with an HF Yagi with insulated elements used for top loading this highest RF voltage point would be at the ends of the Yagi's boom (eg, the 20-meter reflector). In most cases you can simply ground the center of the elements to the boom.

PA3DZN had to do this with his KT-34 tribander, and after having grounded all elements, he could not detect any change in performance of the Yagi, but his arcing problem was solved. John, K9DX, reported that he burned the feed line off his KLM 40-meter beam when it flashed over to the boom. Grounding all the elements solved the problem. Although direct grounding

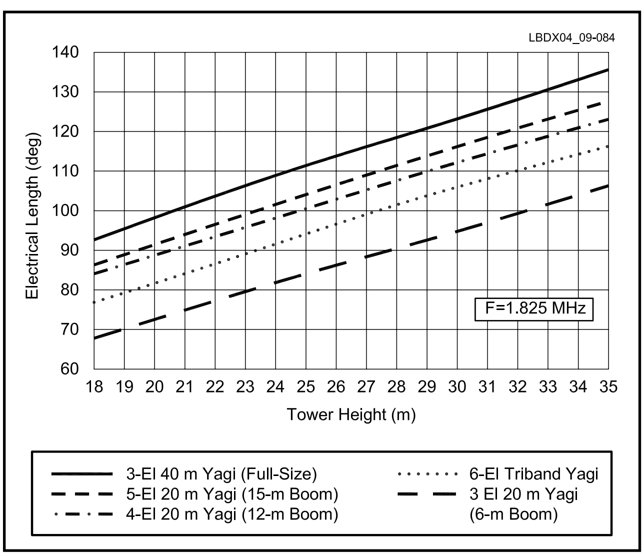


Fig 9-81 — Electrical length of a tower loaded with a Yagi antenna. The chart is valid for 160 meters (1.825 MHz) and a tower with an equivalent diameter of 30 cm. For a larger tower diameter, the electrical length will be shorter (4° to 7° for a tower measuring 60 cm in diameter).

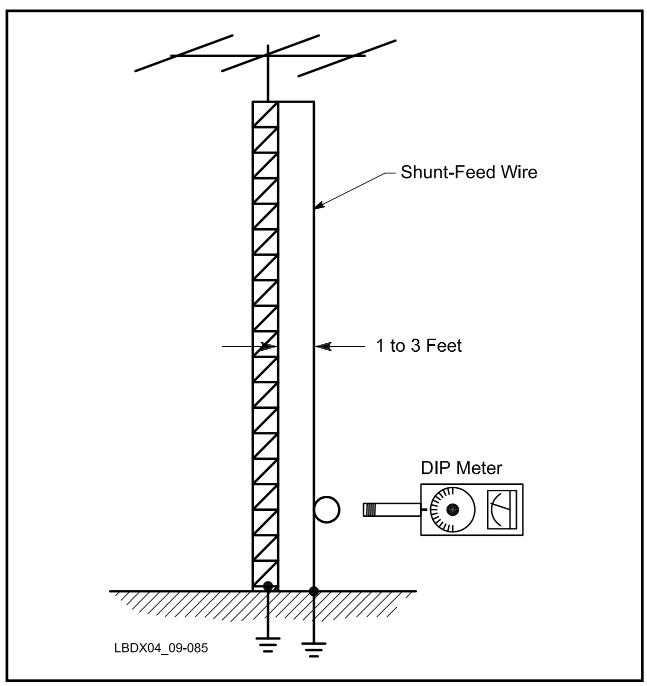


Fig 9-82 — A method of "dipping" a tower with a shunt-feed wire connected to the top.

of the elements should in most cases not upset the functioning of the Yagi, you could connect the elements to the boom using RF inductors having a few hundred ohms reactance on the lowest band the Yagi is used on.

You should be careful grounding the center of the driven element, because that might upset the matching system. Although insulated from the boom, the driven element usually acts as if it were grounded to the boom on 160 meters because of the feed line's coupling to the tower on its way down to the ground. Therefore, the driven element already fully adds to the top-loading and there is no reason to "ground" the driven element to the boom when you shunt feed the tower on 160 meters.

Certain types of baluns used on Yagis can be destroyed by shunt feeding the tower. If you use a balun using ferrite material (eg, W2DU, Hy-Gain, or Force-12) you may have to decouple 160-meter RF from reaching the balun, which is not easily done at a high-impedance point, near the end of the 160-meter antenna (see also Section 6.9.8). Plumber's-delight Yagis (all elements connected to the boom and using a gamma or omega match) are the ideal solutions for Yagi-loaded towers.

6.9.3. Measuring the Electrical Length

A second and very practical method of determining the resonant length of a tower system was given by DeMaw, W1FB (Ref 774). A shunt-feed wire is dropped from the top of the tower to ground level. What you want to do is turn a grounded single-conductor vertical into a folded-element vertical, where you now can easily do measurements in the drop wire. Attach a small 2-turn loop between the end of the wire and ground and couple this loop to the dip meter (Fig 9-82).

The lowest dip found then is the resonant frequency of the tower/beam. The electrical length at the design frequency is given by:

$$L \text{ (in degrees)} = 90° \times \frac{f_{design}}{f_{resonant}}$$

Therefore, if $f_{resonant} = 1.6$ MHz and $f_{design} = 1.8$ MHz, then L = 101°.

6.9.4. Gamma and Omega Matching

There are many approaches to matching a loaded, grounded tower. Three popular methods are:
- Slant-wire shunt feeding (Section 6.9.5)
- Folded-monopole feeding (Section 6.9.6)
- Gamma or omega-match shunt feeding.

Gamma and omega-matching techniques are widely used on loaded towers. The design of gamma matches has often been described in the literature (Ref 1401, 1414, 1421, 1426 and 1441).

Fig 9-83 shows the height of the gamma-match tap, as well as the value of the gamma capacitor for a range of antenna lengths varying from 60° to 180°. The chart was developed using a gamma wire of 10-mm diameter. There are three sets of graphs, for three different wire spacings (0.5, 1.0 and 1.5 meters).

It is a fairly common misconception to think that the tower must be resonant to be able to match it correctly to 50 Ω at the desired frequency. This isn't true. However, there are some advantages to having the tower (with its top loading) resonant near the desired frequency:

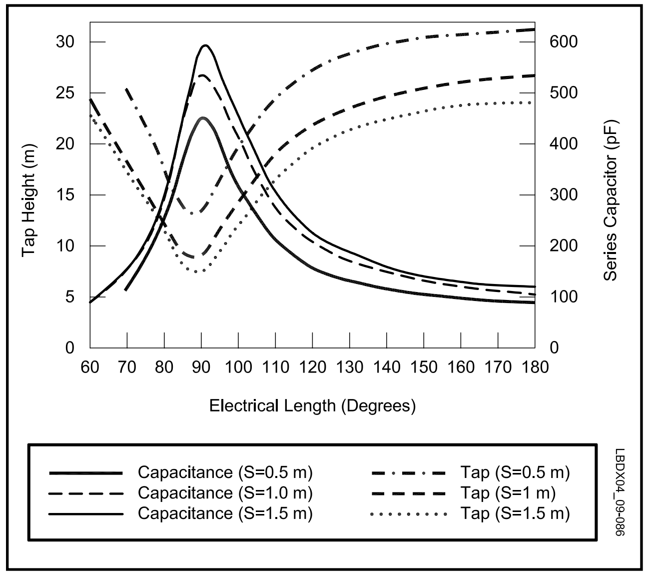


Fig 9-83 — Tap height and values of the gamma series capacitor for a shunt-fed tower at 1.835 MHz. The tower diameter is 250 mm, and the gamma wire has a diameter of 10 mm. Three sets of curves are shown, for three spacings (S). The spacing is the distance from the wire to the tower center.

- A resonant tower makes it possible to design an efficient shunt-feed system.
- A tower resonant slightly off the desired frequency can exhibit a broader SWR bandwidth because there are now two dips in the SWR curve, one caused by the resonant frequency of the tower and another one caused by the pulling of the gamma or omega match of the resonance to a slightly different frequency.
- The 50-Ω tap point is closest to the ground with a λ/4 resonant tower (only about 8 meters high on 160 meters).
- The 2:1 SWR bandwidth is better on tower which is near resonance than for a tower way off resonance. (see also Section 6.9.4.3). That's because the rate of change in the imaginary part of the impedance is smallest around resonance (and greatest at anti-resonance).
- The RF voltage across the gamma capacitor will be lowest.

However, it is not necessary to strive for λ/4-wave resonance. Virtually any size vertical can be successfully shunt-fed and will perform well.

6.9.4.1. Close Spacing Versus Wide Gamma Spacing

The wider the spacing, the shorter the gamma wire needs to be. Shorter gamma wires will naturally show less inductive reactance, which means that the series capacitor must be larger in value.

Electrically very long verticals will require a tap that is 20 to 30 meters up on the tower. The required series capacitor will be small in value (typically 100 to 150 pF). There will be a very high voltage across capacitors of such small value.

In case the required gamma-wire length shown in Fig 9-83 appears to be longer than the physical length of the tower, you will need an omega match.

6.9.4.2. Influence of Gamma-Wire Diameter

The gamma-wire diameter has little influence on the length of the gamma wire (position of the tap on the tower). A larger diameter wire will require a somewhat shorter gamma wire. The wire diameter has a pronounced influence, however, on the required gamma capacitor. It also has some influence on the SWR bandwidth of the antenna system, but less than most believe.

6.9.4.3. SWR Bandwidth

Table 9-16 and **Table 9-17** show the feed-point impedance and the SWR versus frequency for a vertical of 100° electrical length, fed with a gamma match. A spacing of 50 cm is used in Table 9-16, and 150-cm spacing in Table 9-17. Wire diameters of 2 mm (AWG #12), 10 mm AWG 000), 50 mm and 250 mm are included. The 2-mm (#12) wire is certainly not responsible for a narrow bandwidth. It does not seem worth using a "wire cage" gamma-wire to improve the bandwidth.

For loaded towers that are much longer than 100°, the bandwidth behavior is quite different. The longer the electrical length of the vertical, the narrower the SWR bandwidth.

Table 9-18 shows the feed-point impedance and the SWR

Table 9-16

Gamma-Match Data for a Shunt-Fed Tower with 50-cm Gamma-Wire Spacing

Tower electrical height = 100 degrees
Tower diameter = 250 mm (10 inches)

	1.730	*1.765*	*1.800*	*1.835*	*1.870*	*1.905*	*1.940 MHz*
Gamma-wire diameter = 2 mm (AWG 12); tap height = 19.5 m (64.0 ft)							
R	80.6	66.8	56.5	50.0	43.3	39.2	36.0
X	+330	+338	+350	+363	+377	+392	+407
SWR	2.0	1.7	1.3	1.0	1.4	2.0	2.8
Gamma-wire diameter = 10 mm (0.4 in.); tap height = 19.8 m (65.0 ft)							
R	82.9	68.6	58	50	44.8	40.6	37.6
X	+250	+257	+267	+278	+291	+303	+316
SWR	1.9	1.6	1.3	1.0	1.3	1.8	2.4
Gamma-wire diameter = 50 mm (2 in.); tap height =20.0 m (65.6 ft)							
R	80.8	66.9	56.9	50	44.3	40.3	37.3
X	+164	+171	+179	+188	+198	+208	+218
SWR	1.8	1.5	1.2	1.0	1.3	1.6	2.1
Gamma-wire diameter = 250 mm (10 in.); tap height =20.2 m (66.3 ft)							
R	78.8	65.5	56	50	44	41	38.3
X	+75	+82	+90	+98	+105	+113	+121
SWR	1.8	1.5	1.2	1.0	1.2	1.5	1.8

Table 9-17

Gamma-Match Data for a Shunt-Fed Tower with 150-cm Gamma-Wire Spacing

Tower electrical height = 100 degrees
Tower diameter = 250 mm (10 inches)

	1.730	*1.765*	*1.800*	*1.835*	*1.870*	*1.905*	*1.940 MHz*
Gamma-wire diameter = 2 mm (AWG 12); tap height = 11.9 m (39.0 ft)							
R	86.8	71	59	50	43.7	38.8	35
X	+226	+229	+236	+244	+253	+262	+272
SWR	1.8	1.5	1.2	1.0	1.2	1.6	2.1
Gamma-wire diameter = 10 mm (0.4 in.); tap height = 12.0 m (39.4 ft)							
R	87.8	71.7	59.8	50	44.4	39.5	25.7
X	+179	+181	+187	+195	+203	+212	+220
SWR	1.8	1.5	1.3	1.0	1.2	1.6	2.0
Gamma-wire diameter = 50 mm (2 in.); tap height =12.0 m (39.4 ft)							
R	87	71	59	50	44.3	39.5	35.8
X	+120	+132	+137	+144	+152	+159	+166
SWR	1.8	1.5	1.2	1.0	1.2	1.5	1.8
Gamma-wire diameter = 250 mm (10 in.); tap height =11.9 m (39.0 ft)							
R	85.2	69.6	58.3	50	44	39.4	36
X	+79	+82	+87	+93	+100	+1-6	+112
SWR	1.8	1.5	1.2	1.0	1.2	1.5	1.7

Table 9-18
Gamma-Match Data for a Shunt-Fed Tower with 150-cm Gamma-Wire Spacing

Tower electrical height = 150 degrees
Tower diameter = 250 mm (10 inches)

	1.730	*1.765*	*1.800*	*1.835*	*1.870*	*1.905*	*1.940 MHz*
Gamma-wire diameter =10 mm (0.4 in.); tap height = 25.9 m (65.0 ft)							
R	43.1	45.2	47.7	50	54	58	62.7
X	+567	+597	+628	+661	+697	+736	+778
SWR	6.0	3.5	1.9	1.0	2.0	3.7	6.3
Gamma-wire diameter = 250 mm (10 in.); tap height = 24.8 m (81.4 ft)							
R	41.5	43.8	46.6	50	54	58.5	64
X	+286	+303	+320	+340	+362	+384	+409
SWR	3.2	2.2	1.5	1.0	1.5	2.2	3.2

for a vertical of 150° electrical length, fed with a gamma-match and a gamma-wire of both 10 mm and 250 mm OD. In contrast with the effect on the shorter vertical (100°), the wire diameter now has a pronounced influence on the bandwidth. The 10-mm wire yields a 70-kHz bandwidth; the 250-mm wire cage almost 130 kHz. As can be seen from the impedance values listed in Table 9-17, it is the large variation in reactance that is responsible for the steep SWR response. This can be overcome using a motor-driven variable capacitor. The 150°-long antenna with a 10-mm-OD gamma wire shows an SWR of less than 1.3:1 over more than 200 kHz, if a variable capacitor with a tuning range of 100 to 175 pF is used. A high-voltage (eg, 10 kV) vacuum variable is a must.

This simple way of obtaining a very flat SWR does not apply to shorter verticals (90° to 110°), where a much larger variation in the resistive part of the feed-point impedance is responsible for the SWR. **Fig 9-84** shows SWR plots for gamma-fed towers of varying electrical length, using a 10-mm OD gamma wire, spaced 150 cm from the tower.

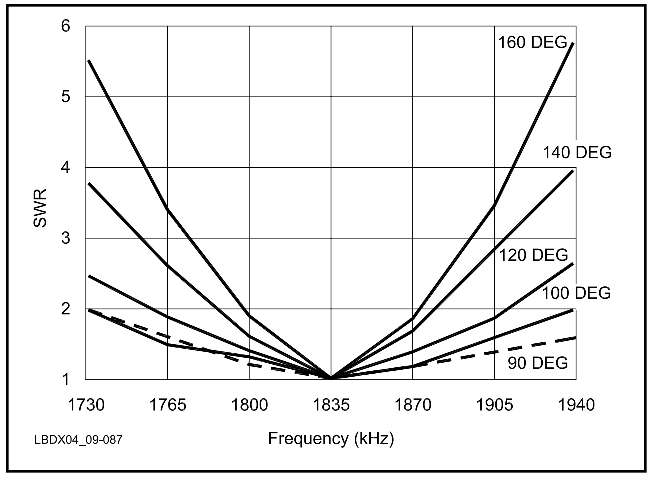


Fig 9-84 — SWR curves for gamma-fed towers using a 10-mm OD gamma wire and a spacing of 50 cm, for electrical tower lengths varying from 90° to 160°. The SWR of the longer vertical can be tuned over a wide bandwidth using a motor-driven variable series capacitor.

6.9.4.4. Adjusting the Gamma-Matching System

The easiest way to fine tune the gamma-matching system is to vary the spacing of the gamma wire. This changes the resistive part of the feed-point impedance. Then you can tune out the inductive reactance using the series gamma capacitor for a 1:1 SWR.

Example

For a vertical with an electrical length of 100° and with a tower diameter of 250 mm, we install the tap at a height of 14 meters. At that point the spacing is 1 meter. Changing the spacing at ground level has the following influence:
Spacing = 0.5 meter: Z = 38 + j 206 Ω
Spacing = 0.75 meter: Z = 44.8 + j 298 Ω
Spacing = 1.0 meter: Z = 49.3 + j 211 Ω
Spacing = 1.25 meters: Z = 56.4 + j 213 Ω
Spacing = 1.5 meters: Z = 61.5 + j 214 Ω

This demonstrates how fine tuning can easily be done on the gamma-matching system.

6.9.4.5. Using the Omega Matching System

If you can tune your tower using a gamma, I would not advise using an omega system. The omega match requires one more component, which means additional losses and additional chances for a component breakdown. It is possible, however, to use a gamma-rod (wire) length that is up to 50% shorter than the length shown in Fig 9-83 when you use an omega match. An omega system is similar to a gamma system except that a parallel capacitor is connected between the bottom end of the gamma wire and ground.

The 100°-long vertical requires a 14-meter long gamma wire, with 100-cm gamma-wire (OD 10 mm) spacing. If we shorten the gamma wire to 8 meters, the transformed impedance becomes 14.1 + j 127 Ω. This can be matched to 50 Ω using an L-network. One of the solutions of this L network consists of two capacitors: the well-known parallel and series capacitor of the omega-matching system. To calculate the omega system, use the following procedure:

- Model the vertical with the shorter gamma rod. Make sure you use enough segments (pulses). For 160 meters, segment lengths of 100 cm give good results. Note the input impedance, which will be lower than 50 Ω and inductive.
- Use the L-Network module of the *New Low Band Software* to calculate the capacitance of the parallel and the series capacitor.

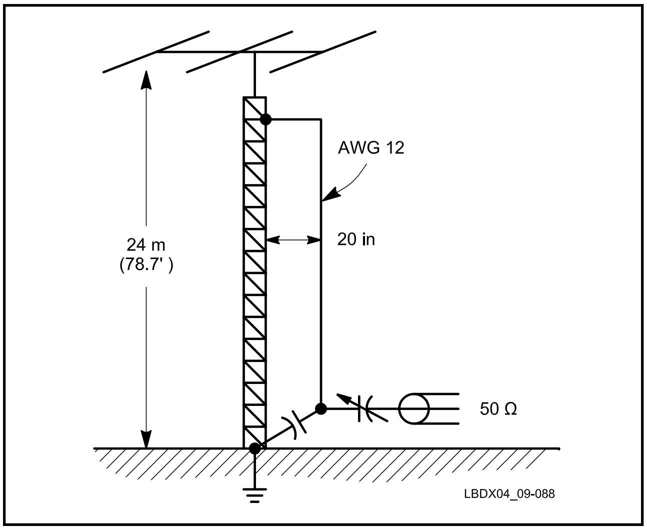


Fig 9-85 — A shunt-fed tower using an omega matching system. The tower is electrically 140° long. An omega match is required, since the tower is physically too short to accommodate a gamma match with a 2-mm gamma wire. Table 9-19 lists the impedances at the end of the gamma wire before and after transformation by the capacitors of the omega-match system.

In our example above, the 8-meter-long gamma wire requires a parallel capacitor of 369 pF and a series capacitor of 323 pF.

If you have a physically short tower with a lot of loading, it may be that the required tap height is greater than your tower height. In this case an omega match is the only solution (if you have already tried a larger spacing).

Example

See **Fig 9-85**. The tower was measured with a dip meter, and the electrical length turned out to be 140°. The physical height is 24 meters. Fig 9-83 shows a required gamma-wire length of 30 meters for a 2-mm OD gamma wire and a 50-cm spacing. In this case we will connect the gamma wire at the top of the tower (h = 24 meters). Using *NEC-2*, we calculate the feed-point impedance as $Z = 17.2 + j\ 579\ \Omega$. From the L-Network software module, the capacitor values are calculated as C_{par} = 62 pF, C_{series} = 88 pF. Note that these very low-value capacitors will carry very high voltages across their terminals with high power.

This L-network is a high-Q network. **Table 9-19** lists the impedances at the end of the gamma wire before and after transformation by the capacitors of the omega-match. Note the very narrow bandwidth of this high-Q matching system. If we adjust the omega-capacitors for a 1:1 SWR on 1835 kHz, the 2:1 bandwidth will be typically only 20 kHz. If we make the series capacitor adjustable (60 to 120 pF), we can tune the antenna to an SWR of less than 1.5:1 over more than 200 kHz.

6.9.4.6. Conclusion

If after modeling or actually measuring your loaded tower it turns out to be longer than about 140°, you might consider using an elevated feed system, with lots of elevated radials (see Section 2.2.10). Within reason, the longer the section below the feed point, the easier to decouple this section. If you keep raising the base of the vertical, however, you will achieve a vertical radiation pattern that is not ideal for most DX work.

If you have an electrically long gamma-matched vertical (more than about 110° high), you can use a large-diameter cage-type gamma wire and a large wire-to-tower spacing. Making the series capacitor remotely tunable will certainly make the antenna much more broadbanded. Do not shorten the gamma wire unless required because of the physical length of the tower.

Fig 9-86 shows the correct wiring of both the gamma and omega-matching networks on a loaded tower. Notice the correct connection of the shunt capacitor in the case of the omega match. Make sure the ground wires all have very low resistance and the lowest possible reactance. Use flat solid-copper strip

Table 9-19
Omega-Match Data for a Shunt-Fed Tower with 50-cm Gamma-Wire Spacing

Tower electrical height = 140 degrees
Tower diameter = 250 mm (10 in.)

	1.730	***1.765***	***1.800***	***1.835***	***1.870***	***1.905***	***1.940 MHz***
Gamma-wire diameter = 2 mm (AWG 12); tap height = 24.0 m (78.7 ft)							
R	16.0	16.3	16.7	17.2	17.9	18.6	19.3
X	+514	+535	+552	+579	+603	+629	+650
With parallel capacitor of 62 pF added							
R	37.4	40.7	44.8	50	56.8	65.3	74.7
X	+785	+845	+910	+986	+1073	+1178	+1287
With fixed series capacitor of 88 pF added							
R	37.4	40.7	44.8	50	56.8	65.3	74.7
X	-261	-180	-95	0	+106	+228	+348
SWR	38.0	17.9	5.9	1.0	5.8	17.9	35
With variable series capacitor, 50 to 125 pF (adjusted to cancel inductive reactance)							
R	37.4	40.7	44.8	50	56.8	65.3	74.7
X	0	0	0	0	0	0	0
SWR	1.3	1.2	1.1	1.0	1.1	1.3	1.5

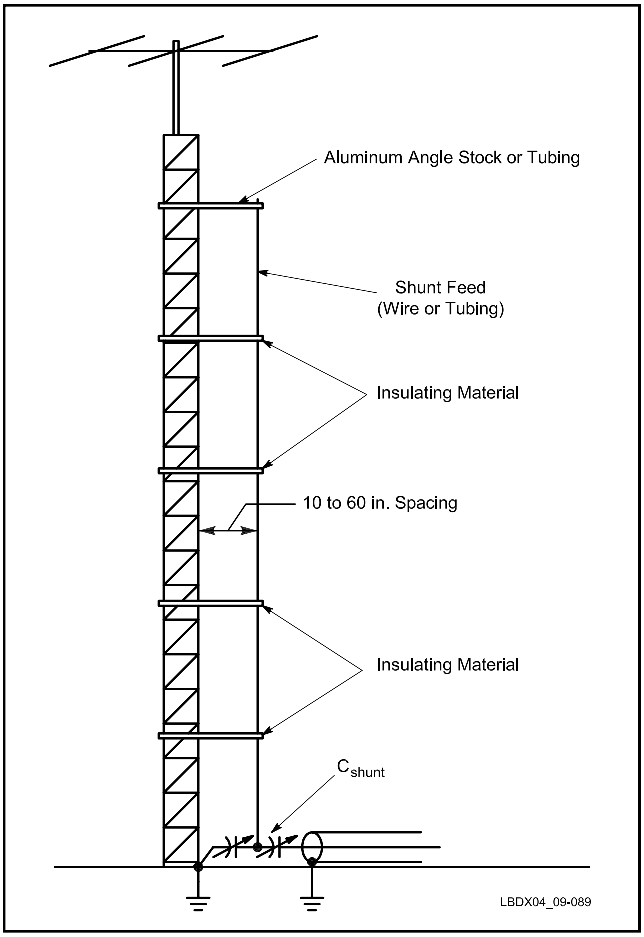


Fig 9-86 — The omega-matching system (a gamma match with an additional shunt capacitor) adds a great deal of flexibility to the shunt-fed-tower arrangement. To maintain maximum bandwidth make the gamma wire as long as possible. If the antenna is electrically longer than 120°, a variable series capacitor will make it possible to obtain a very low SWR over a wide bandwidth.

if possible. Do *not* use flat braided strip, which has high RF losses, contrary to popular belief. Use flat plain copper strip.

The same principles can, of course, be applied to 80 meters, although it is probable that a tower of reasonable height, loaded with a Yagi antenna, will result in too long an antenna for operation on that band.

6.9.4.7. A Few Practical Hints

All cables leading to the tower and up to the rotator and antennas should be firmly secured to a tower leg, on the inside of the tower. All leads from the shack to the tower base should be buried underground to provide sufficient RF decoupling. If there is RF on some of the cables, you should coil up a length of the cable running up your shunt-fed or series-fed tower to make a common-mode choke at the tower base. A coil consisting of 50 turns on a 10 cm diameter form yields about 100 µH of inductance. If you still detect some RF on these cables entering your shack, install another coil at that point or put some ferrite beads on the cable. Coaxial cables running down the tower should preferably be grounded right at the base of the vertical.

Take care to ensure good electrical continuity between the tower sections, and between the rotator, the mast and the tower. Again, use flat-strip copper conductors, not the woven battery-connecting flat strips, which are good only for dc.

A gamma rod can be supported with sections of plastic pipe, attached to the tower with U bolts or stainless-steel radiator hose clamps. If the tower is a crank-up type, heavy, insulated copper wire can be used for the gamma element.

6.9.5. The Slant-Wire Feed System

The slant-wire feed system is very similar to a gamma feed system. The feed wire is attached at a certain height on the tower and slopes at an angle to the ground, where a series capacitor tunes out the reactance. The advantage of this system is that a match can be obtained with a lower tap point, which makes it possible to avoid using an omega match on physically short towers. The disadvantage is that the slant-wire feed also radiates a horizontally polarized component. The slant-wire feed system can easily be modeled using a computer program, just like the gamma and omega-matching systems.

6.9.6. Folded Monopoles

Folded antennas have the following advantages:

- Higher bandwidth due to a larger effective antenna diameter.
- Higher feed-point impedance.

Folded monopoles achieve a higher Z_{feed} but do *not* raise the R_{rad} of the antenna. You will not improve the radiation efficiency by using a folded monopole configuration.

Fig 9-87 shows how you can manipulate the wire diameter and spacing to obtain up-transformation ratios ranging from two to well over 10. (Source: *Kurze Antennen*, by Gerd Janzen, ISBN 3-440-05469-1). The configuration of the two-wire folded monopole is shown in the same figure. One leg is grounded, while the antenna is fed between the bottom of the other leg and ground.

The effective diameter of multi-wire elements can be calculated from the chart shown in Fig 9-50. Three-wire folded-

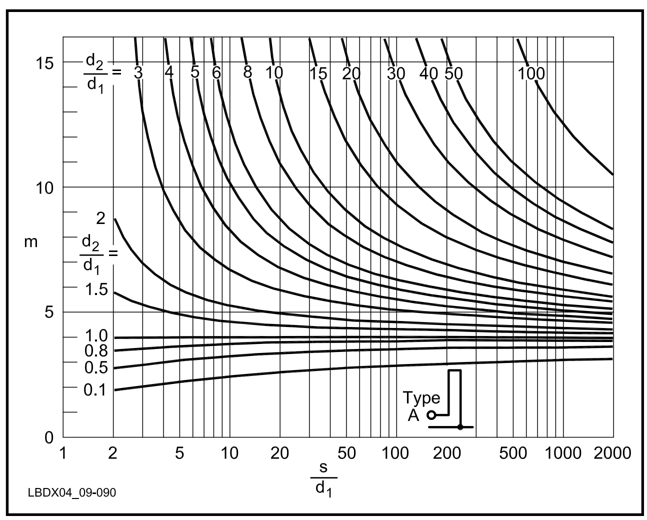


Fig 9-87 — Transformation ratio (m) of a two-wire folded monopole, as a function of the wire spacing (S/d1) and the ratio of the conductor diameter (d2 = diameter of the grounded conductor, d1 = diameter of the fed conductor). (After Gerd Janzen, *Kurze Antennen*.)

element configurations allow even higher transformation ratios, as can be seen in **Fig 9-88**. The effective antenna diameter (which determines the bandwidth of the antenna) is given in Fig 9-64 for the various configurations. The configurations are also shown in the same figure.

6.9.7. Modeling Shunt-Fed Towers

MININEC or *NEC-2* can be used for modeling the gamma, omega and slant-wire matching systems on shunt-fed grounded towers. Satisfactory results are obtained using the following guidelines:

- The horizontal wire connecting the gamma wire to the tower has one segment (the length of the segment is the spacing from the gamma wire to the tower).
- All segments should have the same length; this is determined by the length of the horizontal wire connecting the gamma-wire to the tower.
- Do not try to model the capacitance top load. It is much easier to first grid-dip the tower (see Fig 9-82), calculate the electrical length of the loaded tower and then use an equivalent straight tower to do the gamma-match modeling.

Example:

A tower dips at 1.42 MHz. The required operating frequency is 1.835 MHz. The electrical length is:

$$\text{L (in degrees)} = 90° \frac{1.835}{1.42} = 116°$$

The physical length of a λ/4 tower (equivalent diameter = 250 mm) is 39 meters. The equivalent tower length for 116° is:

$$39\,\text{m} \times \frac{116°}{90°} = 50.3 \text{ meters}$$

Now model a vertical with a diameter of 250 mm and of 50.3 meters length. According to Fig 9-83, the tap will be at a height of between 17 and 25 meters, depending on the wire spacing.

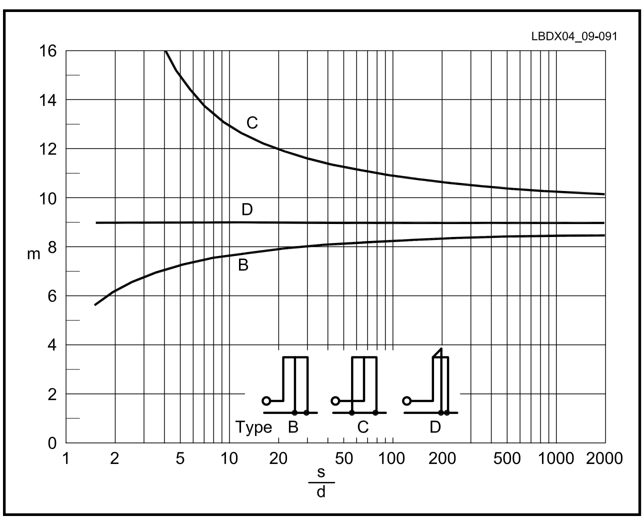


Fig 9-88 — Transformation ratio (m) of a three-wire folded monopole, as function of the spacing between the wire (S/d) and the configuration B, C or D). In this case three conductors of equal diameter are assumed. (After Gerd Janzen, *Kurze Antennen*.)

6.9.8. Decoupling Antennas at High-Impedance Points

When shunt feeding a tower that supports various antennas, including wire antennas, you may wish to decouple the wire antennas from the vertical radiator. Otherwise, the wire antennas will act as top loading to the vertical. For relatively short towers, the extra loading may be welcome, but in other cases the loaded vertical may become too long with the additional loading of large wire antennas, such as a 160-meter or an 80-meter inverted V.

These wire antennas are usually installed at the top of the tower, at a high-impedance point. This makes decoupling of the wire antennas more difficult. Conventional common-mode current baluns are not suitable, since they do not have enough inductance to effectively decouple the antenna. Such baluns can lead to unexpected changes in the feed-point impedance of the loaded vertical while transmitting. When first transmitting the SWR may be normal, but soon the ferrite material used in the current balun will heat to the point where the Curie temperature is reached, resulting in a sudden drop in magnetic susceptibility of the ferrite material. The balun will no longer represent enough impedance, causing the dipole to load the tower with a change in SWR as a result.

For effective decoupling in this application, a high-impedance balun is required. This balun is rather like a trap — a parallel-tuned resonant circuit — tuned to the frequency of the loaded vertical. The RF currents that flow from the vertical to the dipole (which we want to decouple) are common-mode currents, which means they flow only on the outside of the coaxial feed line of the inverted V dipole.

The trap is made by winding a single-layer coil of coax onto a suitable form, and resonating the coil with a suitable capacitor. Jim Jorgenson, K9RJ, made such a trap for 160 meters. It consisted of 21 turns of RG-213, wound close-spaced on a PVC pipe 10 cm in diameter. The coil is about 33 cm long, and the measured inductance of this coil is 33 µH. The coil is held in place on the form by drilling close-fitting holes at an angle through the PVC pipe and passing the coax through these holes into the interior of the pipe at both ends. At one end the shield and the inner connector are separated and connected to stainless steel eyebolts that are used to connect the two legs of the inverted V antenna. At the other (bottom) end the coax passes out through a standard PVC end cap and a PL-259 connector is attached at that end.

The capacitance needed to resonate this coil on 1830 kHz is about 200 pF. You could use a quality transmitting-type ceramic capacitor, but a suitable capacitor can be made from a short piece of coax. RG-213 coax has a capacitance of 100 pF/m, which means that an open-ended piece of RG-213, 2 meters long, will resonate the coil on 160 meters. The resonant frequency of the trap can easily be measured using a grid-dip meter. **Fig 9-89** shows the layout of the trap and **Fig 9-90** shows it deployed on the tower. To tune the trap, you can deliberately make the coaxial-line capacitor too long, and then cut small pieces at a time until resonance is obtained at the desired frequency. The stub capacitor can then be folded inside the PVC tube before putting on the bottom end cap. Jim reports that since he has been using this balun there

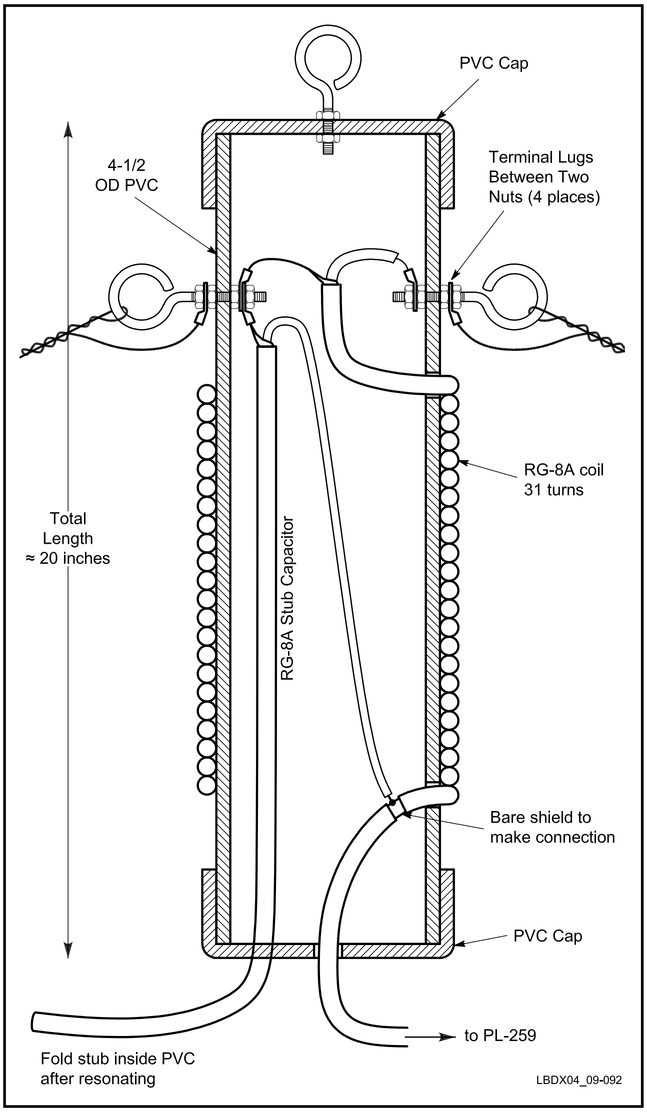


Fig 9-89 — Construction details of the K9RJ decoupling trap.

Fig 9-90 — The K9RJ coaxial trap feeds a 160-meter inverted V hanging from the top of a tower, which is also used as a shunt-fed vertical for 160 meters. The entire construction looks somewhat like an oversized center insulator or balun.

has been no change of the shunt-fed tower impedance, with the 160-meter inverted V attached. This proves that the trap is now fully decoupling the inverted V from the vertical.

Whether or not this concept is good enough to decouple the inverted V-antenna (in this case) from the tower depend on where along the tower this system is installed. If it is installed right at the top of the tower with no additional top loading, the impedance at that point is very high, meaning extremely high voltages at that point. This system worked for a 160-meter inverted-V mounted at the 80-foot level on K9RJ's 100-foot shunt-fed and top-loaded tower. It probably would work well enough for most typical shunt-fed and top loaded tower installations on 160 meters, but not when the antenna is installed right at the top of an unloaded tower.

7. INVERTED-L ANTENNAS

The ever-so popular inverted-L is analyzed in this section and a few practical designs, such as the well-known "AKI Special," are given particular attention.

The inverted-L is a popular antenna, especially on 160 meters. These antennas are not truly verticals, as part of the antenna is horizontal and thus radiates a horizontally polarized component. We often form the wrong mental picture of what actually happens because most antenna modeling programs only express the field in two distinct polarizations. We wrongly picture two distinct fields. The actual field is the vector sum of the two fields, and is a single polarization wave with a tilt and a distinct total null at 90-degrees from the peak response.

Most inverted-Ls are of the $\lambda/4$ (total electrical length) variety, although this does not necessarily need to be the case. The vertical portion of an inverted-L can be put up alongside a tower supporting HF antennas. In such a setup one must take care that the tower plus HF antenna does not resonate near the design frequency of the inverted-L. Grid dip your supporting tower using the method shown in Fig 9-82. If it dips anywhere near the operating frequency, maybe you should shunt feed the tower instead of using it as a support for an inverted-L.

If you nevertheless choose to use the inverted-L, you must detune the tower to make sure the highest possible current flows in the parallel-tuned structure (see Section 3.10 in Chapter 7).

The longer the vertical part of the antenna, the better the low-angle radiation characteristics of the antenna and the

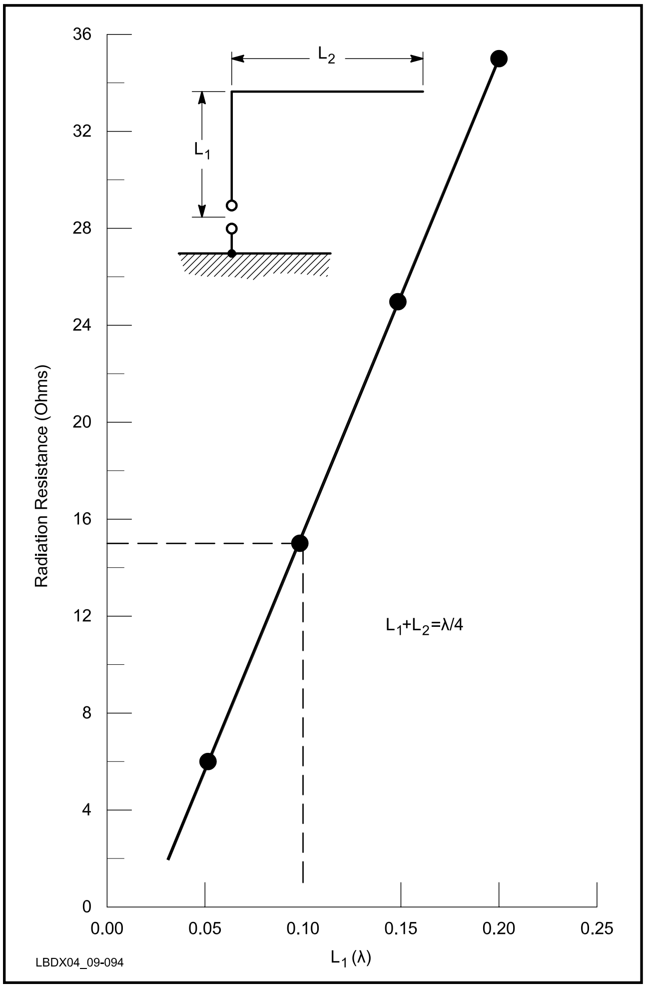


Fig 9-91 — Radiation resistance of an inverted-L antenna as a function of the lengths of the horizontal wire versus the vertical conductor size.

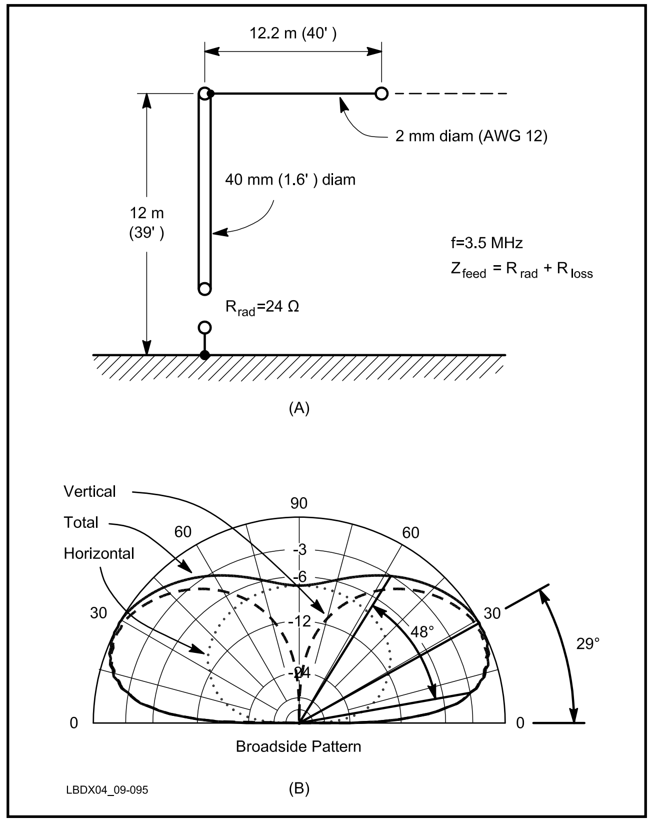


Fig 9-92 — At A, a 3.5-MHz inverted L with a 12-meter vertical mast. The vertical radiation pattern is shown at B. The pattern has both vertically and horizontally polarized components and these components are also plotted at B. The pattern is generated over average ground, using 60 λ/4 radials. Note that the angle of maximum radiation is 29°, not bad for a DX antenna.

higher the radiation resistance (see **Fig 9-91**). The horizontal part of the antenna accounts for the high-angle radiation that the antenna produces but this normally is low, since the bulk of the radiation comes from the bottom part of the antenna, where the current is highest. Since it is a top-loaded monopole, an inverted-L requires a good ground system, just like any vertical (more so the short loaded verticals).

Fig 9-92 shows the vertical and horizontal radiation patterns for a practical design of an inverted-L antenna for 3.5 MHz, one having a 12-meter long vertical mast. Notice how the vertical part of the antenna takes care of the low-angle radiation, while the horizontal part gives high-angle output. The radiation pattern shown is for the direction perpendicular to the plane of the inverted-L.

An inverted-L is also an attractive solution for the operator who wants to turn his 80-meter vertical antenna into a 160-meter antenna (**Fig 9-93A**).

If you want to make it a two-band antenna, the easiest solution is to insert a trap at the top of the 80-meter vertical (Ref 659). The exact L/C ratio is not important, but it influences the length of the loading wire and the SWR behavior of the antenna on both 80 and 160 meters (Fig 9-93B). Details on the trap can be found in Section 6.4.

The inverted-L has been extensively described in amateur literature as a good antenna for producing a low-angle signal on Top Band (Ref 798 and 799). The Battle Creek Special, described in Section 6.7 is an example of an inverted-L (on 80 and 160 meters).

Sloping Loading Wire

The large majority of inverted-L antennas employ a sloping loading wire. The effect is that the radiation resistance of the antenna will be much lower and the loading wire slightly longer to achieve the same resonant frequency. An 18-meter vertical with a 25-meter-long horizontal loading wire yield R_{rad} = 17.5 Ω. The same vertical with a sloping loading wire (25.8 meters long, tip 3 meters above ground) yields a R_{rad} of only 9 Ω.

Conclusion: keep the loading wire as horizontal as possible (see also Section 3.7).

Making the Inverted-L Longer than λ/4

We have seen people making the loading wire longer than necessary to obtain resonance with the idea of raising R_{rad} and making a direct feed to a 50-Ω feed line possible. In this case you will have to make the loading wire so long that the majority of the radiation will be horizontally polarized and

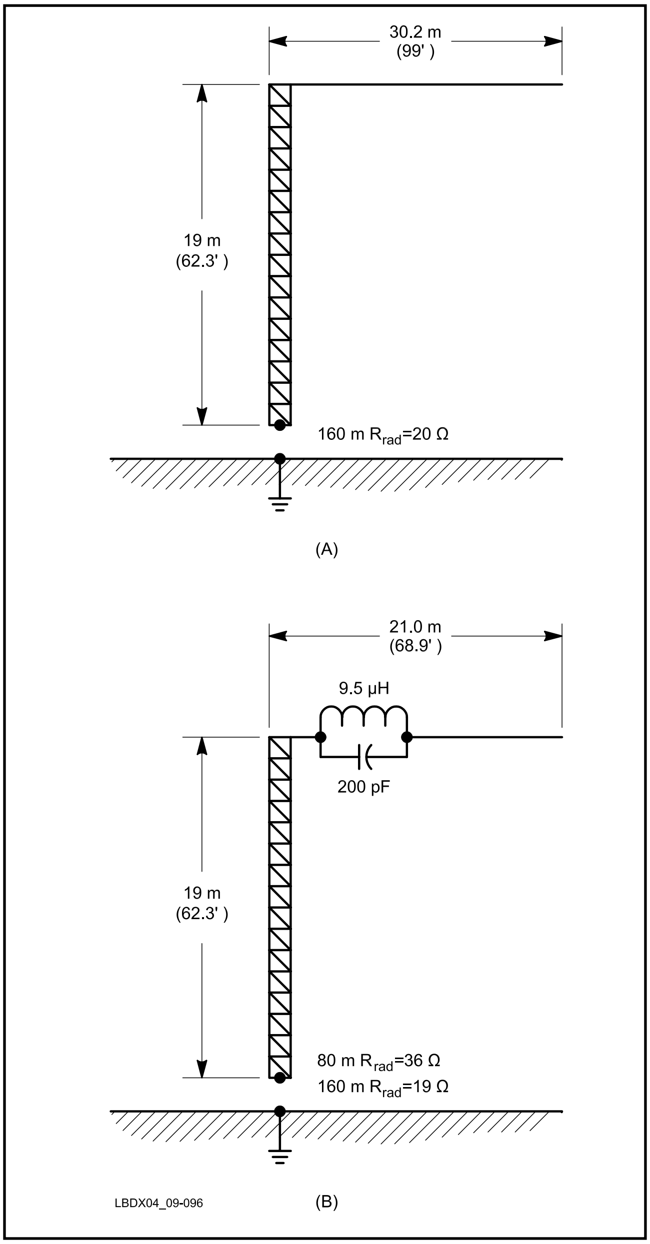


Fig 9-93 — At A, an inverted-L antenna for 160 meters, using a 19-meter vertical tower. To cover both 80 and 160 meters, a trap can be installed at the top of the tower as shown at B. With the trap installed, the loading wire is shorter, because the trap shows a positive reactance (loading effect) on 160 meters.

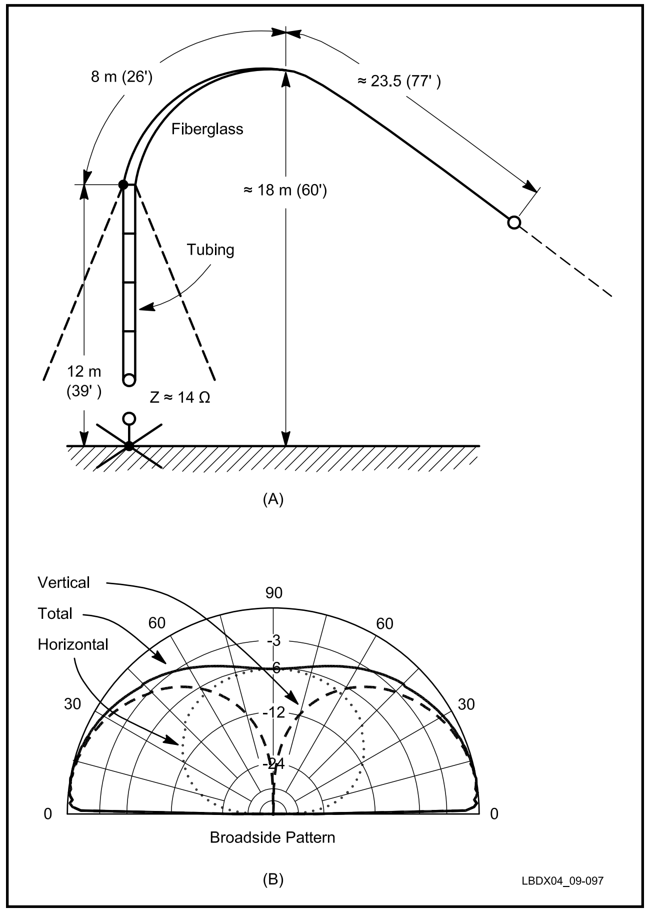


Fig 9-94 — The "AKI Special," a typical DXpedition type 160-meter inverted L. A collapsible fiberglass fishing rod (available in Europe in lengths of up to 12 meters) is used on top of a 12-meter aluminum mast. A 2 mm OD wire is attached to the rod, and slopes to a distant point to make the sloping (horizontal) part of the antenna. The radiation pattern is over saltwater. (That's where the island DXpeditioners put these antennas.)

the radiation angle will be straight up — in other words a very poor DX antenna.

The Inverted L, a Poor Man's T-Loaded Vertical

If at all possible, use a symmetrical top hat loading system, and keep it as horizontal as possible. Inverted Ls radiate considerable high angle energy depending on the length of the vertical. If the vertical is only 18 meters high, the horizontally polarized high angle radiation will only be approximately 6 dB down from the vertically polarized component! Because of the current distribution in both arms of the loading device, a symmetrical capacitive loading system (such as a T hat) will not radiate any horizontally polarized component in the far field. Compared to a top loaded vertical, the inverted-L is what the inverted V is compared to a dipole.

7.1. The AKI Special

The AKI Special is another DXpedition-style inverted-L, as used by Aki Nago, JA5DQH, during his operations on 160 meters from several rare DX spots. From Kingman Reef (May 1988), Nago used the inverted-L shown in **Fig 9-94**.

The vertical part is made of a 12-meter long aluminum mast, which is extended by an 8-meter long fiberglass fishing rod, to which a copper wire has been attached. From the tip of the (bent) fishing rod, the sloping wire extends another 23.5 meters, to be terminated with a fishing line supported by a 3-meter long pole at some distance.

Aki used about 800 meters of radials running into the Pacific Ocean. He used a very similar 160-meter antenna successfully from Palmyra during the same DXpedition trip in 1988, and during a more recent DXpedition to Ogasawara by JA5AUC. The excellent signals from VKØIR (1996) on 80-meter SSB were also produced with an AKI-type inverted-L, using two elevated radials, above a large number of ground wires (not connected to the radials or the feed system). The calculated radiation resistance of this antenna is approximately 14 Ω. The main radiation angle (over seawater) is 10°, but due to the relatively long horizontal (sloping) wire, the radiation at higher angles is only slightly suppressed.

7.1.1. Tuning Procedure

When cutting the length of the sloping wire, cut it at first 2 meters too long. Put up the antenna, and connect one of the popular antenna analyzers between the bottom of the antenna and the ground system. Adjust the length of the sloping top wire for minimum SWR. Now read the resistance value off the scale of your analyzer. If it is between 35 Ω and 70 Ω, the SWR will be pretty acceptable (1.5:1) and you may want to feed the antenna directly with 50-Ω feed line. From the difference between the R value and the calculated 14-Ω radiation resistance, you can calculate the effective ground-loss resistance of the ground radial system.

If, at resonance, the feed-point impedance is above 50 Ω, you really need to improve the radial system. At 50 Ω the efficiency would be 14/50 = 28%. Any value higher than 50 Ω indicates an even lower efficiency. If you want a perfect match you can use an L network or an unun (Ref 1522).

8. THE T ANTENNA

The current-fed T antenna is a top-loaded short vertical, as covered earlier in this chapter. The voltage-fed T antenna is given special attention here, as well as different top-loading structures.

8.1. Current-Fed T Antennas

T-wire loading (flat-top wire) is covered in detail in Section 3.6.2.3. when dealing with top loading of short verticals. The advantage of the horizontal T-wire loading system over the inverted-L system is that the top-wire does not contribute to the total radiation pattern. **Fig 9-95** shows a practical current-fed T design, where a 12-meter-long vertical is loaded with a horizontal top-load wire to achieve resonance at 3.5 MHz. The R_{rad} of this design is approximately 23.5 Ω.

Fig 9-56 gives a design chart for λ/4 T antennas. If there is not enough room for a single flat-top wire, two wires (or any number of wires positioned in equal increments on a 360° circle) can be used. If you use two in-line wires the length of the wire will be about 60% of the length of a single wire.

8.2. Voltage-Fed T Antennas

Voltage-fed T antennas are loaded vertical antennas with a current minimum at ground level. A specific case consists of a quarter-wave vertical, loaded with a half-wave top wire. **Fig 9-96** shows the configuration of this antenna and the current distribution. In this case, the impedance at the base of the antenna is high and purely resistive. The current maximum is at the antenna top. The antenna is sometimes called an inverted vertical, as it has its current maximum at the top. In theory, the current in both halves of the flat-top wire is such that radiation from that wire is zero. (In practice there is a very small amount of horizontal radiation.) The disadvantage of this construction is that the antenna requires a very long (λ/2 long) flat-top wire. Fig 9-96 also shows the dimensions for such a vertical for a practical design on 3.5 MHz.

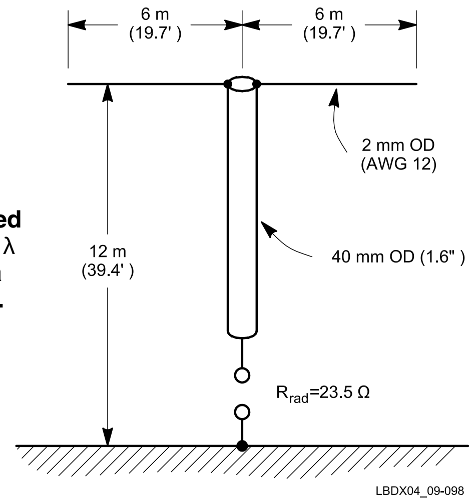


Fig 9-95 — T-wire loaded current-fed λ λ/4 antenna for 3.5 MHz.

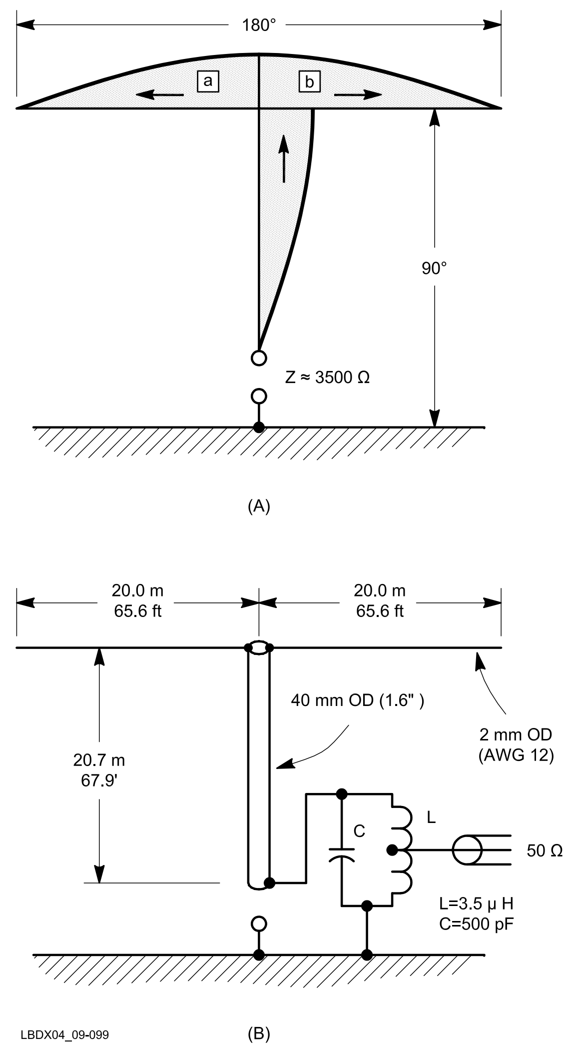


Fig 9-96 — Voltage-fed 80-meter λ/4 vertical (also called inverted vertical), using a λ/2 long top-loading wire. The T wire has a twofold function — providing a low impedance at the top of the vertical, and having a configuration whereby horizontally polarized radiation is essentially canceled (area A = area B, hence no radiation). C is ~500 pF and L is ~3.5 µH.

Hille, DL1VU, dramatically improved the T antenna design by folding the λ/2 flat-top section in such a way that the radiation from the flat-top section is effectively suppressed. **Fig 9-97** shows the configuration of this antenna. It can easily be proved that the area under the current distribution line for the central part (which is λ/12 long) is the same as the area for the remaining part of the loading device (which is λ/6 long). Because of the way the wires are folded, the radiation from the horizontal loading device is effectively canceled.

The latest design of a T-type top-load by Hille requires only a single λ/4 flat top. To cancel all possible horizontal radiation from this flat-top wire, the λ/4 is folded back as shown in **Fig 9-98**. Notice that the top load is asymmetrical.

A single quarter-wave flat top acts as a short circuit at the top of the vertical, the same way that radials provide a low-impedance attachment point for the outer conductor of the coax feed line in the case of a ground-plane antenna.

Hille also described a vertical with a physical length of only 0.39 λ, using the λ/4-long top-load wire configuration described above (Ref 7991). This antenna produces the same field strength as a 5λ/8 (0.64-λ) vertical antenna.

The T antenna can also be seen as a Bobtail Curtain antenna with the two vertical end sections missing. As such, this antenna is a poor performer with respect to the Bobtail antenna, where the directivity and gain is obtained through the use of three vertical elements.

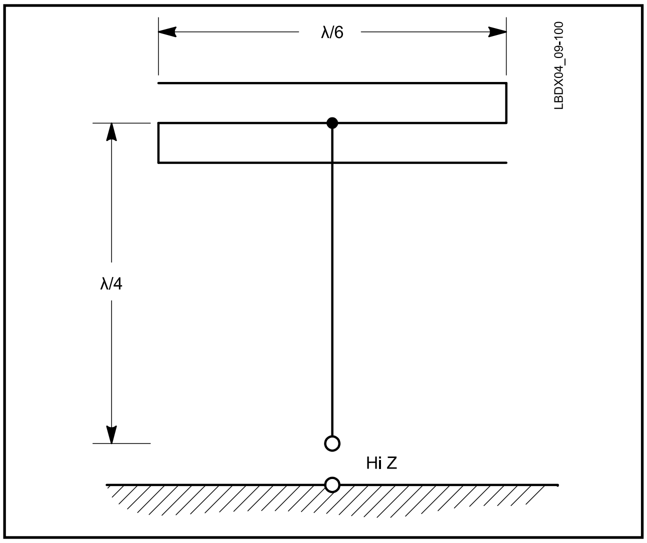


Fig 9-97 — Voltage-fed λ/4 T antenna with the λ/2 flattop wire folded to have a total span of only λ/6. The current distribution in the folded top load is such that radiation from the top load is effectively canceled. The advantage of this design over the original voltage-fed T antenna is that it requires a much shorter top-load space.

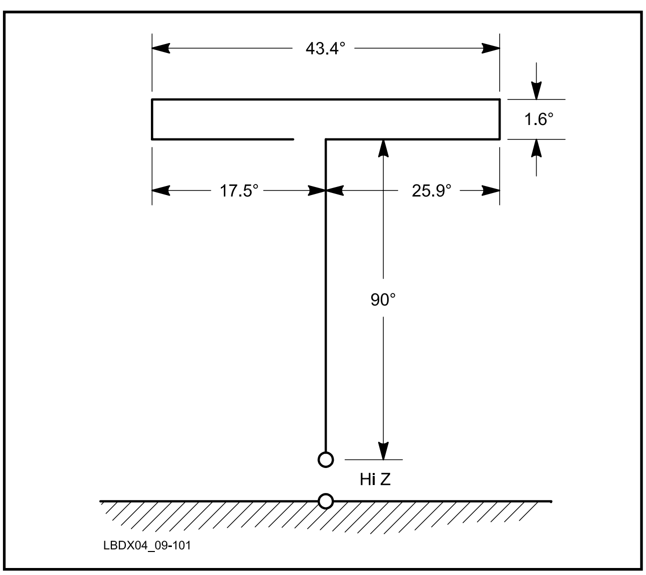


Fig 9-98 — Voltage-fed T antenna with a λ/4 long top load, arranged in such a way that there is no radiation from the flat-top section.

8.2.1. Feeding the Antenna

The voltage-fed T antenna can best be fed by means of a parallel tuned circuit (see Fig 9-96). You can either tap the coax on the coil for the lowest SWR point or tap the antenna near the top of the coil. Either method is valid.

8.2.2. The Required Ground and Radial System

The ground and radial requirements are identical to those required for a λ/2 vertical (see Section 4.3).

8.3. Close-Spaced Short Vertical and Reduced Losses

If you are in a situation where you cannot put down a good radial system (such as 100 λ/4 radials) for your short verticals, but you can erect several of those verticals close together (eg, with λ/16 spacing), this can be a way of improving the efficiency of your antenna.

John, W1FV, wrote: "I've been doing this for years on 160 with three 60-foot verticals (actually my 80-meter vertical system) spaced 35 feet apart. When fed in-phase, the feed-point radiation resistance at each vertical is around 18 ohms without a top hat. A single 60-foot vertical system has a radiation resistance of around one third of that. When the total system resistive loss (ground loss plus other component losses) is high (much bigger than 5-6 ohms), the efficiency of the three vertical system would be improved as much as a factor of three (5 dB) over a single vertical. For two in-phase verticals, the improvement would be around 3 dB. When the system loss starts out low and the single vertical efficiency is pretty good, there is obviously less to be gained, but that's also true when other loading schemes are used with short verticals."

This concept of using close-spaced in-phase verticals dates from 1920 (described in Jasik's *Antenna Engineering Handbook*, 1st Edition, page 19-9). Ground loss remains constant for a given area of ground system and antenna, because the sum of currents from each vertical flowing into that fixed size ground system remains exactly the same no matter how many verticals are added.

The only improvement occurs when multiple antennas are far enough apart so that return currents at the base of one vertical are not influenced by currents from another vertical. This means that each antenna must have a small ground system area, well separated from the ground systems of the other verticals. The end result, however, is no better than making a single large ground system the exact size of the sum of the small ground systems.

To have an efficiency advantage in such a configuration there must be loss present in the verticals, including resistive ground loss (for example, a very small radial system). Resistive loss is proportional to current squared, so reducing current for constant power reduces loss. Since the drive currents to the verticals are split equally (in the ideal case) between N verticals, the current per vertical is 1/N times the current that would flow in one of the verticals by itself. This neglects the effects of mutual coupling, which are usually rather insignificant between short monopoles. This means the loss per vertical is $(1/N)^2$ times the loss of the single vertical. Since there are N verticals, the net system loss is N times $(1/N)^2$, or just 1/N times the loss of one vertical. This can be a significant improvement over a single vertical that would otherwise be lossy or inefficient by itself.

Tom, W8JI pointed out a possible application: "Where this would help is when a driveway would be in the middle of an area, and you couldn't cross the driveway with radials. You could build an antenna consisting of two verticals, with one on either side of the driveway, and separate "half" ground systems on either side that are not connected. In this example efficiency would be identical to a single vertical in the middle of the driveway with a full radial system that covers exactly the same physical area, but you can still have a driveway."

Another application of the principle would be where you would use four 80-meter verticals forming an 80-meter Four Square, each vertical using a radial system designed for 80 meters, and where you would feed all four verticals in phase. That system acts like a single vertical of the same height placed in the exact middle of the 80-meter array. That would be better than feeding only one element at the edge of the ground system. But it gains nothing over a single vertical loaded the same way with the same area ground system, except convenience. W8JI pointed out: "A Four Square works the same way. The center two elements combine to effectively make one element in the middle of the array. That is why we can feed a four-square with a 1:1:1:1 current ratio when a three-element array requires a 1:2:1 ratio! The center two elements (being in-phase) form one "radiation fat" element."

W8JI concluded saying: "If it were a 160-only array, he almost certainly would be better off putting the same effort into a single vertical and one big ground system covering the same overall area. RCA found this to be true in an actual test at a VLF station, where they initially used multiple antenna elements over multiple distributed grounds to obtain the same 1/N efficiency as described above. When they pulled the multiple elements and the multiple independent ground systems out and replaced everything with a conventional system of radials filling the same area, efficiency actually went up a considerable amount (and they got rid of many maintenance headaches)."

Tom also points out that in recent tests (on VLF) the Air Force did at Marion the conclusion was a normal large radial ground system resulted in considerably less ground loss than had previously been obtained with a combination of multiple verticals using independent smaller ground systems, with a complex overhead distribution and equalizing system.

9. LOCATION OF A VERTICAL ANTENNA

Let's tackle the often-asked question, "Will a vertical work in my particular location?" Verticals for working DX on the low bands are certainly not space-saving antennas but to the contrary, require a lot of space and a good ground. Many low-band DXers have wondered why some verticals don't work well at all, while others work "like gangbusters." The poor performers generally have the poor locations. To repeat, a vertical is not a space-saving antenna! A good vertical takes a lot of real estate. In addition, it must be real estate with a good RF ground!

The standard for buried radials is that for best radiation efficiency you need 120 λ/2 radials. This means that for 80 meters, you need about an 80×80 m lot in which to place all the radials. The radials are there to provide a low-resistance return path for the antenna current to achieve good efficiency.

The area beyond the ends of the radials is at least as important, because that's where the low-angle reflection at ground level takes place (the Fresnel zone). This is where the reflection efficiency is determined.

Up to λ/2 away from the vertical, most of the reflection will take place that is responsible for the 25° radiation (main angle) of a typical λ/4 vertical over average ground. Therefore, beyond this point, a clear path should be available for these low-angle rays to obtain maximum low-angle radiation. It is clear that for even lower angles of radiation, the ground at even greater distances becomes important. As explained earlier, this is even more so with "long" verticals (eg, λ/2 vertical), where the Fresnel reflection takes place up to 100 λ away from the antenna (for wave angles down to 0.25°).

To avoid excessive absorption, verticals should be kept at least λ/4 away from residential houses. This means, for instance, that at a point 60 meters from a 3.5-MHz antenna, the maximum height of a structure should be limited to 9.1 meters. What about trees closer in? Trees can be reasonably good conductors and can be very lossy elements in the near field of a radiator. A case has been reported in the literature where a λ/4 vertical with an excellent ground system showed a much lower radiation resistance than expected. It was found that trees in the immediate area were coupling heavily with the vertical and were causing the radiation resistance of the vertical to be very low. Under such circumstances of uncontrolled coupling into very lossy elements, far from optimum performance can be expected. Of course, if the trees are short in relation to the quarter wavelength, it is reasonable to assume that the result of such coupling will be minimal.

Even though neighboring (lossy) structures such as trees may not be resonant, they will always absorb some RF to an unknown degree. Other objects that are likely to affect the performance of a vertical are nearby antennas and towers. Mutual coupling can be considered the culprit if the radiation resistance of the vertical is lower than expected. Another way of checking for coupling with other antennas is to alternately open and short-circuit the suspected antenna feed lines while watching the SWR or the radiation resistance of the vertical antenna. If there is any change, you are in trouble. Checking for resonance of towers has been described in Section 7.

It may come as a surprise that a vertical is so demanding of space. Most amateur verticals are not anywhere near ideal, yet good performance can still be obtained from practical setups. But the builder of a vertical should understand which factors are important for optimum performance, and why.

10. 160-METER DXPEDITION ANTENNAS

I have talked at great length with well-known DXpeditioners who have been especially successful on the low bands. I'd like to share the following rules with candidate DXpeditioners with respect to the low bands.

If you're on an island, erect the station on that side of the island where you will have the most difficult propagation path or where you are facing the most stations (eg, if you are on an island in the South Indian Ocean, try a shore on the northwest side of the island, looking into both Europe and North America). By all means erect the antenna very close to saltwater (which means no further than ¼ wave from the water), or over (or in) saltwater. This will help you lower the pseudo-Brewster angle, and will ensure a good low-angle take off. Use two in-line gull wing radials, parallel to the water line (see Section 2.2.5., Fig 9-27)

Unless you have a very tall support of at least 30 meters height, use a vertical. Good choices are the Battle Creek Special, the BC Trapper, the AKI Special, the Titanex V160E or any top loaded vertical (T antenna) or inverted L, for which you should try to make the vertical part as long as possible. The vertical section should be at least 15 meters tall, 12 meters being an absolute minimum for 160 meters. If there are some trees, you may try to climb a tall tree, and use a collapsible fiberglass fishing rod (they exist in 12 meter lengths) to extend the effective support height. Use as many radials as you can. Very thin wire is just fine if you use many (current is shared by the many wires). A small spool of #28 enameled copper magnet wire can hold a lot of wire and takes little space. If you are within ¼ λ of the saltwater shore , just use two elevated radials (gull wing). These make switching from the CW to the phone band very simple by merely adjusting the length of the two radials.

The Titanex verticals are very special in that they are made of an aluminum-titanium alloy that is very strong and extremely lightweight. The model V160E vertical is a 26.7-meter long vertical that weighs only 7.5 kg. See **Fig 9-99**. The maximum section length is only 2.1 meters, and the total antenna can easily be erected by two or three persons. This, as well as the low weight, makes this a very attractive antenna for DXpeditions. The guy wires are 2-mm Kevlar, and guys are placed at heights of 6, 9, 12, 15 and 18 meters. The upper 8 meters of the vertical swings freely in the wind. With a total length of 26 meters, this antenna has a very respectable radiation resistance of 12 Ω, which is 50% higher than that of the Battle Creek Special (which is 10 meters shorter!). The antenna is 3λ/8 on 80, and 5λ/8 on 40. Also on 40 this should make it a killer antenna if erected over saltwater. The V80E vertical measures only 20 meters tall, which is good for a R_{rad} of about 8.5 Ω, which is similar to the R_{rad} of the Battle Creek Special. Titanex also provides a three-band relay-switched matchbox providing a 1:1 SWR on the three low bands. More info at **www.titanex.de**. The Titanex antennas are expensive mechanical marvels but they have been used extremely successfully during a number of expeditions including VK9CR, VK9XY, C56CW, FW2OI, S21XX, P29VXX, DL7FD/HR3, K7K, K4M, T31BB, 9MØC, TJ1GB, ZL7DK, YJØADJ, FOØFI, FOØFR, 3B7RF, CN8WW, 3B9R, CEØZY, A52A, D68C, 3GØY and 3B9C.

Fig 9-99 — The Titanex V160E antenna on the beach on VK9CR, surrounded by beautiful coconut trees. This 26.8-meter-long special DXpedition vertical weighs only 7.5 kg and disassembles into 2.1-meter long sections, ideal for traveling!

Don't bother putting up a Beverage over saltwater; it won't work well. If you set it up along the sea shore, stay at least one quarter, preferably a half wavelength away from the perfect underground (saltwater). You can terminate it in saltwater though, as long as it does not run across saltwater or ground that is nearly a conductor!

If there is a tall support, you may want to use a sloping half-wave vertical, especially if you are near the sea (see Chapter 8 on dipole antennas). The sloping vertical builds up its image as far as 100 λ away from the antenna. If there is no saltwater nearby and ground conductivity is poor, use a high support for an inverted-V dipole. Don't try an inverted V or any other horizontally polarized antenna at a height of 15 meters or less. All you will get is very high-angle radiation.

Here is a hint from Rudi, DK7PE: If you are on a DXpedition in a country with a substantial tourist business, choose the tallest hotel (Hilton, Sheraton or Intercontinental hotels usually do well in this respect). Slope a dipole from the top of the building to some distant point and let the feed line come to your room, which can be a few stories below the roof. Make the dipole as vertical as possible. This is by far the best antenna if you are in such a situation.

DK7PE proved it during his operation from D2CW (August 1992) where he had his sloping dipole attached some 60 meters above street level, facing north, and within 1 λ of the South Atlantic Ocean. Rudi's signals were always S9 in Europe on 160 meters. During his more recent operation from Ethiopia and Eritrea (9F2CW), he proved it again. Rudi's total antenna system for his DXpeditions (covering 160 through 10 meters) can be packed in a small handbag. The RG-58 cable takes up 80% of the volume. The antenna consists of precut lengths of flexible insulated wire, with small insulators and a variety of alligator clips that let him change bands. On the higher bands he can configure the wire into a 2-element Yagi.

R. E. Tanaka, 9M2AX, well-known 160-meter operator

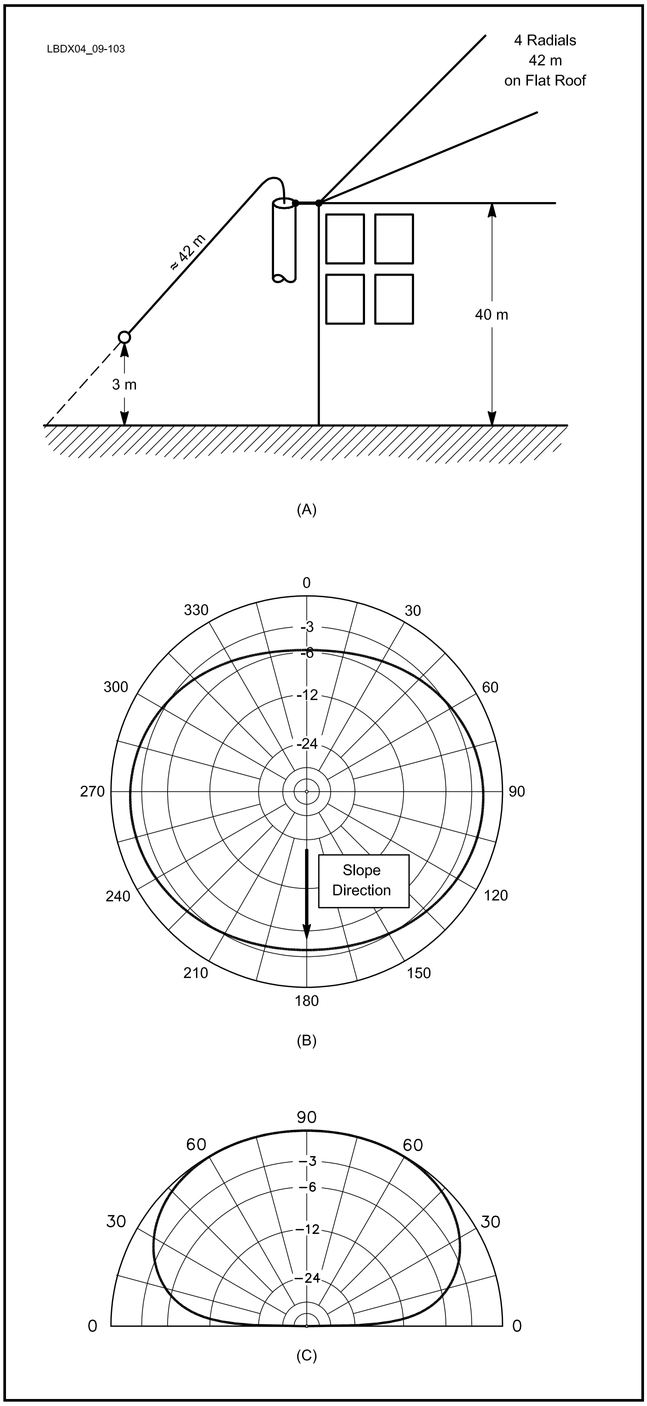


Fig 9-100 — Configuration and radiation patterns of the inverted λ/4 sloper antenna used by 9M2AX. The azimuth pattern is shown at B for an elevation angle of 30°, and at C is the elevation pattern. (The elevation pattern is taken in the 90-270° direction as displayed in the azimuth pattern.) Note the relative high amount of high-angle radiation. Using just two radials in-line would improve this situation considerably.

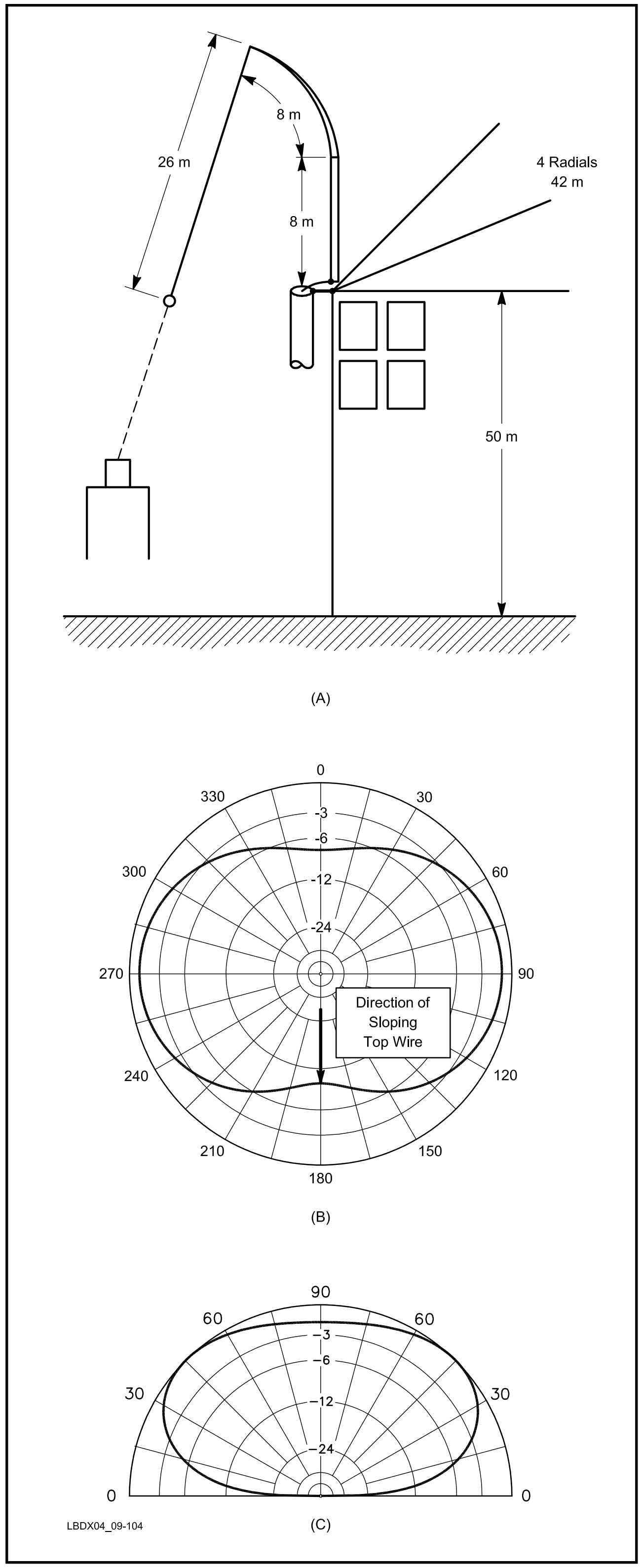


Fig 9-101 — Configuration and radiation patterns of the inverted λ/4 sloper antenna used by 9M2AX during his operation from East Malaysia (Sarawak) as 9M8AX. The azimuth pattern is shown at B, and the elevation pattern at C. (The elevation pattern is taken in the 90-270° direction as displayed in the azimuth pattern.) The antenna was installed on the edge of a 50-meter high flat roof. Four λ/4 long radials were laid on the roof. The metal mast plus the fiberglass rod are 16 meters long. The sloping wire was adjusted for minimum SWR at resonance.

from the Far East, sent me the sketches of the antennas he is using in 9M2 as well as when he operated from 9M8AX. The antennas Ross was using can be put up at any tall hotel, and should be excellent suggestions for 160-meter DXpeditioners.

Fig 9-100 and **Fig 9-101** show the layouts of the two antenna setups and their radiation patterns. The radial system covering only one quadrant (90°) results in a significant high-angle radiation component with the 9M2AX version. The low-angle radiation is very pronounced as well. From modeling, the "inverted sloping wire vertical" from the 9M2 QTH has a feed-point impedance of about 75 Ω.

The 9M8AX configuration is an inverted L with a sloping flat-top. The calculated impedance from modeling is nearly 60 Ω. This antenna has better low-angle radiation than the 9M2AX version, which is normal. In order to eliminate the high-angle radiation for the 9M2AX version, it would be necessary to install just two radials (in-line with one another), so that the radiation from these wires would be canceled. The radials are *not* there to provide a ground plane, but are merely serving to provide a low-impedance point to which to connect the outer shield of the feed line. One λ/4 radial would serve that purpose, but would radiate a lot of horizontal component. Two radials in-line would provide a low impedance point just as well, but would not radiate a high-angle horizontal component.

11. BUYING A COMMERCIAL VERTICAL

I sincerely hope that this chapter on verticals has incited you to build your own antenna. You cannot believe how much more satisfaction you get out of using something you made or designed yourself, rather than going to the store, opening your wallet and then playing the appliance-type ham.

Anyhow, if you choose not to make your own, here are a few rules to help you select your new low-band commercial vertical:

1) Most, if not all companies advertising their products, largely exaggerate the performance, especially if it's something different from a straightforward vertical.
2) A short vertical with a large bandwidth means there are a lot of losses. With short antennas a large bandwidth is a direct measure of its poor efficiency (lots of losses).
3) The efficiency of a vertical is in the first place determined by the physical length of the vertical (and the ground system, which you will have to install yourself anyhow).
4) Only top loading is efficient.
5) Verticals with coil loading are bound to be inefficient. An 8-meter-long (short) vertical with center-coil loading is bound to be a very poor performer on 160 meters as a transmitting antenna.
6) To be a reasonable performer a minimum physical vertical length of about 15 m is needed on 160 meters.
7) Good hardware (stainless steel, good finishing, etc) are no guarantee for a good antenna.
8) A fancy feed system or folded elements that claim to reduce losses and increase efficiency are a total fallacy.
9) A producer of a 160-meter vertical who prescribes using only a few 10-meter long radials does not know what he is talking about.
10) Advertisers bragging that their product is bought by government agencies are not proving anything. Remember the Maxcom "dummy load" antenna-matching network used extensively by the armed forces?
11) An advertiser specifying his 8-meter-long 160-meter vertical has 75% efficiency, without specifying the ground radial system is telling you stories.
12) Advertisers selling their product by telling how many new countries one of their customers has worked with it, are… Well, you know. Maybe, with a good homemade vertical he would have worked double the number of new countries! Not very scientific advertising, anyhow.

Spending nearly $500 for a 9-meter-long radiator is a heck of a lot of money. You could buy some simple aluminum tubing (TV-type push-up mast, about $70), some copper wire to make a number of top-loading wires (add another $10), some nylon guy rope (another $10), maybe an (empty) Coke bottle for an insulator (free), and you have exactly the same for about 20% of the price of the commercial thing. It won't work any better, but at least you won't feel like you've been robbed. And spending nearly $400 for an 8-meter-long 160-meter vertical, with a slim (and thus very lossy) loading coil, is even worse, of course.

Amateur Radio is a technical hobby. It is true that the progress of microelectronics has made it very difficult for the average ham to do much home designing and home building in the field of receivers and transmitters. Building antennas is one of the few fields where we can, ourselves, through our own knowledge, understanding and expertise, do as well and usually much better than the commercial companies. Let's grab this opportunity with both hands, and build our own vertical for the low bands. This will give you the ultimate kick, I promise you!

Contents

CHAPTER 10

Large Loop Antennas

The delta-loop antenna is a superb example of a high-performance compromise antenna. The single-element loop antenna is almost exclusively used on the low bands, where it can produce low-angle radiation, requiring only a single quarter-wave high support. We will see that a vertically polarized loop is really an array of two phased verticals, and that the ground requirements are the same as for any other vertically polarized antenna.

This means that with low delta loops, the horizontal wire will couple heavily to the lossy ground and induce significant losses, unless we have improved the ground by putting a ground screen under the antenna. (See Chapter 9, Section 1.3.3 and Section 2.) I have seen it stated in various places that delta loops don't require a good ground system. This is as true as saying that verticals with only a single or a few elevated radials don't require a good ground system, which is not true.

Loop antennas have been popular with 80-meter DXers for nearly 40 years. Resonant loop antennas have a circumference of 1 λ. The exact shape of the loop is not particularly important. In free space, the loop with the highest gain, however, is the loop with the shape that encloses the largest area for a given circumference. This is a circular loop, which is difficult to construct. Second best is the square loop (quad), and in third place comes the equilateral triangle (delta) loop (Ref 677).

The maximum gain of a 1-λ loop over a λ/2 dipole in free space is approximately 1.35 dB. Delta loops are used extensively on the low bands at apex heights of λ/4 to 3λ/8 above ground. At such heights the vertically polarized loops far outperform dipoles or inverted-V dipoles for low-angle DXing, assuming good ground conductivity.

Loops are generally erected with the plane of the loop perpendicular to the ground. Whether or not the loop produces a vertically or a horizontally polarized signal (or a combination of both) depends only on how (or on which side) the loop is being fed.

Another type of large loop antennas comprises the horizontally mounted loops, which have the plane of the loop parallel to the ground. These antennas produce horizontal radiation with takeoff angles determined, as usual, by the height of the horizontal loop over ground.

1. QUAD LOOPS

Belcher, WA4JVE, Casper, K4HKX, (Ref 1128), and Dietrich, WAØRDX, (Ref 677), have published studies comparing the horizontally polarized vertical quad loop with a dipole. A horizontally polarized quad loop antenna (**Fig 10-1A**) can be seen as two short, end-loaded dipoles stacked λ/4 apart, with the top antenna at λ/4 and the bottom one just above ground level.

There is no broadside radiation from the vertical wires

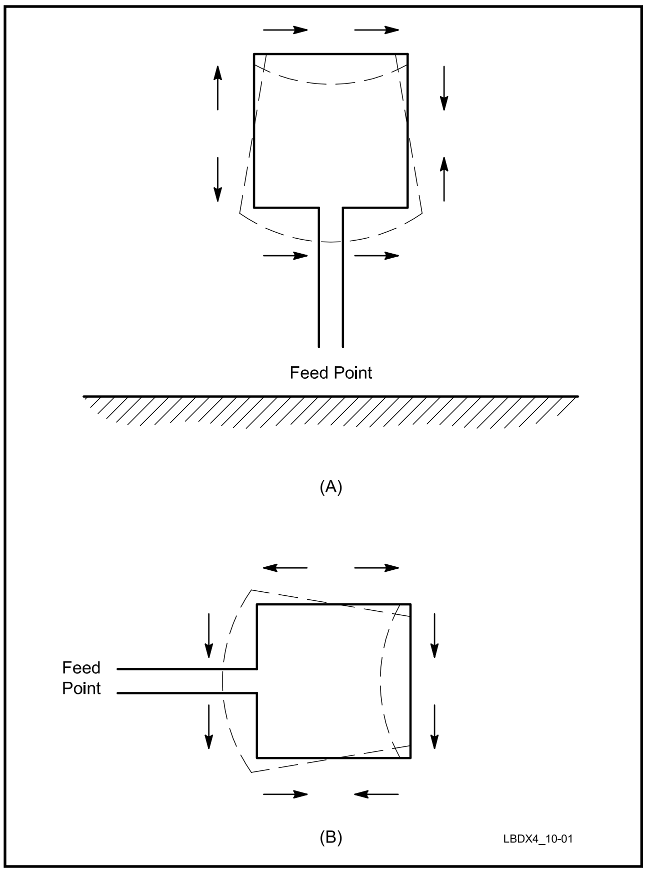


Fig 10-1 — Quad loops with a 1-λ circumference. The current distribution is shown for (A) horizontal and (B) vertical polarization. Note how the opposing currents in the two legs result in cancellation of the radiation in the plane of those legs, while the currents in the other legs are in-phase and reinforce each other in the broadside direction (perpendicular to the plane of the antenna).

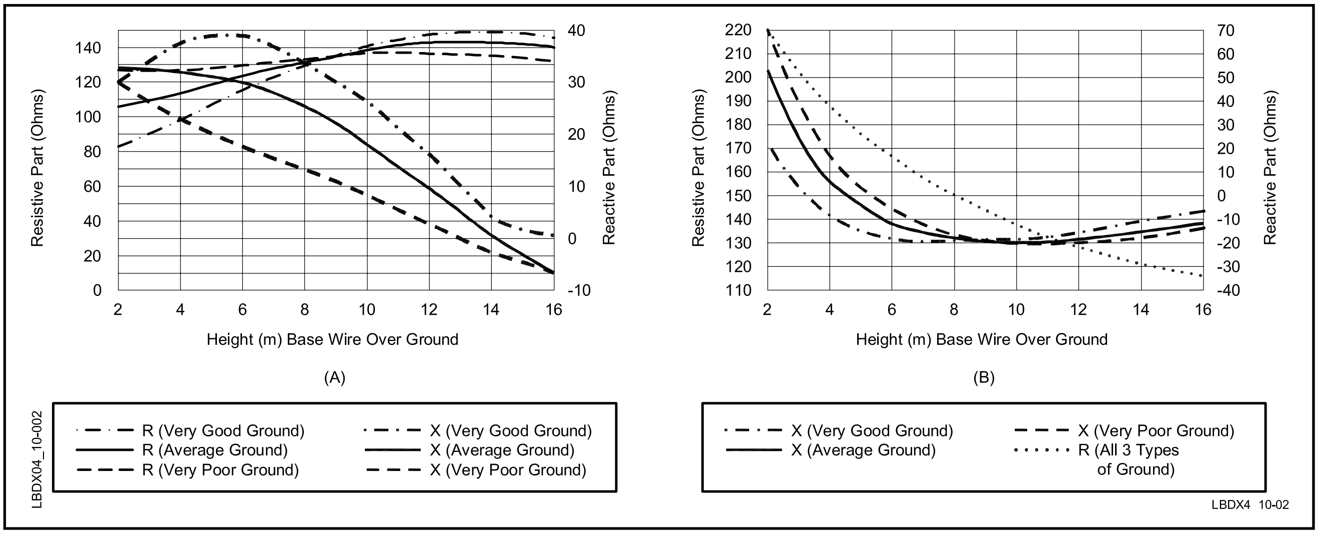


Fig 10-2 — Radiation resistance and feed-point resistance for square loops at different heights above real ground. The loop was first dimensioned to be resonant in free space (reactance equal to zero), and those dimensions were used for calculating the impedance over ground. At A, for horizontal polarization, and at B, for vertical polarization. Analysis was with *NEC* at 3.75 MHz.

of the quad because of the current opposition in the vertical members.

In a similar manner, the vertically polarized quad loop consists of two top-loaded, λ/4 vertical dipoles, spaced λ/4 apart. Fig 10-1B shows how the current distribution along the elements produces cancellation of radiation from certain parts of the antenna, while radiation from other parts (the horizontally or vertically stacked short dipoles) is reinforced.

The square quad can be fed for either horizontal or vertical polarization merely by placing the feed point at the center of a horizontal arm or at the center of a vertical arm. At the higher frequencies in the HF range, where the quads are typically half to several wavelengths high, quad loops are usually fed to produce horizontal polarization, although there is no specific reason for this except maybe from a mechanical standpoint. Polarization by itself is of little importance at HF (except — to a certain degree — on 160 meters! See Chapter 1, Section 3.5), because it becomes random after ionospheric reflection.

1.1 Impedance

The radiation resistance of an equilateral quad loop in free space is approximately 120 Ω. The radiation resistance for a quad loop as a function of its height above ground is given in **Fig 10-2**. The impedance data were obtained by modeling an equilateral quad loop over three types of ground (very good, average and very poor ground) using *NEC*. *MININEC* cannot be used for calculating loop impedances at low heights (see Section 2.9).

The reactance data can assist you in evaluating the influence of the antenna height on the resonant frequency. The loop antenna was first modeled in free space to be resonant at 3.75 MHz and the reactance data was obtained with those free-space resonant-loop dimensions.

For the vertically polarized quad loop, the resistive part of the impedance changes very little with the type of ground under the antenna. The feed-point reactance is influenced by the ground quality, especially at lower heights. For the horizontally polarized loop, the radiation resistance is noticeably influenced by the ground quality, especially at low heights. The same is true for the reactance.

1.2. Square Loop Patterns

1.2.1. Vertical Polarization

The vertically polarized quad loop, Fig 10-1B, can be considered as two shortened top-loaded vertical dipoles, spaced λ/4 apart. Broadside radiation from the horizontal elements of the quad is canceled, because of the opposition of currents in the vertical legs. The wave angle in the broadside direction will be essentially the same as for either of the vertical members. The resulting radiation angle will depend on the quality of the ground up to several wavelengths away from the antenna, as is the case with all vertically polarized antennas.

The quality of the reflecting ground will also influence the gain of the vertically polarized loop to a great extent. The quality of the ground is as important as it is for any other vertical antenna, meaning that vertically polarized loops close to the ground will not work well over poor soil.

Fig 10-3 shows both the azimuth and elevation radiation patterns of a vertically polarized quad loop with a top height of 0.3 λ (bottom wire at approximately 0.04 λ). This is a very realistic situation, especially on 80 meters. The loop radiates an excellent low-angle wave (lobe peak at approximately 21°) when operated over average ground. Over poorer ground, the wave angle would be closer to 30°. The horizontal directivity, Fig 10-3C, is rather poor, and amounts to approximately 3.3 dB of side rejection at any wave angle.

1.2.2. Horizontal Polarization

A horizontally polarized quad-loop antenna (two stacked short dipoles) produces a wave angle that is dependent on the height of the loop. The low horizontally polarized quad (top at 0.3 λ) radiates most of its energy right at or near zenith angle (straight up).

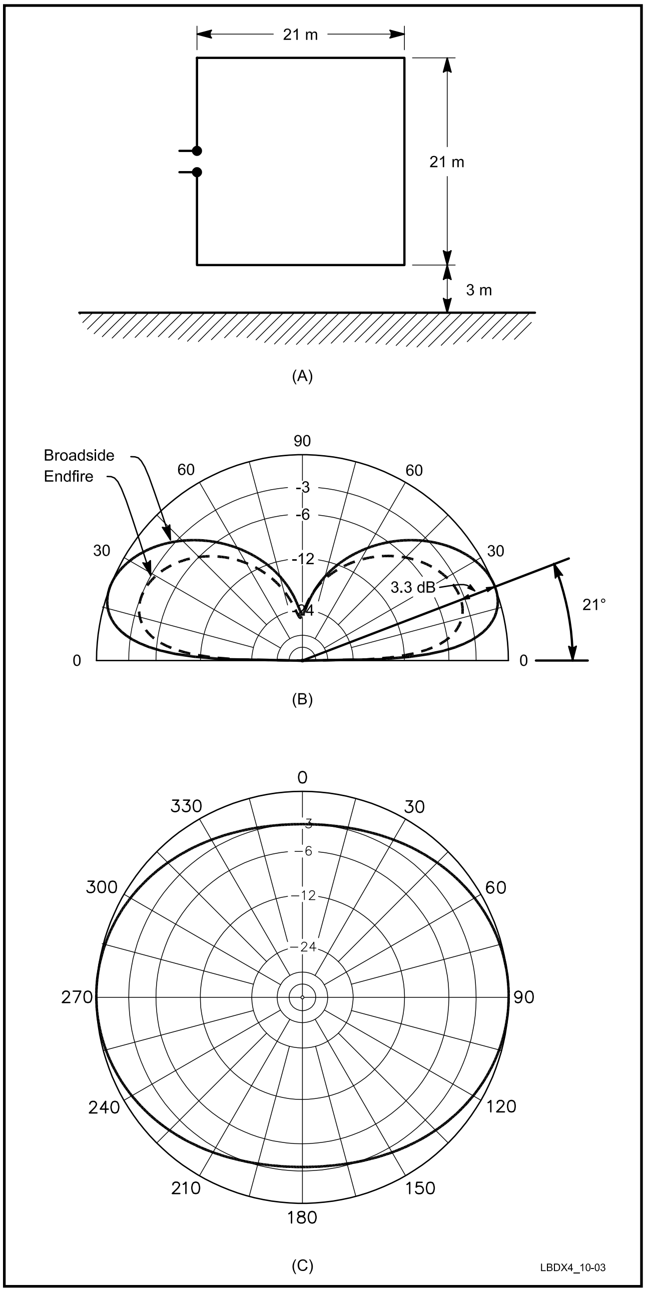


Fig 10-3 — Shown at A is a square loop, with its elevation-plane pattern at B and azimuth pattern at C. The patterns are generated for good ground. The bottom wire is 0.0375 λ above ground (3 meters or 10 feet on 80 meters). At C, the pattern is for a wave angle of 21°.

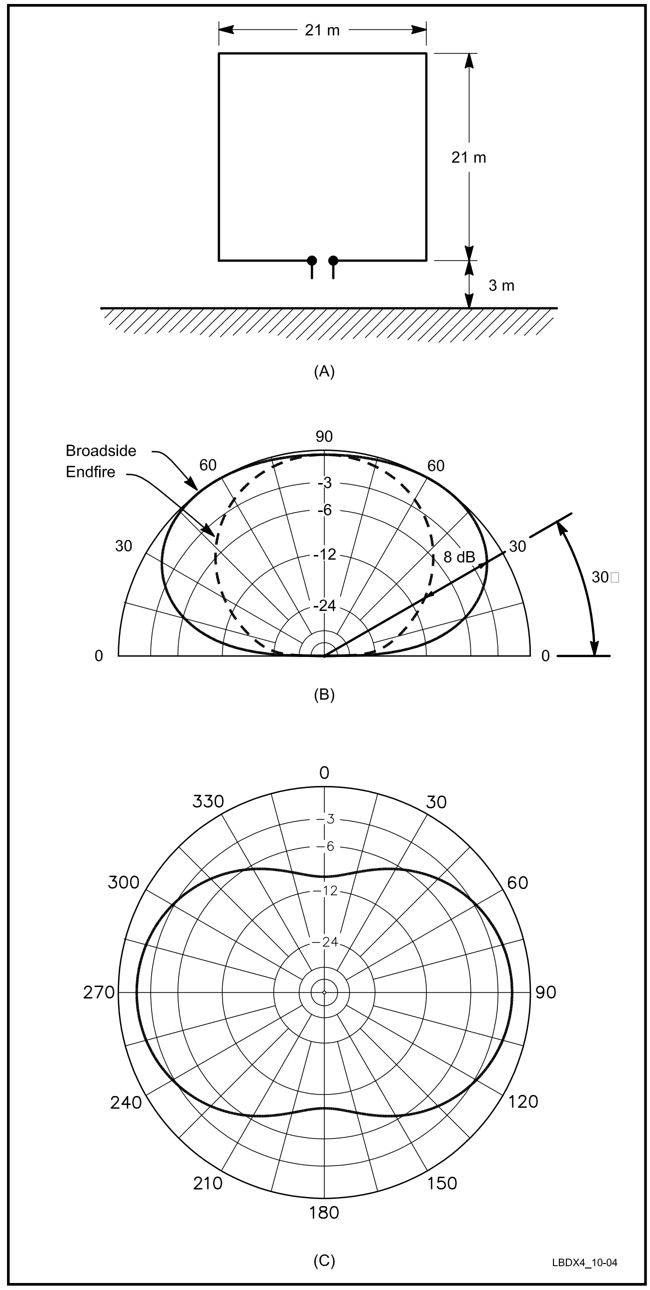


Fig 10-4 — Azimuth and elevation patterns of the horizontally polarized quad loop at low height (bottom wire 0.0375 λ above ground). At an elevation angle of 30°, the loop has a front-to-side ratio of approximately 8 dB.

Fig 10-4 shows directivity patterns for a horizontally polarized loop. The horizontal pattern, Fig 10-4C, is plotted for a takeoff angle of 30°. At low wave angles (20° to 45°), the horizontally polarized loop shows more front-to-side ratio (5 to 10 dB) than the vertically polarized rectangular loop.

1.2.3. Vertical versus Horizontal Polarization

Vertically polarized loops should be used only where very good ground conductivity is available. From **Fig 10-5A** we see that the gain of the vertically polarized quad loop, as well as the wave angle, does not change very much as a function of the antenna height. This makes sense, since the vertically polarized loop is in the first place two phased verticals, each with its own radial.

However, the gain is drastically influenced by the quality of the ground. At low heights, the gain difference between very poor ground and very good ground is a solid 5 dB! The wave angle for the vertically polarized quad loop at a low height (bottom wire at 0.03 λ) varies from 25° over very poor ground to 17° over very good ground.

I have frequently read in Internet messages that a delta loop has certain advantages over a vertical antenna (or arrays

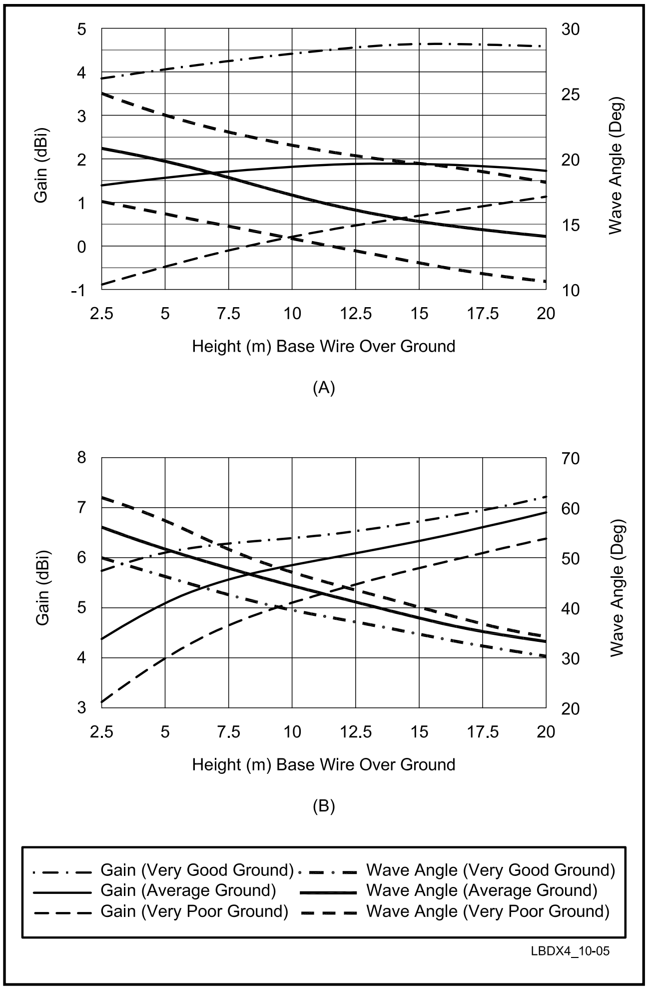


Fig 10-5 — Radiation angle and gain of the horizontally and the vertically polarized square loops at different heights over good ground. At A, for vertical polarization, and at B, for horizontal polarization. Note that the gain of the vertically polarized loop never exceeds 4.6 dBi, but its wave angle is low for any height (14 to 20°). The horizontally polarized loop can exhibit a much higher gain provided the loop is very high. Modeling was done over average ground for a frequency of 3.75 MHz, using *NEC*.

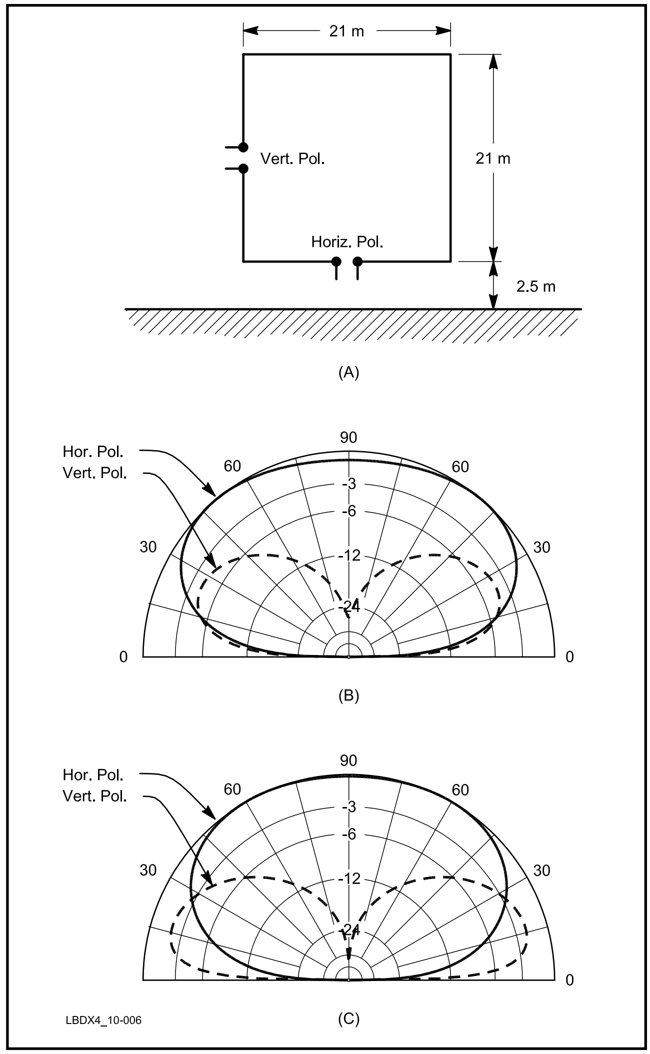


Fig 10-6 — Superimposed (same dB scale) patterns for horizontally and vertically polarized square quad loops (shown at A) over very poor ground (B) and very good ground (C). In the vertical polarization mode the ground quality is of utmost importance, as it is with all verticals. See also Fig 10-14.

of vertical antennas) since the loop antenna does not require any radials. This statement is really quite misleading. It is like saying that a vertical with just one or a few elevated radials does not require a good ground underneath. Indeed, in a delta loop (and a quad loop), the "element" that takes care of the return current is part of the antenna itself just like with a dipole! Vertically polarized delta loops at low height always require a good ground screen underneath the antenna (unless they are over excellent or perfect ground), exactly in the same way that a vertical with only one or two radials requires a good ground underneath the radials.

With a horizontally polarized quad loop the wave angle is very dependent on the antenna height, but not so much on the quality of the ground. At very low heights, the main wave angle varies between 50° and 60° (but is rather constant all the way up to 90°), but these are rather useless radiation angles for DX work. As far as gain is concerned, there is a 2.5-dB gain difference between very good and very poor ground, which is only half the difference we found with the vertically polarized loop. Comparing the gain to the gain of the vertically polarized loop, we see that at very low antenna heights the gain is about 3-dB better than for the vertically polarized loop. But this gain exists at a high wave angle (50° to 90°), while the vertically polarized loop at very low heights radiates at 17° to 25°.

Fig 10-6 shows the vertical-plane radiation patterns for both types of quad loops over very poor ground and over very good ground on the same dB scale. For more details see Section 2.3.

1.3. A Rectangular Quad Loop

A rectangular quad loop, with unequal side dimensions, can be used with very good results on the low bands. An impressive signal used to be generated by 5NØMVE from Nigeria with such a loop antenna. The single quad-loop element is strung between two 30-meter high coconut trees, some 57 meters apart in the bush of Nigeria. 5NØMVE fed the loop

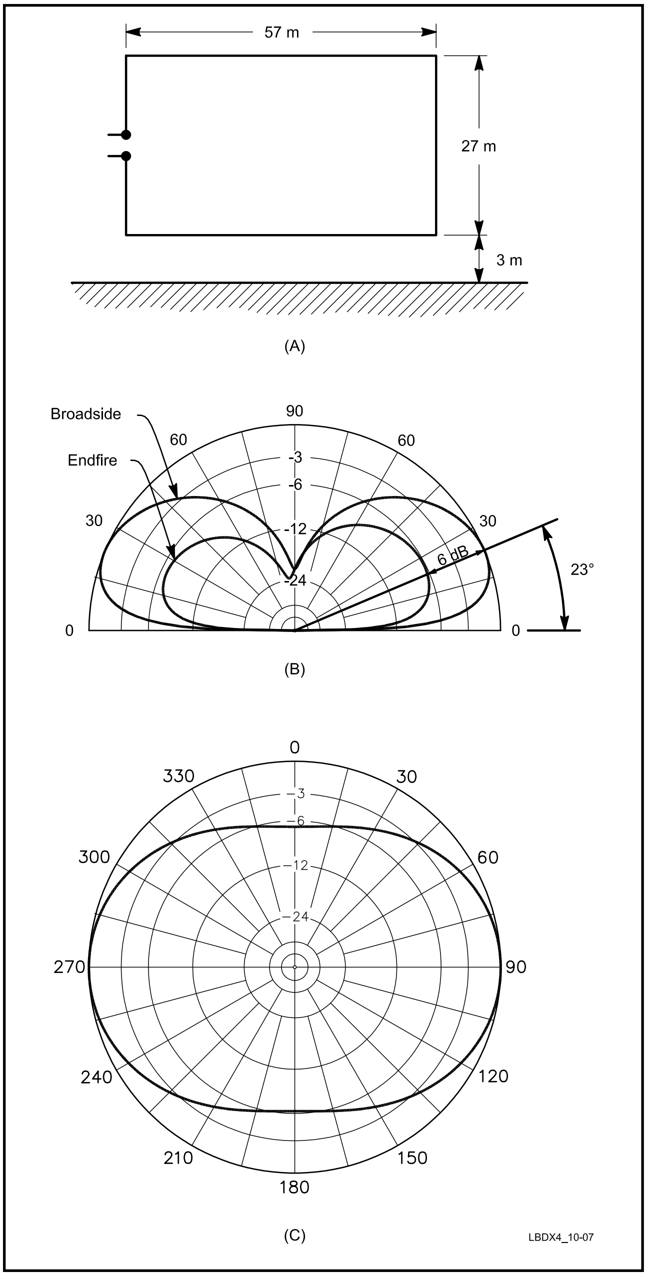


Fig 10-7 — At A, a rectangular loop with its baseline approximately twice as long as the vertical height. At B and C, the vertical and horizontal radiation patterns, generated over good ground. The loop was dimensioned to be resonant at 1.83 MHz. The azimuth pattern at C is taken at a 23° elevation angle.

in the center of one of the vertical members. He first tried to feed it for horizontal polarization but he says it did not work well. The vertical and the horizontal radiation patterns for this quad loop over good ground are shown in **Fig 10-7**. The horizontal directivity is approximately 6 dB (front-to-side ratio).

Even in free space, the impedance of the two varieties of this rectangular loop is not the same. When fed in the center of a short (27-meter) side, the radiation resistance at resonance is 44 Ω. When fed in the center of one of the long (57-meter) sides, the resistance is 215 Ω. Over real ground the feed-point impedance is different in both configurations as well; depending on the quality of the ground, the impedance varies between 40 and 90 Ω.

1.4. Loop Dimensions

The total length for a resonant loop is approximately 5 to 6% longer than the free-space wavelength.

1.5. Feeding the Quad Loop

The quad loop feed point is symmetrical, whether you feed the quad in the middle of the vertical or the horizontal wire. A balun should be used. Baluns are described in Chapter 6 on matching and feed lines.

Alternatively, you could use open-wire feeders (for example, 450-Ω line). The open-wire-feeder alternative has the advantage of being a lightweight solution. With a tuner you will be able to cover a wide frequency spectrum with no compromises.

2. DELTA LOOPS

Just as the inverted-V dipole has been described as the poor man's dipole, the delta loop can be called the poor man's quad loop. Because of its shape, the delta loop with the apex on top is a very popular antenna for the low bands; it needs only one support.

In free space the equilateral triangle produces the highest gain and the highest radiation resistance for a three-sided loop configuration. As we deviate from an equilateral triangle toward a triangle with a long baseline, the effective gain and the radiation resistance of the loop will decrease for a bottom-corner-fed delta loop. In the extreme case (where the height of the triangle is reduced to zero), the loop has become a half-wavelength-long transmission line that is shorted at the end, which shows a zero-Ω input impedance (radiation resistance), and thus zero radiation (well-balanced open-wire line does not radiate).

Just as with the quad loop, we can switch from horizontal to vertical polarization by changing the position of the feed point on the loop. For horizontal polarization the loop is fed either at the center of the baseline or at the top of the loop. For vertical polarization the loop should be fed on one of the sloping sides, at λ/4 from the apex of the delta. **Fig 10-8** shows the current distribution in both cases.

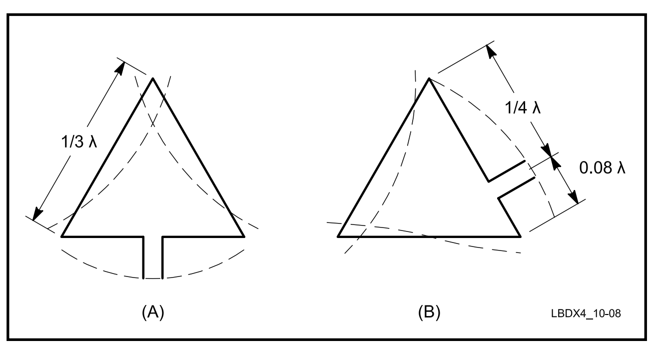


Fig 10-8 — Current distribution for equilateral delta loops fed for (A) horizontal and (B) vertical polarization.

2.1. Vertical Polarization

2.1.1. How it Works

Refer to **Fig 10-9**. In the vertical-polarization mode the delta loop can be seen as two sloping quarter-wave verticals (their apexes touch at the top of the support), while the baseline (and the part of the sloper under the feed point) takes care of feeding the "other" sloper with the correct phase. The top connection of the sloping verticals can be left open without changing anything about the operation of the delta loop. The same is true for the baseline, where the middle of the baseline could be opened without changing anything. These two points are the high-impedance points of the antenna. Either the apex or the center of the baseline must be shorted, however, in order to provide feed voltage to the other half of the antenna. Normally, of course, we use a fully closed loop in the standard delta loop, although for single-band operation this is not strictly necessary.

Assume we construct the antenna with the center of the horizontal bottom wire open. Now we can see the two half baselines as two λ/4 radials, one of which provides the necessary low-impedance point for connecting the shield of the coax. The other radial is connected to the bottom of the second sloping vertical, which is the other sloping wire of the delta loop.

This is similar to the situation encountered with a λ/4 vertical using a single elevated radial (see Chapter 9 on vertical antennas). The current distribution in the two quarter-wave radials is such that all radiation from these radials is effectively canceled. The same situation exists with the voltage-fed T antenna, where we use a half-wave flat top (equals two λ/4 radials) to provide the necessary low-impedance point to raise the current maximum to the top of the T antenna.

The vertically polarized delta loop is really an array of two λ/4 verticals, with the high-current points spaced 0.25 λ to 0.3 λ, and operating in phase. The fact that the tops of the verticals are close together does not influence the performance to a large degree. The reason is that the current near the apex of the delta is at a minimum (it is current that takes care of radiation!).

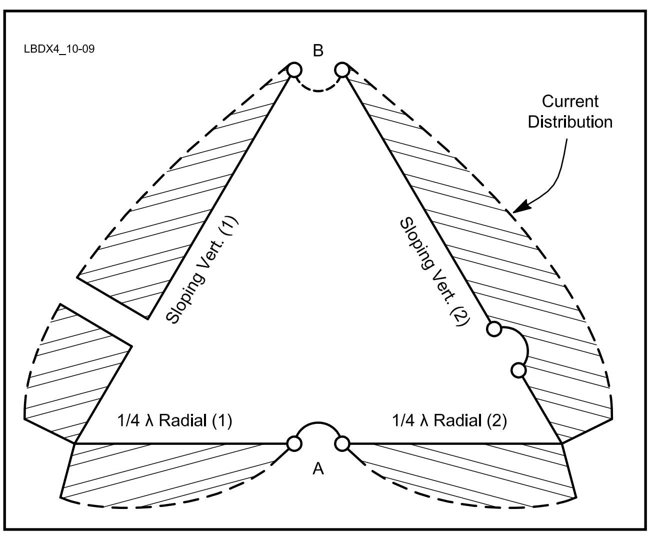


Fig 10-9 — The delta loop can be seen as two λ/4 sloping verticals, each using one radial. Because of the current distribution in the radials, the radiation from the radials is effectively canceled.

You can open the apex and move the vertical wires apart if you have a very tall support, in which case you will increase the gain of the antenna somewhat.

Considering a pair of phased verticals, we know from the study on verticals that the quality of the ground will be very important as to the efficient operation of the antenna. This does not mean that the delta loop requires radials. It has two elevated radials that are an integral part of the loop and take care of the return currents. The presence of the (lossy) ground under the antenna is responsible for near-field losses, unless we can shield it from the antenna by using a ground screen or a radial system, which should not be connected to the antenna.

As with all vertically polarized antennas, the quality of the ground within a radius of several wavelengths will determine the low-angle radiation of the loop antenna.

2.1.2. Radiation Patterns

2.1.2.1. The Equilateral Triangle

Fig 10-10 shows the configuration as well as both the broadside and the end-fire vertical radiation patterns of the vertically polarized equilateral-triangle delta loop antenna. The model was constructed for a frequency of 3.75 MHz. The baseline is 2.5 meters above ground, which puts the apex at 26.83 meters. The model was made over good ground. The delta loop shows nearly 3 dB front-to-side ratio at the main wave angle of 22°. With average ground the gain is 1.3 dBi.

2.1.2.2. The Compressed Delta Loop

Fig 10-11 shows an 80-meter delta loop with the apex at 24 meters and the baseline at 3 meters. This delta loop has a long baseline of 30.4 meters. The feed point is again located λ/4 from the apex.

The front-to-side ratio is 3.8 dB. The gain with average ground is 1.6 dBi. In free space the equilateral triangle gives a higher gain than the "flat" delta. Over real ground and in the vertically polarized mode, the gain of the flat delta loop is 0.3 dB better than the equilateral delta, however. This must be explained by the fact that the longer baseline yields a wider separation of the two "sloping" verticals, yielding a slightly higher gain.

For a 100-kHz bandwidth (on 80 meters) the SWR rises to 1.4:1 at the edges. The 2:1 SWR bandwidth is approximately 175 kHz.

Bill, W4ZV, used what he calls a "squashed" delta loop very successfully on 160 meters. The apex is 36 meters high and Bill claims that this configuration actually has improved gain over the equilateral delta loop, which can indeed be verified by accurate modeling. The antenna was also fed a quarter wavelength from the apex, using a λ/4 75-Ω matching stub. Bill says that this loop can actually be installed on a 27-meter tower by pulling the base away from the tower. By pulling the base away about 8 meters from a tower, you can actually use a full-wave delta on a 24-meter high tower, with very little trade-off.

2.1.2.3. The Bottom-Corner-Fed Delta Loop

Fig 10-12 shows the layout of the delta loop being fed at one of the two bottom corners. The antenna has the same apex and baseline height as the loop described in Section 2.1.2.2. Because of the "incorrect" location of the feed point, cancellation of radiation from the base wire (the two "radials") is not 100% effective, resulting in a significant horizontally polarized radiation component. The total field has a very uniform gain coverage (within 1 dB) from 25° to 90°. This may be a disadvantage for the rejection of high-angle signals when working DX at low wave angles.

Due to the "incorrect" feed-point location, the end-fire radiation (radiation in line with the loop) has become asymmetrical. The horizontal radiation pattern shown in Fig 10-12D is for a wave angle of 29°. Note the deep side null (nearly 12 dB) at that wave angle. The loop actually radiates its maximum signal about 18° off the broadside direction.

All this is to explain that this feed-point configuration (in the corner of the compressed loop) is to be avoided, as it really deteriorates the performance of the antenna.

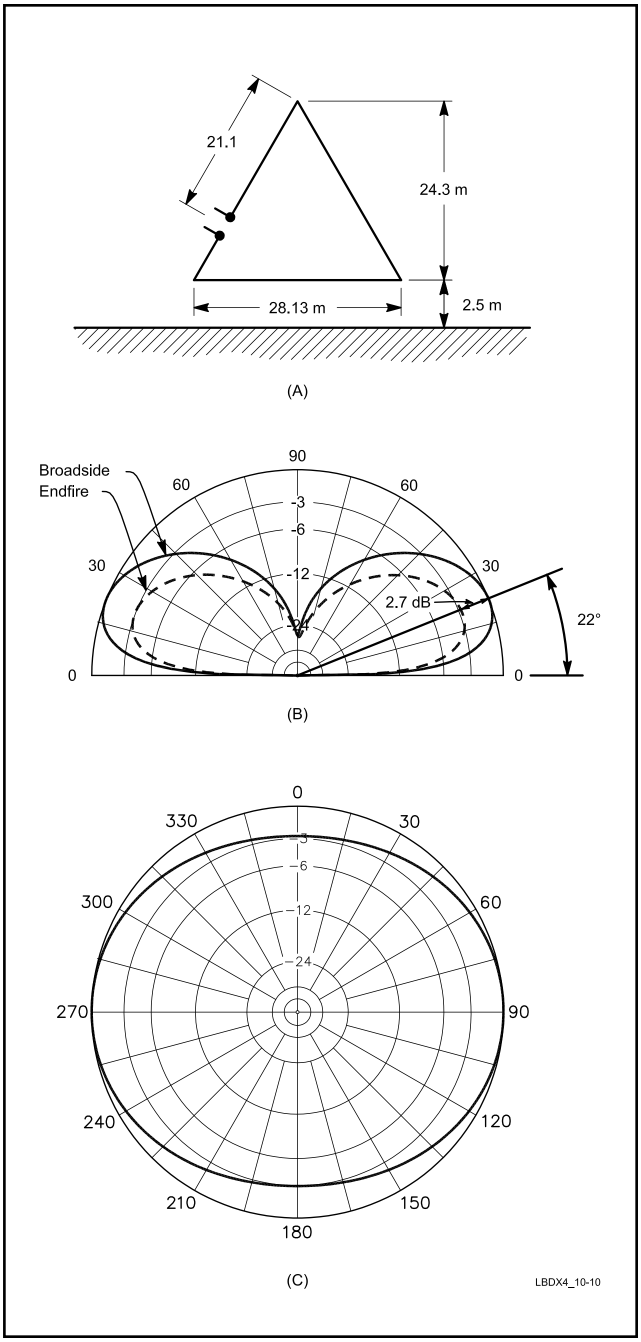


Fig 10-10 — Configuration and radiation patterns for a vertically polarized equilateral delta loop antenna. The model was calculated over good ground, for a frequency of 3.8 MHz. The elevation angle for the azimuth pattern at C is 22°.

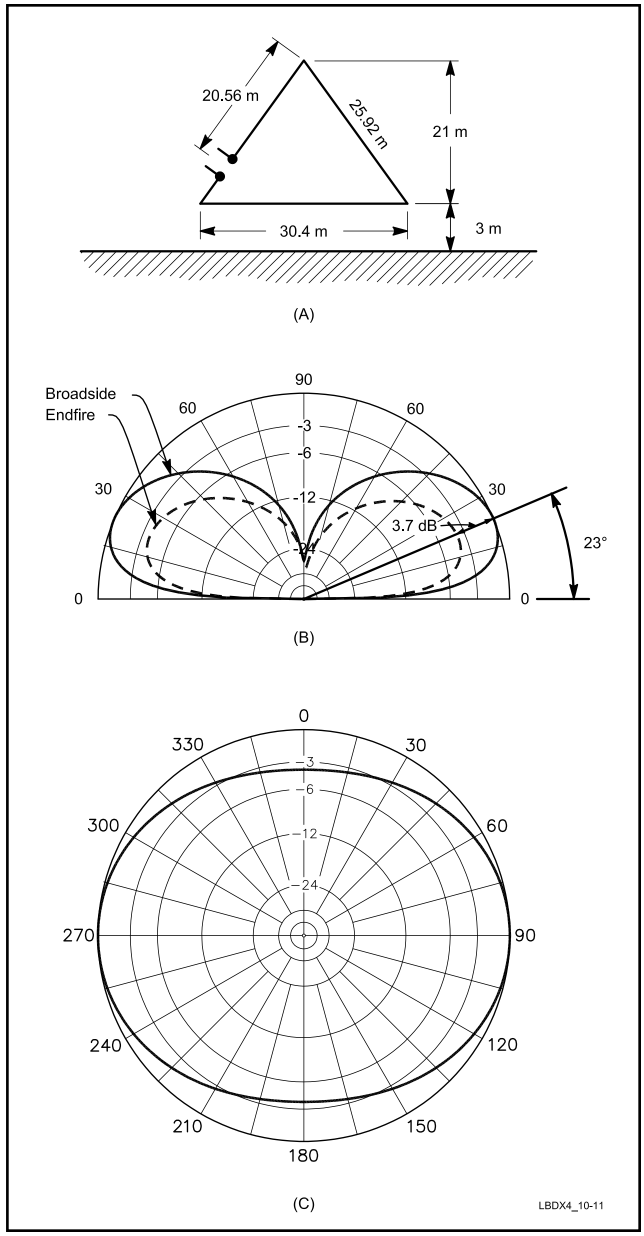


Fig 10-11 — Configuration and radiation patterns for the "compressed" delta loop, which has a baseline slightly longer than the sloping wires. The model was dimensioned for 3.8 MHz to have an apex height of 24 meters and a bottom wire height of 3 meters. Calculations are done over good ground at a frequency of 3.8 MHz. The azimuth pattern at C is for an elevation angle of 23°. Note that the correct feed point remains at $\lambda/4$ from the apex of the loop.

2.2. Horizontal Polarization

2.2.1. How it Works

In the horizontal polarization mode, the delta loop can be seen as an inverted-V dipole on top of a very low dipole with its ends bent upward to connect to the tips of the inverted V. The loop will act as any horizontally polarized antenna over real ground; its wave angle will depend on the height of the antenna over the ground.

2.2.2. Radiation Patterns

Fig 10-13 shows the vertical and the horizontal radiation patterns for an equilateral-triangle delta loop, fed at the center of the bottom wire. As anticipated, the radiation is maximum at zenith. The front-to-side ratio is around 3 dB for a 15 to 45° wave angle. Over average ground the gain is 2.5 dBi.

Looking at the pattern shape, one would be tempted to say that this antenna is no good for DX. So far we have only spoken about relative patterns. What about real gain figures from the vertically and the horizontally polarized delta loops?

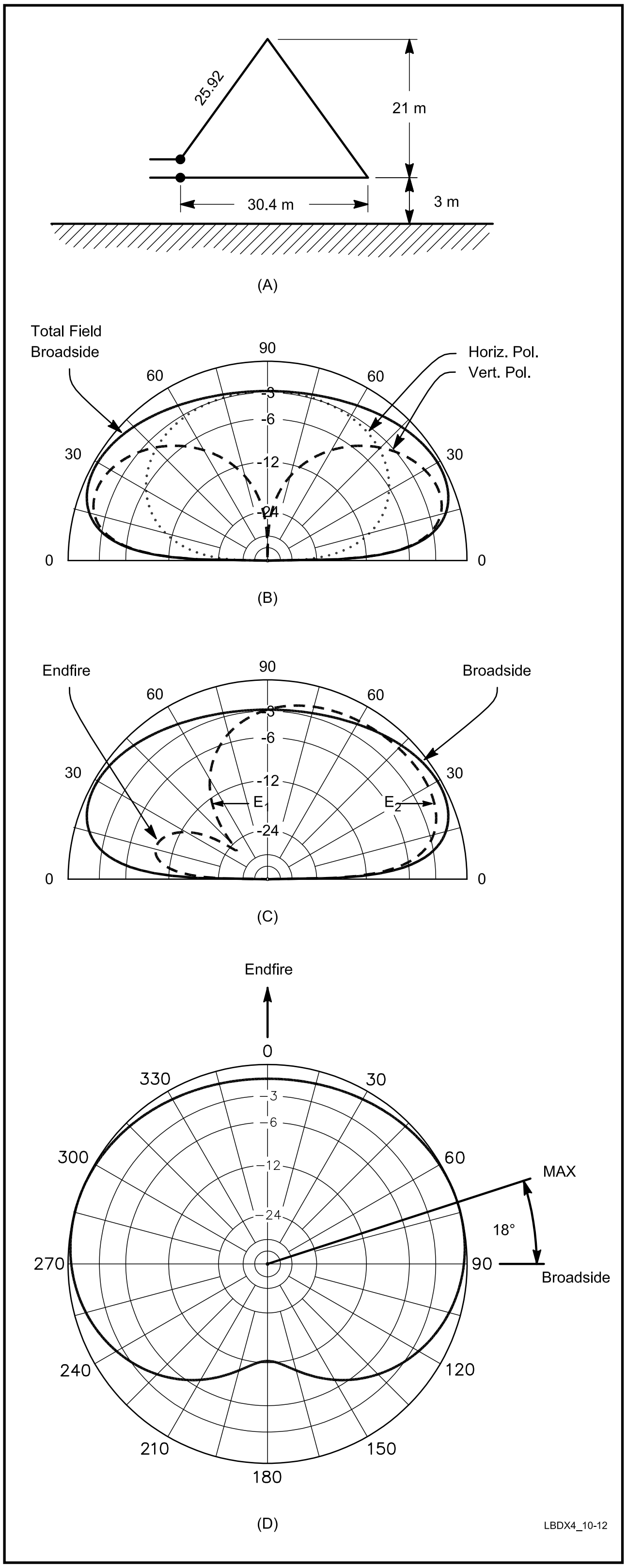


Fig 10-12 — Configuration and radiation patterns for the compressed delta loop of Fig 10-11 when fed in one of the bottom corners at a frequency of 3.75 MHz. Improper cancellation of radiation from the horizontal wire produces a strong high-angle horizontally polarized component. The delta loop now also shows a strange horizontal directivity pattern (at D), the shape of which is very sensitive to slight frequency deviations. This pattern is for an elevation angle of 29°.

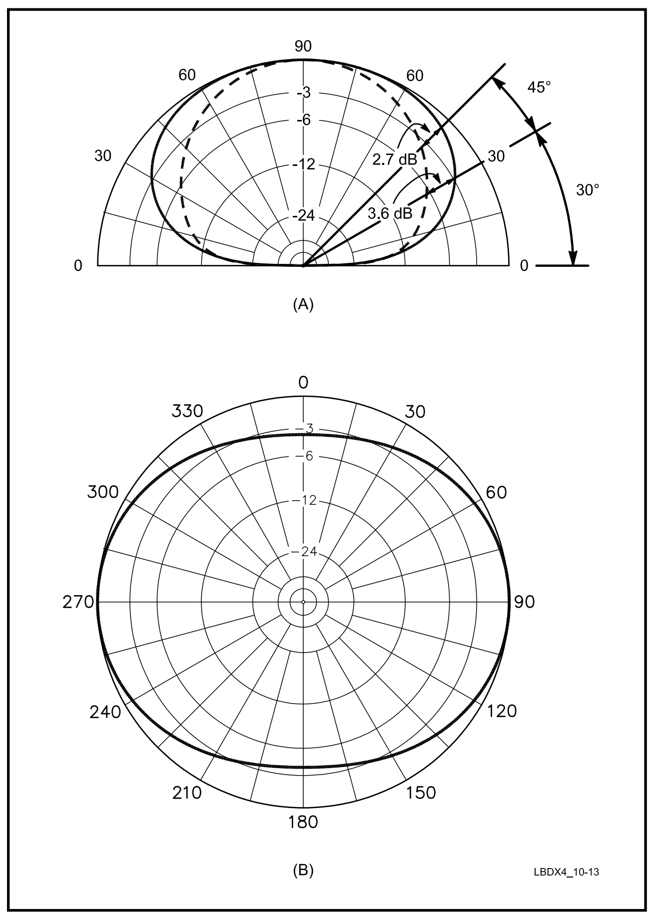


Fig 10-13 — Vertical and horizontal radiation patterns for an 80-meter equilateral delta loop fed for horizontal polarization, with the bottom wire at 3 meters. The radiation is essentially at very high angles, comparable to what can be obtained from a dipole or inverted-V dipole at the same (apex) height.

2.3. Vertical versus Horizontal Polarization

Fig 10-14 shows the superimposed elevation patterns for vertically and horizontally polarized low-height equilateral-triangle delta loops over two different types of ground (same dB scale). *MININEC*-based modeling programs cannot be used to compute the gain figures of these loops, since impedance and gain figures are incorrect for very low antenna heights.

2.3.1. Over Very Poor Ground

The horizontally polarized delta loop is better than the vertically polarized loop for all wave angles above 35°. Below 35° the vertically polarized loop takes over, but quite marginally. The maximum gain of the vertically and the horizontally polarized loops differs by only 2 dB, but the big difference is that for the horizontally polarized loop, the gain occurs at almost 90°, while for the vertically polarized loop it occurs at 25°.

One might argue that for a 30° elevation angle, the horizontally polarized loop is as good as the vertically polarized loop. It is clear, however, that the vertically polarized antenna gives good high-angle rejection (rejection against local signals), while the horizontally polarized loop will not.

2.3.2. Over Very Good Ground

The same thing that happens with any vertical happens with our vertically polarized delta: The performance at low angles is greatly improved with good ground. The vertically polarized loop is still better at any wave angle under 30° than when horizontally polarized. At a 10° radiation angle, the difference is as high as 10 dB. We have learned, in Chapter 5, that on the low bands very low angles (down to just a few degrees) are often involved, and in that respect the vertically polarized delta over good ground is far superior.

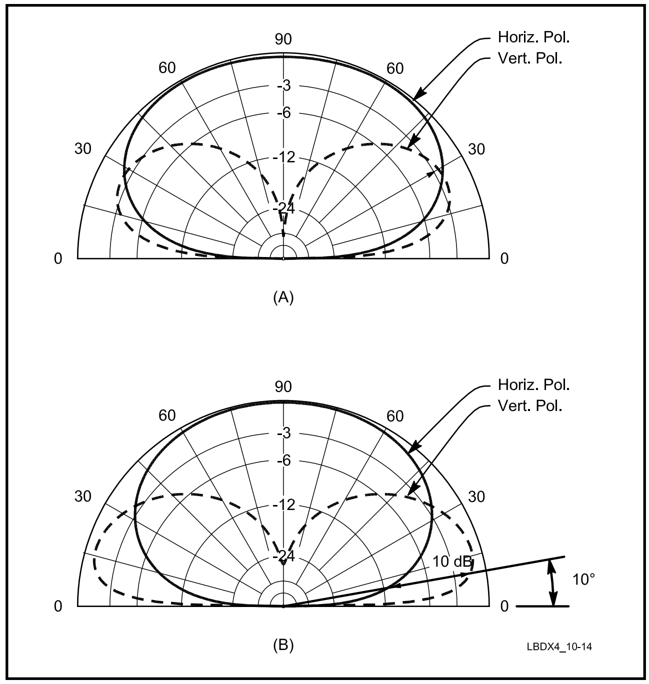


Fig 10-14 — Radiation patterns of vertically and horizontally polarized delta loops on the same dB scale. At A, over very poor ground, and at B, over very good ground. These patterns illustrate the tremendous importance of ground conductivity with vertically polarized antennas. Over better ground, the vertically polarized loop performs much better at low radiation angles, while over both good and poor ground the vertically polarized loop gives good discrimination against high-angle radiation. This is not the case for the horizontally polarized loop.

2.3.3. Conclusion

Over very poor ground, the vertically polarized loops do not provide much better low-angle radiation when compared to the horizontally polarized loops. They have the advantage of giving substantial rejection at high angles, however.

Over good ground, the vertically polarized loop will give up to 10-dB and more gain at low radiation angles as compared to the horizontally polarized loop, in addition to its high-angle rejection. See Fig 10-14B.

2.4. Dimensions

The length of the resonant delta loop is approximately 1.05 to 1.06 λ. When putting up a loop, cut the wire at 1.06 λ, check the frequency of minimum SWR (it is always the resonant frequency), and trim the length. The wavelength is given by

$$\lambda = \frac{299.8}{f\ (\text{MHz})} \qquad \text{(Eq 10-1)}$$

2.5. Feeding the Delta Loop

The feed point of the delta loop in free space is symmetrical. At high heights above ground the loop feed point is to be considered as symmetrical, especially when we feed the loop in the center of the bottom line (or at the apex), because of its full symmetry with respect to the ground.

Fig 10-15 shows the radiation resistance and reactance for both the horizontally and the vertically polarized equilateral delta loops as a function of height above ground. At low heights, when fed for vertical polarization, the feed point is to be considered as asymmetric, whereby the "cold" point is the point to which the "radials" are connected. The center conductor of a coax feed line goes to the sloping vertical section. Many users have, however, used (symmetric) open-wire line to feed the vertically polarized loop (eg, 450-Ω line).

Most practical delta loops show a feed-point impedance between 50 and 100 Ω, depending on the exact geometry and coupling to other antennas. In most cases the feed point can be reached, so it is quite easy to measure the feed-point impedance using, for example, a good-quality noise bridge connected directly to the antenna terminals. If the impedance is much higher than 100 Ω (equilateral triangle), feeding via a 450-Ω open-wire feeder may be warranted. Alternatively, you could use an unun (unbalanced-to-unbalanced) transformer, which can be made to cover a very wide range of impedance ratios (see Chapter 6 on feed lines and matching). With somewhat compressed delta loops, the feed-point impedance is usually between 50 and 100 Ω. Feeding can be done directly with a 50 or 70-Ω coaxial cable, or with a 50-Ω cable via a 70-Ω quarter-wave transformer (Zant = 100 Ω).

To keep RF off the feed line, use a balun or current choke, although the feed point of the vertically polarized delta loop is

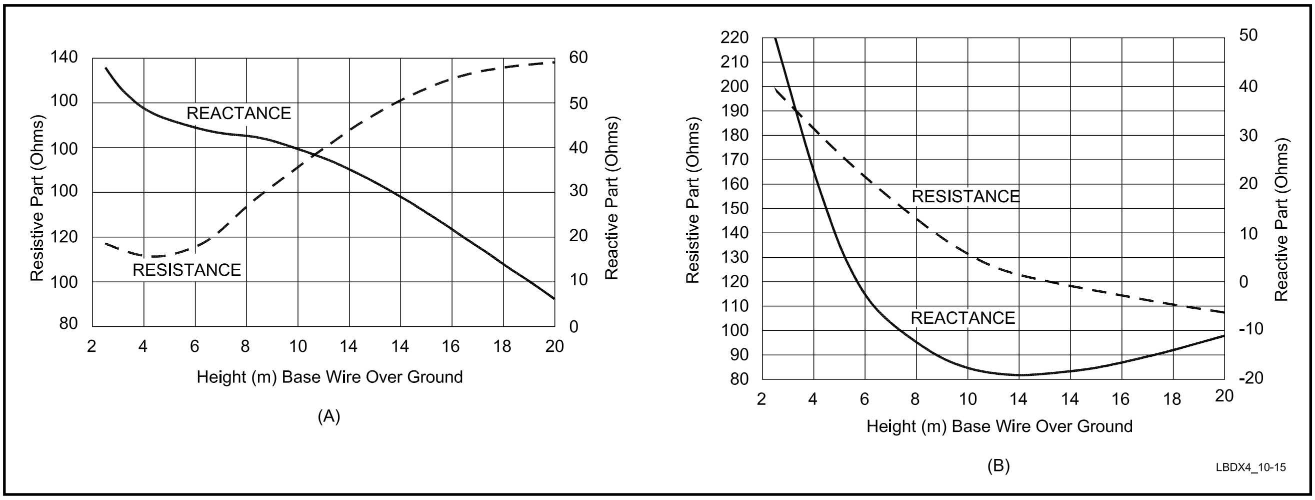


Fig 10-15 — Radiation resistance of (A) horizontally and (B) vertically polarized equilateral delta loops as a function of height above average ground. The delta loop was first dimensioned to be resonant in free space (reactance equals zero). Those dimensions were then used for calculating the impedance over real ground. Modeling was done at 3.75 MHz over good ground, using *NEC*.

not strictly symmetrical. In this case, however, we want to keep any RF current from flowing on the outside of the coaxial feed line, as these parasitic currents could upset the radiation pattern of the delta loop. For details on those baluns/common mode chokes see Chapter 8, Section 1.5. and Chapter 6, Section 7.

2.6. Gain and Radiation Angle

Fig 10-16 shows the gain and the main-lobe radiation angle for the equilateral delta loop at different heights. The values were obtained by modeling a 3.8-MHz loop over average ground using *NEC*.

Earl Cunningham, K6SE (SK), investigated different configurations of single element loops for 160 meters, and came up with the results listed in **Table 10-1** (modeling done with *EZNEC* over good ground). These data correspond surprisingly well with those shown in Fig 10-16 (where the ground was average), which explains the slight difference in gain.

2.7. Two Delta Loops at Right Angles

If the 4 to 5 dB front-to-side ratio bothers you, and if you have sufficient space, you can put up two delta loops at right angles on the same tower. You must, however, make provisions to open up the feed point of the antenna not in use, as well as its apex. This results in two non-resonant wires that do not influence the loop in use. If you would leave the unused delta loop in its connected configuration, the two loops would influence one another to a very high degree, and the results would be very disappointing.

2.8. Loop Supports

Vertically polarized loop antennas are really an array of two (sloping) verticals, each with an elevated radial. This means that if you support the delta loop from a metal tower, this tower may well influence the radiation pattern of the loop if it resonates anywhere in the vicinity of the loop. You can investigate this by modeling, but when the tower is loaded with Yagis, it is often difficult to exactly model the Yagis and their influence on the electrical length of the tower.

The safest thing you can do is to detune the tower to make sure the smallest possible current flows in the structure.

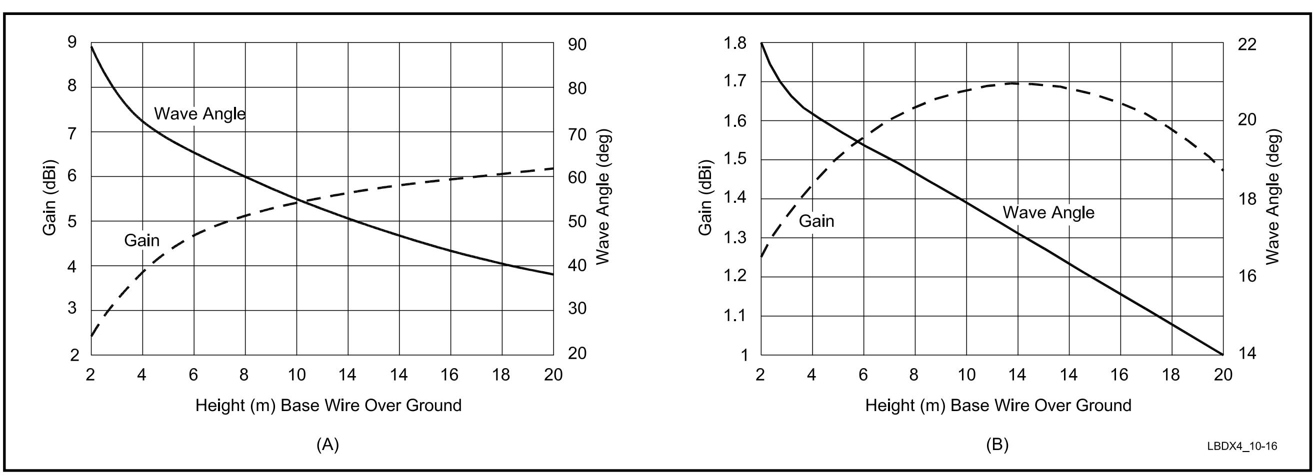
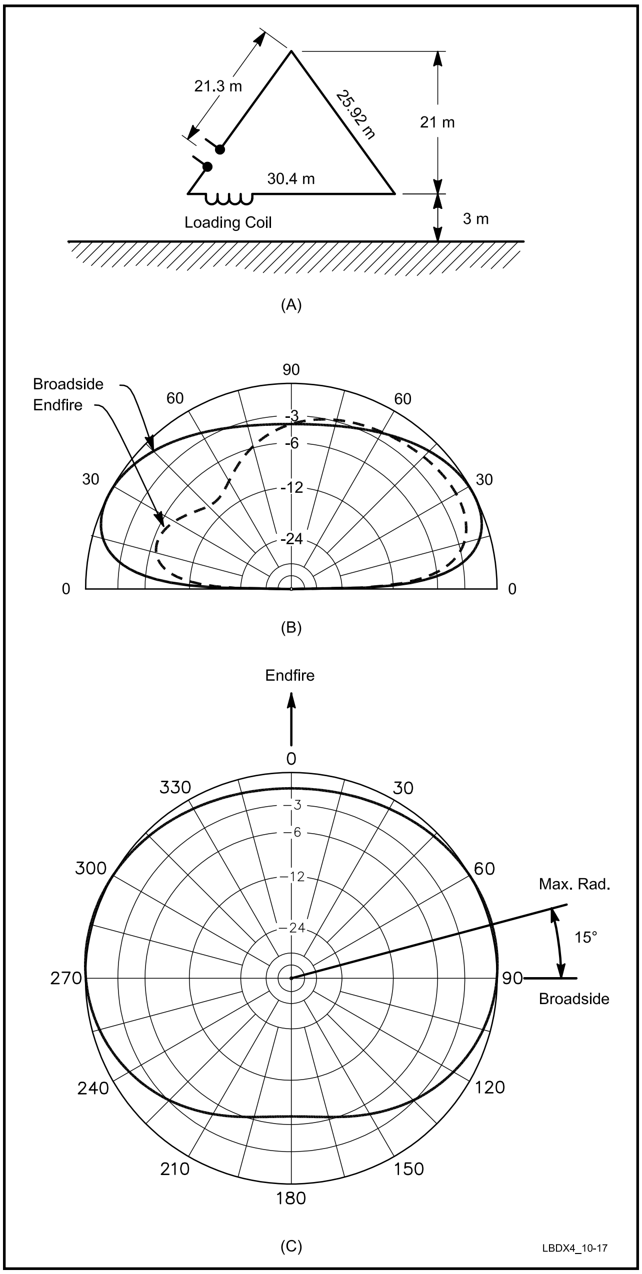


Fig 10-16 — Gain and radiation angle of (A) horizontally and (B) vertically polarized equilateral delta loops as a function of the height above ground. Modeling was done at 3.75 MHz over average ground, using *NEC*.

Table 10-1
Loop Antennas for 160 Meters

Description	*Feeding Method*	*Gain (dBi)*	*Elevation Angle (degrees)*
Diamond loop, bottom 2.5 meters high	Fed in side corner	2.15	18.0
Square loop, bottom 2.5 meters high	Fed in center of one vertical wire	2.06	20.5
Inverted equilateral delta loop (flat wire on top)	Fed λ/4 from bottom	1.91	20.9
Regular equilateral delta loop	Fed λ/4 from top	1.90	18.1

The easiest way I found to do this is to drop a wire from the top of the tower, parallel with the tower (at 0.5 to 1.0 meter distance) and terminate this wire via a 2000 pF variable capacitor to ground. Next use a current probe (such as is shown in Fig 11-17 in Chapter 11 on Vertical Arrays) and adjust the variable capacitor for *maximum* current while transmitting on the delta loop. The capacitor can be replaced with a fixed one (using a parallel combination of several values, if necessary) having the same value. This procedure will guarantee minimum mutual coupling between the loop and the supporting tower.



2.9. Modeling Loops

Loops can be modeled with *MININEC* or *NEC-2* when it comes to radiation patterns. Because of the acute angles at the corners of the delta loop, special attention must be paid to the length of the wire segments near the corners. Wire segments that are too long near wire junctions with acute angles will cause pulse overlap (the total conductor will look shorter than it actually is). The wire segments need to be short enough in order to obtain reliable impedance results. Wire segments of 20 cm length are in order for an 80-meter delta loop if accurate results are required. To limit the total number of pulses, the segment-length tapering technique can be used: The segments are shortest near the wire junction, and get gradually longer away from the junction. *ELNEC* as well as *EZNEC* have a special provision that automatically generates tapered wire segments (Ref 678).

At low heights (bottom of the antenna below approximately 0.2 λ), the gain and impedance figures obtained with a *MININEC*-based program are incorrect. The gain is too high, and the impedance too low. This is because *MININEC* calculates using a perfect ground under the antenna. Correct gain and impedance calculations at such low heights require modeling with a *NEC*-based program, such as *EZNEC*. All gain and impedance data listed in this chapter were obtained by using such a *NEC*-based modeling program.

3. LOADED LOOPS

3.1. CW and SSB 80-Meter Coverage

An 80-meter delta loop or quad loop will not cover 3.5 through 3.8 MHz with an SWR below 2:1. There are two ways to achieve wide-band coverage:

Fig 10-17 — To shift the resonant frequency of the delta loop from 3.75 MHz to 3.55 MHz, a loading coil (or stub) is inserted in one bottom corner of the loop, near the feed point (A). This has eliminated the reactive component, but has also upset the symmetrical current distribution in the bottom wire. Vertical patterns are shown at B, and the horizontal pattern is shown at C for a 27° elevation angle. As with the loop shown in Fig 10-12, high-angle radiation (horizontal component) has appeared, and the horizontal pattern exhibits a notch in the endfire direction. Maximum radiation is again slightly off from the broadside direction.

1) Feed the loop with an open-wire line (450 Ω to a matching network).
2) Use inductive or capacitive loading on the loop to lower its resonant frequency.

3.1.1. Inductive Loading

For more information about inductive loading, you can refer to the detailed treatment of short verticals in Chapter 9 on vertical antennas. There are three principles:

1) The required inductance of the loading devices (coils or stubs) to achieve a given downward shift of the resonant frequency will be minimum if the devices are inserted at the maximum current point (similar to base loading with a vertical). At the minimum current point the inductive loading devices will not have any influence. This means that for a vertically polarized delta loop, the loading coil (or stub) cannot be inserted at the apex of the loop, nor in the middle of the bottom wire.
2) Do not insert the loading devices in the radiating parts of the loop. Insert them in the part where the radiation is canceled. For example, in a vertically polarized delta loop, the loading devices should be inserted in the bottom (horizontal) wire near the corners.
3) Always keep the symmetry of the loop intact, including after having added a loading device.

From a practical (mechanical) point of view it is convenient to insert the loading coil (stub) in one of the bottom corners. **Fig 10-17A** shows the loaded, compressed delta loop (with the same physical dimensions as the loop shown in Fig 10-11), where we have inserted a loading inductance in the bottom corner near the feed point.

A coil with a reactance of 240 Ω (on 3.5 MHz) or an inductance of 10.9 μH will resonate the delta on 3.5 MHz. The 100-kHz SWR bandwidth is 1.5:1. Note again the high-angle fill in the broadside pattern (no longer symmetrical baseline configuration), as well as the asymmetrical front-to-side ratio of the loop.

Although a well-designed and well made loading coil (see Chapter 9, Section 3.3) can have a much higher Q than a linear loading stub, it sometimes is easier to quickly tune the loop with a stub. The 10.9-μH coil can be replaced with a shorted stub. The inductive reactance of the closed stub is given by:

$$X_L = Z \tan \ell \qquad \text{(Eq 10-2)}$$

where

Z = characteristic impedance of stub (transmission line)
ℓ = length of line (degrees)
X_L = inductive reactance

From this,

$$\ell = \arctan\left(\frac{X_L}{Z}\right) \qquad \text{(Eq 10-3)}$$

In our example:

$X_L = 240\ \Omega$
$Z = 450\ \Omega$
$\ell = \arctan(240/450) = 28°$

Assuming a 95% velocity factor for the transmission line, we can calculate the physical length of the stub as follows:

$$\text{Wavelength} = \frac{299.8}{3.5} = 85.66 \text{ meters (for 360°)}$$

$$\text{Physical length} = 85.66 \text{ meters} \times 0.95 \times \frac{28°}{360°} = 6.33 \text{ meters}$$

Parts B and C of Fig 10-17 show the radiation patterns resulting from the insertion of a single stub (or coil) in one of the bottom corners of the delta loop. The insertion of the single loading device has broken the symmetry in the loop,

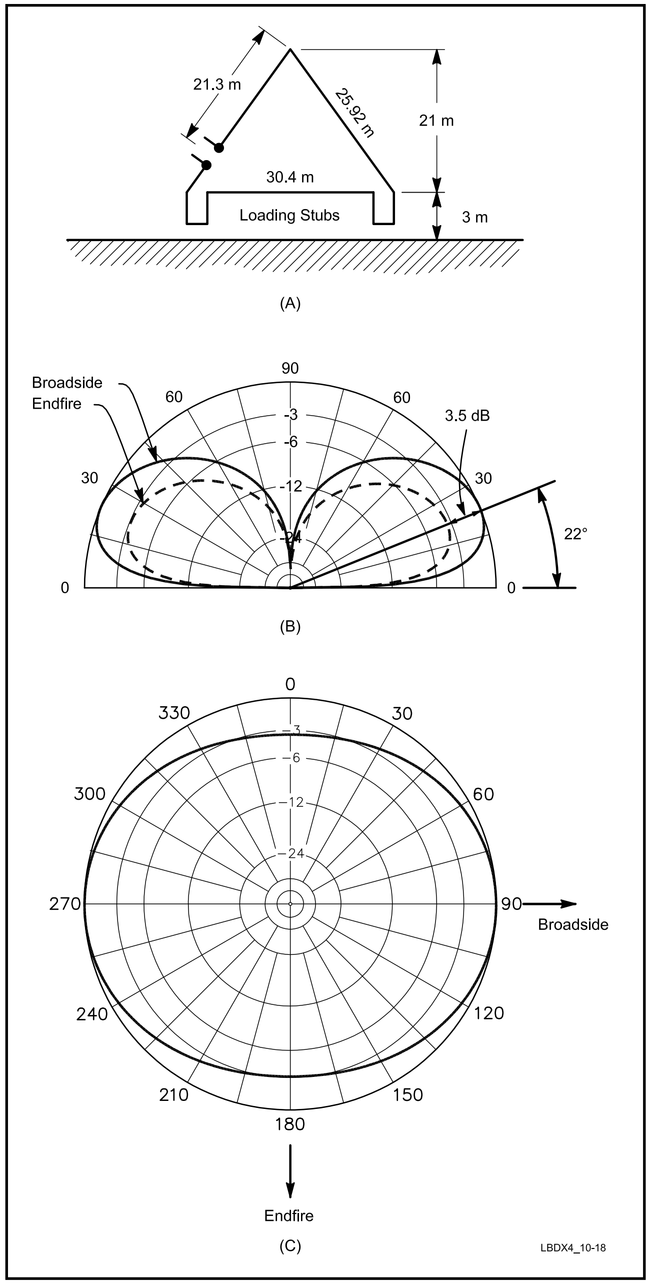


Fig 10-18 — The correct way of loading the delta loop is to insert two loading coils (or stubs), one in each bottom corner. This keeps the current distribution in the baseline symmetrical, and preserves a "clean" radiation pattern in the horizontal as well as the vertical plane. The horizontal pattern at C is for an elevation angle of 22°.

and the bottom wire now radiates as well, upsetting the pattern of the loop.

This can be avoided by using two loading coils or stubs, located symmetrically about the center of the baseline. The example in **Fig 10-18A** shows two stubs, one located in each bottom corner of the loop. Each loading device has an inductive reactance of 142 Ω. For 3.5 MHz this is:

$$\frac{142}{2\pi \times 3.5} = 6.46\ \mu\text{H}$$

A 450-Ω short circuited line is 3.96 meters long (see calculation method above). The corresponding radiation patterns in Fig 10-18 are now fully symmetrical, and the annoying high-angle radiation is totally gone. The 100-kHz SWR bandwidth is 1.45:1. The 2:1 SWR bandwidth is 170 kHz.

Fig 10-19 shows the practical arrangement that can be used for installing the switchable stubs at the two delta-loop bottom corners. A small plastic box is mounted on a piece of epoxy printed-circuit-board material that is also part of the guying system. In the high-frequency position the stub should be completely isolated from the loop. Use a good-quality open-wire line and DPDT relay preferably with ceramic insulation. The closed stub can be attached to the guy lines, which must be made of insulating material. If at all possible, make a high-Q coil, and replace the loading stub with the coil!

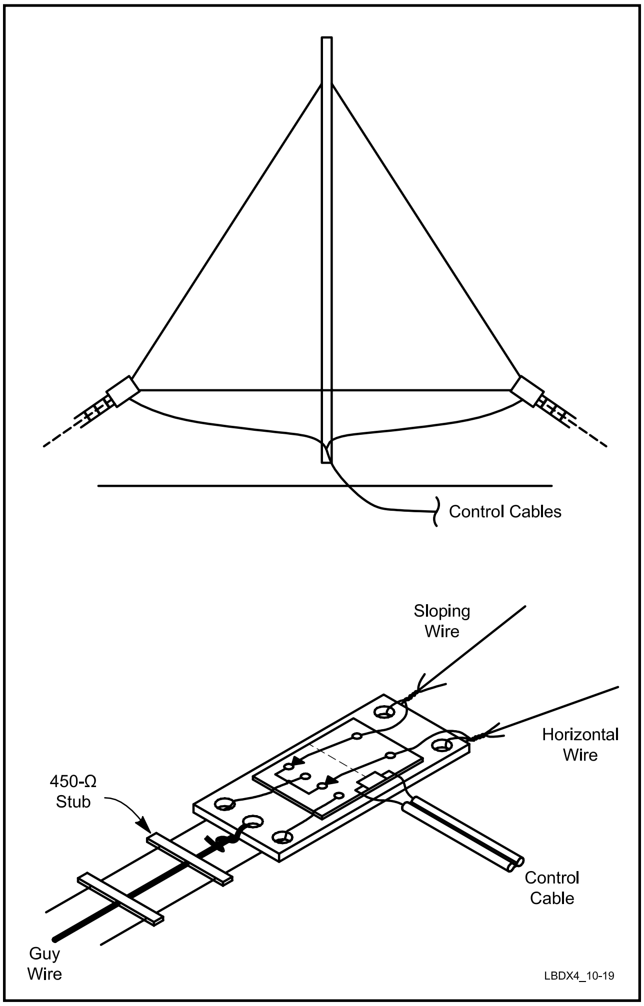


Fig 10-19 — Small plastic boxes, mounted on a piece of glass-epoxy board, are mounted at both bottom corners of the loop, and house DPDT relays for switching the stubs in and out of the circuit. The stubs can be routed along the guy lines (guy lines must be made of insulating material). The control-voltage lines for the relays can be run to a post at the center of the baseline and from there to the shack. Do not install the control lines parallel to the stubs.

3.1.2. Capacitive Loading

You can also use capacitive loading in the same way that we employ capacitive loading on a vertical. Capacitive loading is preferred over inductive loading because it is essentially lossless. Capacitive loading has the most effect when applied at a voltage antinode (also called a voltage point).

This capacitive loading is much easier to install than the inductive loading, and requires only a single-pole (high-voltage!) relay to switch the capacitance wires in or out of the circuit. *Keep the ends of the wires out of reach of people and animals, as extremely high voltage is present.*

Fig 10-20 shows different possibilities for capacitive loading on both horizontally and vertically polarized loops. If installed at the top of the delta loop as in Fig 10-20C, a

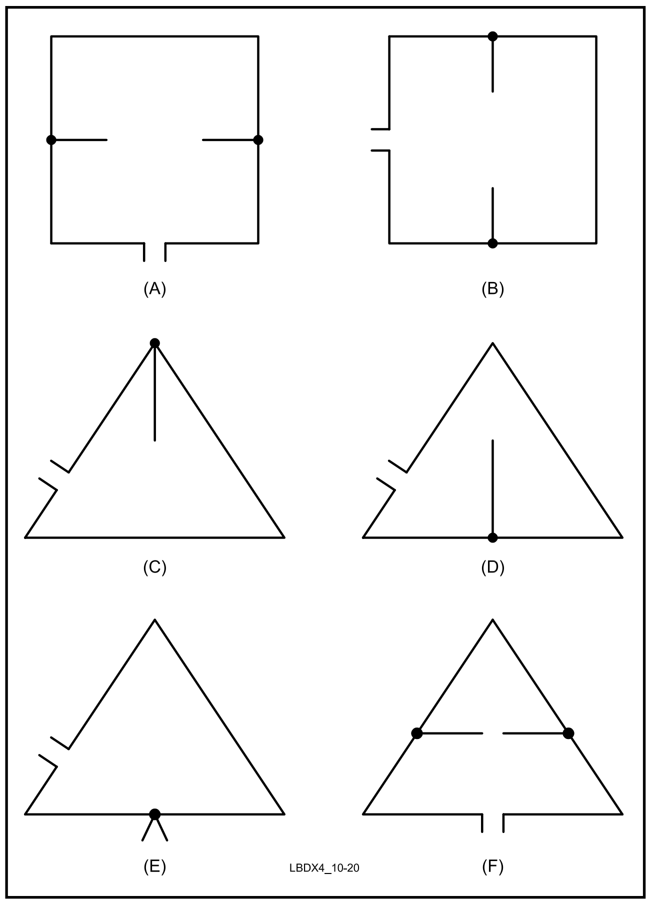


Fig 10-20 — Various loop configurations and possible capacitive loading alternatives. Capacitive loading must be applied at the voltage maximum points of the loops to have maximum effectiveness. The loading wires carry very high voltages, and good insulators should be used in their insulation.

9-meter long wire inside the loop will shift the 3.8-MHz loop (from Fig 10-11) to resonance at 3.5 MHz. For installation at the center of the baseline, you can use a single wire (Fig 10-20D), or two wires in the configuration of an inverted V (Fig 10-20E). Several wires can be connected in parallel to increase the capacitance. (*Watch out, since there is very high voltage on those wires while transmitting!* As the loading devices are applied at voltage maxima, switching the loading wire(s) in and out requires relays that can take very high voltages. Current will not be the issue.)

The same symmetry guidelines should be applied as explained in Section 3.1 to preserve symmetrical current distribution.

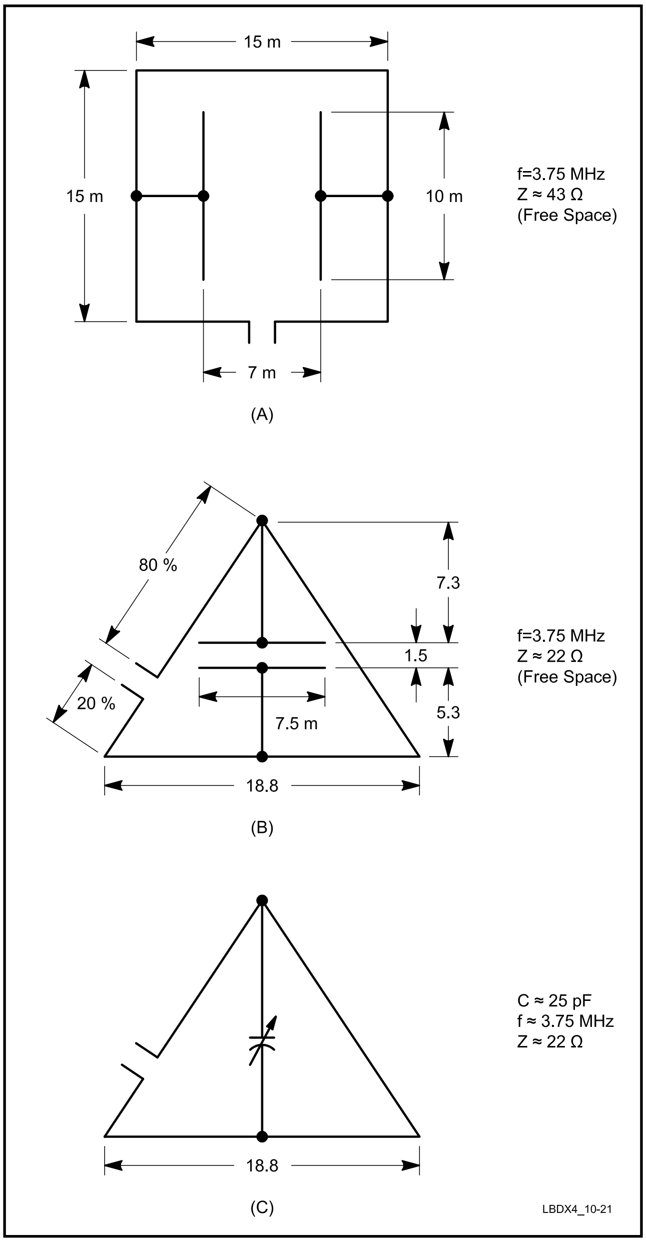


Fig 10-21 — Capacitive loading can be used on loops of approximately 2/3 full size. See text for details.

3.1.3. Adjustment

Once the loop has been trimmed for resonance at the high-frequency end of the band, just attach a length of wire with a clip at the voltage point and check the SWR to see how much the resonant frequency has been lowered. It should not take you more than a few iterations to determine the correct wire length. If a single wire turns out to require too much length, connect two or more wires in parallel, and fan out the wire ends to create a higher capacitance.

3.1.4. Bandwidth

By using one of the above-mentioned loading methods and a switching arrangement, a loop can be made that covers the entire 80-meter band with an SWR below 2:1.

3.2. Reduced-Size Transmitting Loop Antennas

Reduced-size loops have been described in amateur literature (Refs 1115, 1116, 1121, 1129). **Fig 10-21** shows some of the possibilities of applying capacitive loading to loops, whereby a substantial shift in frequency can be obtained. G3FPQ uses a reduced-size 2-element 80-meter quad that makes use of capacitive-loaded square elements as shown in Fig 10-21A. The fiberglass spreaders of the quad support the loading wires.

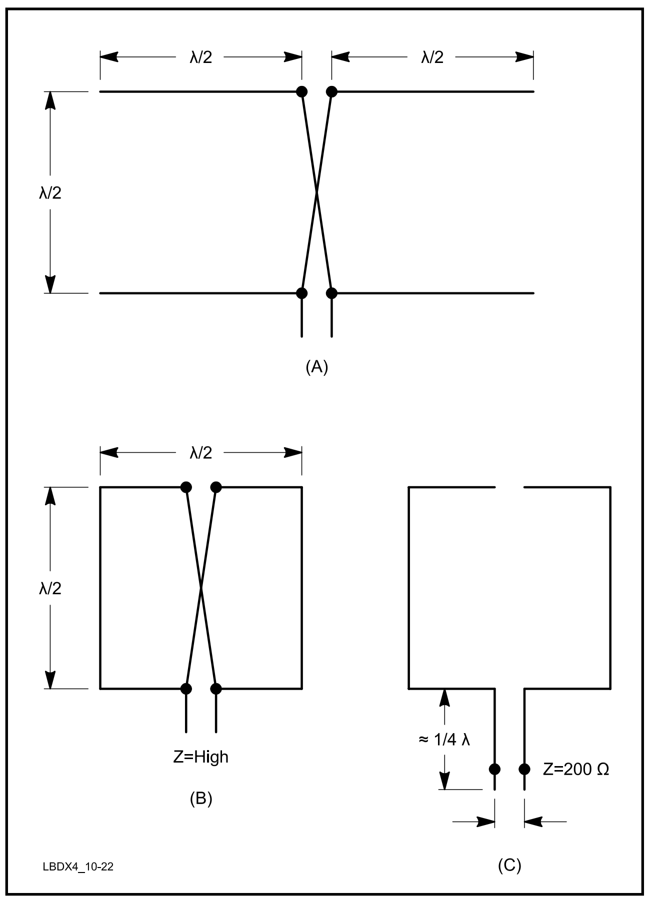


Fig 10-22 — The bi-square antenna is a lazy-H antenna (two λ/2 collinear dipoles, stacked λ/2 apart and fed in phase), with the ends of the dipoles bent down (or up) and connected. The feed-point impedance is high and the array can best be fed via a λ/4 stub arrangement.

It is possible to lower the frequency by a factor of 1.5 with this method, without lowering the radiation resistance to an unacceptable value (a loop dimensioned for 5.7 MHz can be loaded down to 3.8 MHz). The triangular loop can also be loaded in the same way, although the mechanical construction may be more complicated than with the square loop. See Fig 10-21B.

In principle, we can replace the parallel wires with a (variable) capacitor. This would allow us to tune the loop. The example in Fig 10-21C requires approximately 30 pF to shift the antenna from 5.7 to 3.8 MHz. Beware, however, that extremely high voltages exist across the capacitor. It would certainly not be over-engineering to use a 50-kV or higher capacitor for the application.

4. BI-SQUARE

The bi-square antenna has a circumference of 2 λ and is opened at a point opposite the feed point. A quad antenna can be considered as a pair of shortened dipoles with λ/4 spacing. In a similar way, the bi-square can be considered as a lazy-H antenna with the ends folded vertically, as shown in **Fig 10-22**. Not many people are able to erect a bi-square antenna, as the dimensions involved on the low bands are quite large.

In free space the bi-square has 3-dB gain over two λ/2 dipoles in phase (collinear), and almost 5 dB over a single λ/2 dipole. Over real ground, with the bottom wire λ/8 above ground (10 meters for an 80-meter bi-square), the gain of the bi-square is the same as for the two λ/2 dipoles in phase. The bottom two λ/2 sections do not contribute to low-angle radiation of the antenna.

The bi-square has the advantage over two half-waves in phase that the antenna does not exhibit the major high-angle sidelobe that is present with the collinear antenna when the height is over λ/2. **Fig 10-23** shows the radiation patterns of the bi-square and the collinear with the top of the antenna 5λ/8 high. Notice the cleaner low-angle pattern of the bi-square. Of course you could obtain almost the same result by lowering the collinear from 5λ/8 to λ/2 high!

The bi-square can be raised even higher in order to further reduce the wave angle without introducing high-angle lobes, up to a top height of 2 λ. At that height the wave angle is 14°, without any secondary high-angle lobe. With the top at 5λ/8, the takeoff angle is 26°.

To exploit the advantages of the bi-square antenna, you need quite impressive heights on the low bands. N7UA is one of the few stations using such an antenna, and he produces a most impressive signal on the long path into Europe on 80 meters. With a proper switching arrangement, the antenna can be made to operate as a full-wave loop on half the frequency (eg, 160 meters for an 80-meter bi-square).

The feed-point impedance is high (a few thousand ohms), and the recommended feed system consists of 600-Ω line with a stub to obtain a 200-Ω feed point. By using a 4:1 balun, a

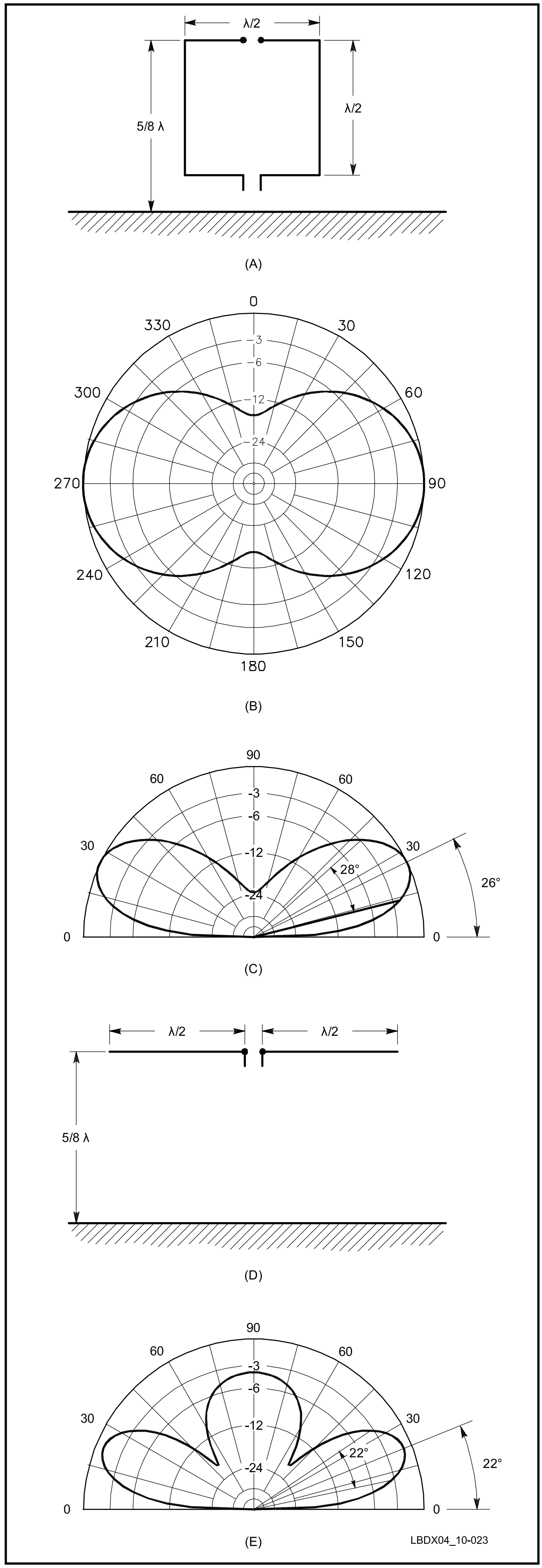


Fig 10-23 — The bi-square antenna (A) and its radiation patterns (B and C). The azimuth pattern at B is for an elevation angle of 25°. At D, two half waves in phase and at E, its radiation pattern. Note that for a top-wire height of 5λ/8, the bi-square does not exhibit the annoying high-angle lobe of the collinear antenna.

coaxial cable can be run from that point to the shack. Another alternative is to run the 600-Ω line all the way to the shack into an open-wire antenna tuner.

5. THE HALF LOOP

The half loop was first described by Belrose, VE2CV (Ref 1120 and 1130). This antenna, unlike the half sloper, cannot be mounted on a tall tower supporting a quad or Yagi. If this were done, the half loop would shunt-feed RF to the tower and the radiation pattern would be upset. This can be avoided by decoupling the tower using a λ/4 stub (Ref 1130). The half loop as shown in **Fig 10-24** can be fed in different ways.

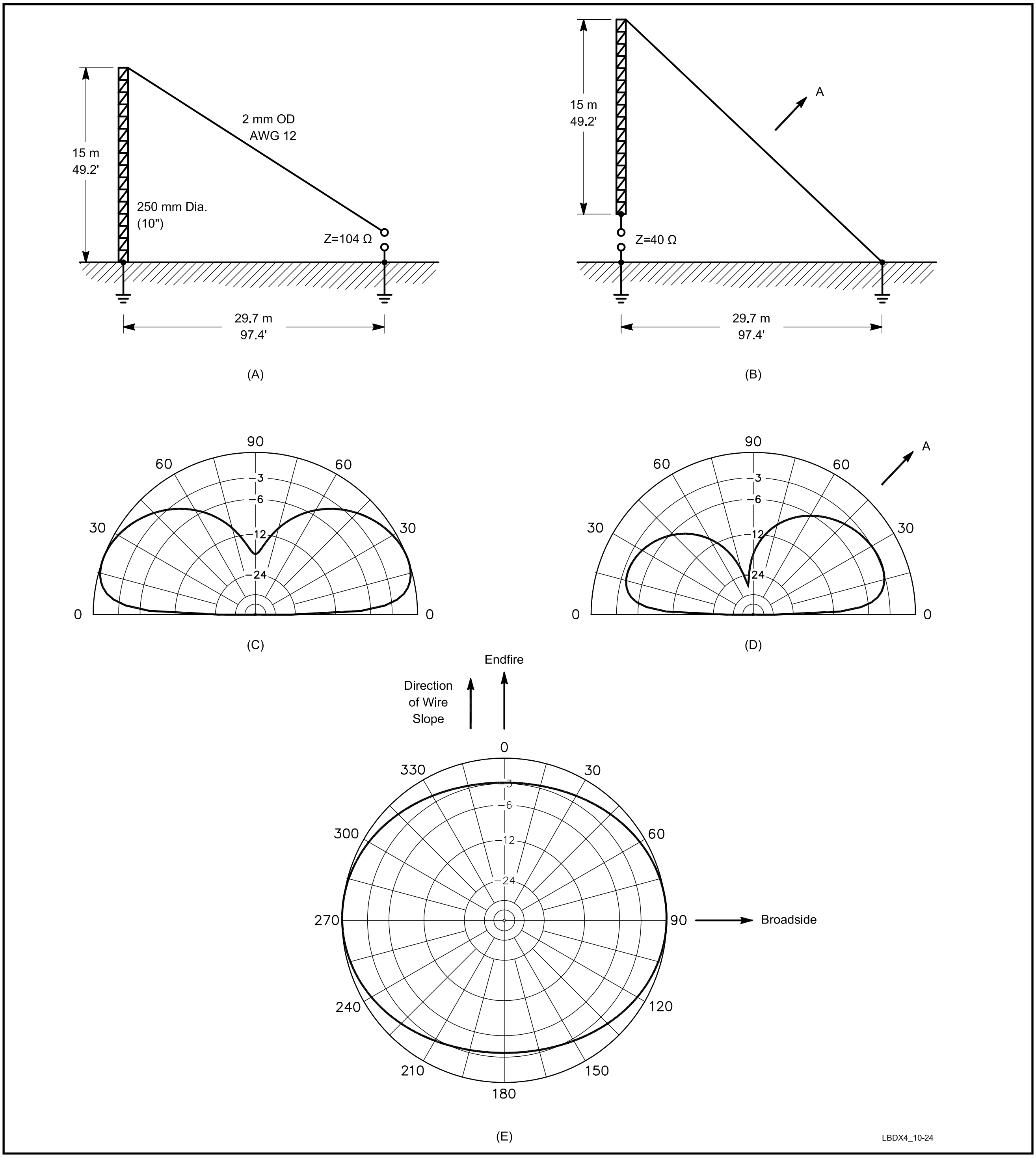


Fig 10-24 — Half-loop antenna for 3.75 MHz, fed for low-angle radiation. The antenna can be fed at either end against ground (A and B). The grounded end must be connected to a good ground system, as must the ground-return conductor of the feeder. Radials are essential for proper operation. Note that while the feed-point locations are different, the radiation patterns do not change. C shows the broadside vertical pattern, D is the end-fire vertical pattern, and E is the azimuth pattern for an elevation angle of 20°.

5.1. The Low-Angle Half Loop

For low-angle radiation, the feed point can be at the end of the sloping wire (with the tower grounded), or else at the base of the tower (with the end of the sloping wire grounded). The radiation pattern in both cases is identical. The front-to-side ratio is approximately 3 dB, and the antenna radiates best in the broadside direction (the direction perpendicular to the plane containing the vertical and the sloping wire).

There is some pattern distortion in the end-fire direction, but the horizontal radiation pattern is fairly omnidirectional. Most of the radiation is vertically polarized, so the antenna requires a good ground and radial system, as for any vertical antenna. As such, the half loop does not really belong to the family of large loop antennas, but as it is derived from the full-size loop, it is treated in this chapter rather than as a top-loaded short vertical.

The exact resonant frequency depends to a great extent on the ratio of the diameter of the vertical mast to the slant wire. The dimensions shown in Fig 10-24 are only indicative. Fine-tuning the dimensions will have to be done in the field.

5.2. The High-Angle Half Loop

The half delta loop antenna can also be used as a high-angle antenna. In that case you must isolate the tower section from the ground (use a good insulator because it now will be at a high-impedance point) and feed the end of the sloping wire. Alternatively, you can feed the antenna between the end of the sloping wire and ground, while insulating the bottom of the tower from ground. Using the same dimensions that made the low-angle version resonant no longer produces resonance in these configurations.

Fig 10-25 shows the low-angle configurations with the radiation patterns. Note that the alternative where the end of the slant wire is fed against ground produces much more high-angle radiation than the alternative where the bottom end of the tower is fed. In both cases, the other end of the aerial is left floating (not connected to ground).

Dimensional configurations other than those shown in the relevant figures can be used as well, such as with a higher tower section and a shorter slant wire. If you move the end of the sloping wire farther away from the tower, you will need to decrease the height of the tower to keep resonance, and the radiation resistance will decrease. This will, of course, adversely influence the efficiency of the antenna. If the bottom of the sloping wire is moved toward the tower, the length of the vertical will have to be increased to preserve resonance. When

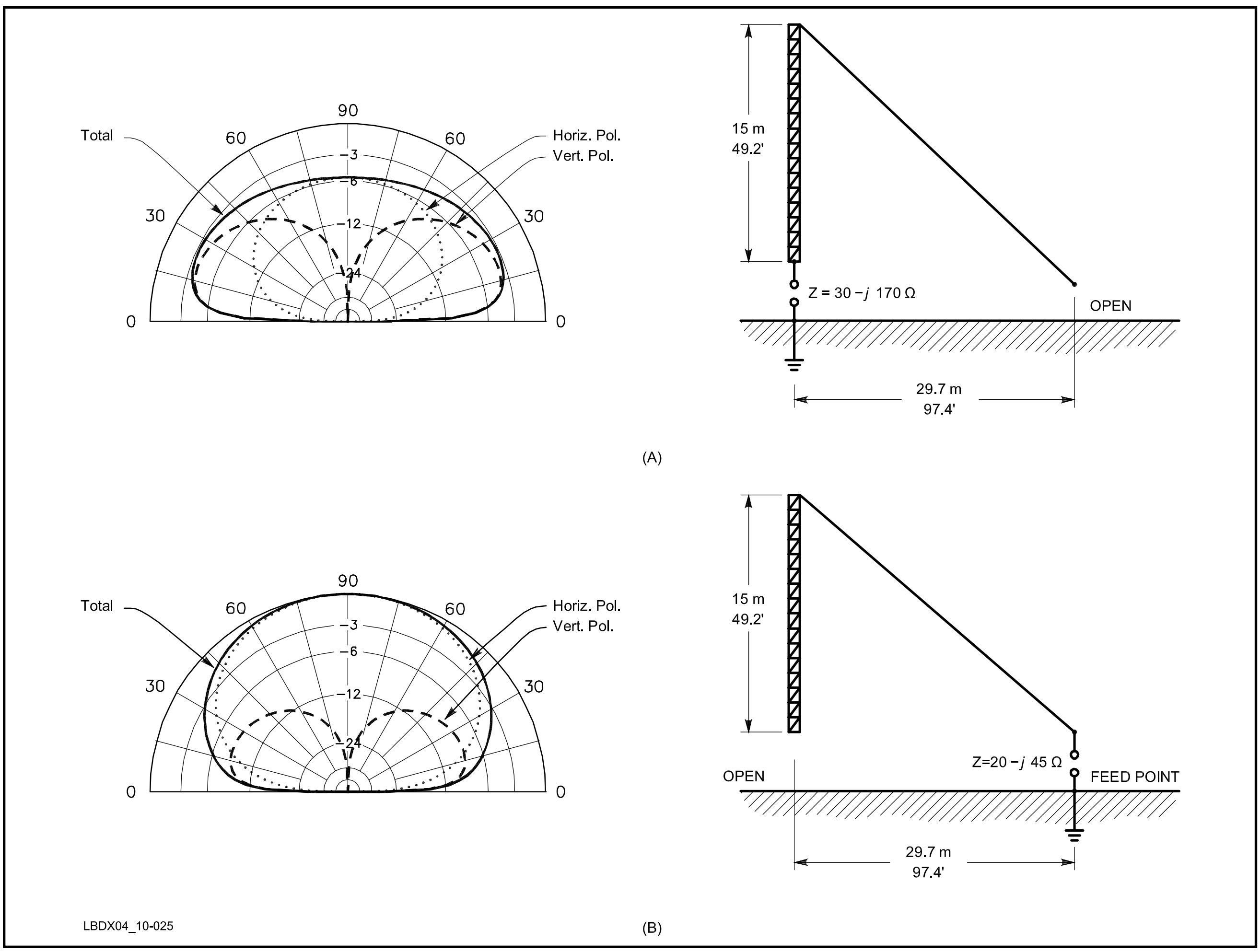


Fig 10-25 — High-angle versions of the half delta loop antenna for 3.75 MHz. As with the low-angle version, the antenna can be fed at either end (against ground). The other end, however, must be left floating. The two different feed points produce different high-angle patterns as well as different feed-point impedances.

the end of the sloping wire has been moved all the way to the base of the tower we have a λ/4 vertical with a folded feed system. The feed-point impedance will depend on the spacing and the ratio of the tower diameter to the feed-wire diameter.

Compared to a loaded vertical, this antenna has the advantage of giving the added possibility for switching to a high-angle configuration. For a given height, the radiation resistance is slightly higher than for the top-loaded vertical, whereby there is no radiation from the top load. The sloping wire in this half-loop configuration adds somewhat to the vertical radiation, hence the increase (10% to 15%) in R_{rad}.

Being able to feed the antenna at the end of the sloping wire may also be an advantage: This point may be located at the transmitter location, so the sloping wire can be directly connected to an antenna tuner. This would enable wide-band coverage by simply retuning the antenna tuner. Switching from a high to a low-angle antenna in that case consists of shorting the base of the tower to ground (for low-angle radiation).

6. THE HALF SLOPER

Although the so-called half sloper of **Fig 10-26A** may look like a half delta, it really does not belong with the loop antennas. As we will see, it is rather a loaded vertical with a specific matching system and current distribution.

Quarter-wave slopers are the typical result of ham ingenuity and inventiveness. Many DXers, short of space for putting up large, proven low-angle radiators, have found their half slopers to be good performers. Of course they don't know how much better other antennas might be, as they have no room to try them. Others have reported that they could not get their half sloper to resonate on the desired frequency (that's because they gave up trying before having found the proverbial needle in the haystack). Of course resonating and radiating are two completely different things. It's not because you cannot make the antenna resonant that it will not radiate well. Maybe they need a matching network?

To make a long story short, half slopers seem to be very unpredictable. There are a large number of parameters (different tower heights, different tower loading, different slope angles, and so forth) that determine the resonant frequency and the feed-point impedance of the sloper.

Unlike the half delta loop, the half sloper is a very difficult antenna to analyze from a generic point of view, as each half sloper is different from any other. Belrose, VE2CV, thoroughly analyzed the half sloper using scale models on a professional test range (Ref 647). His findings were confirmed by DeMaw, W1FB (Ref 650). Earlier, Atchley, W1CF, reported outstanding performance from his half sloper on 160 meters (Ref 645).

I have modeled an 80-meter half sloper using *MININEC*. After many hours of studying the influence of varying the many parameters (tower height, size of the top load, height of the attachment point, length of the sloper, angle of the sloper,

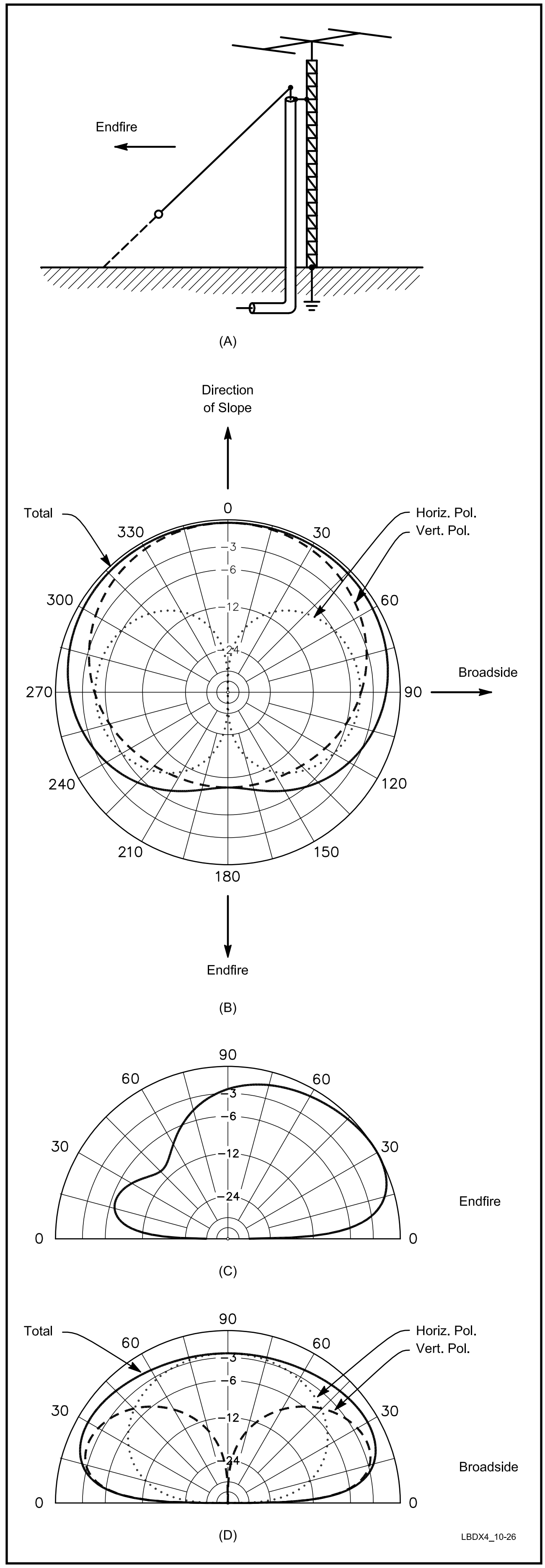


Fig 10-26 — At A, a half-sloper mounted on an 18-meter tower that supports a 3-element full-size 20-meter Yagi. See text for details. At B, the azimuth pattern for a 45°elevation angle with vertically and horizontally polarized components, and at C and D, elevation patterns. The antenna shows a modest F/B ratio in the end-fire direction at a 45° elevation angle.

ground characteristics, etc), I came to the following conclusions:

- The so-called half sloper is made up of a vertical and a slant wire. Both contribute to the radiation pattern. The radiation pattern is essentially omnidirectional. The low-angle radiation comes from the loaded tower, the high-angle radiation from the horizontal component of the slant wire. The antenna radiates a lot of high-angle signal (coming from the slant wire).
- Over poor ground the antenna has some front-to-back advantage in the direction of the slope, ranging from 10 to 15 dB at certain wave angles. Over good and excellent ground the F/B ratio is not more than a few dB.

An interesting testimony was sent on the Internet by Rys, SP5EWY, who wrote "Well, I had previously used my tower without radials and the half sloper favored the South, with the wire sloping in that direction, by at least 1 S-unit. Later I added 20 radials and since then it seems to radiate equally well in all directions."

In essence the half sloper is a top-loaded vertical, which is fed at a point along the tower where the combination of the tower impedance and the impedance presented by the sloping wire combine to a 50-Ω impedance (at least that's what we want). The sloping wire also acts as a sort of radial to which the other conductor of the feed line is connected (like radials on a vertical to push against). In other words, the sloping wire is only a minor part of the antenna, a part that helps to create resonance as well as to match the feed line. Belrose (Ref 647) also recognized that the half sloper is effectively a top-loaded vertical. Fig 10-26 (B through D) shows the typical radiation patterns obtained with a half-sloper antenna.

While modeling the antenna, it was very critical to find a point on the tower and a sloper length and angle that give a good match to a 50-Ω line. The attachment point on the tower need not be at the top. It is not important how high it is, as you are not really interested in the radiation from the slant wire.

Changing the attachment point and the sloper length does not appreciably change the radiation pattern. This indicates that it is the tower (capacitively loaded with the Yagi) that does the bulk of the radiating. As the antenna mainly produces a vertically polarized wave, it requires a good ground system, at least as far as its performance as a low-angle radiator is concerned.

From my experience in spending a few nights modeling half slopers, I would highly recommend that any prospective user first model the antenna using *EZNEC*, which is great for such a purpose and which has the most user-friendly interfaces for multiple iterations.

There is an interesting analysis by D. DeMaw (Ref 650). DeMaw correctly points out that the antenna requires a metal support, and that a tree or a wooden mast will not do. But he does not emphasize anywhere in his study that it is the metal support that is responsible for most of the desirable low-angle radiation. DeMaw, however, recognizes the necessity of a good ground system on the tower, which implicitly admits that the tower does the radiating. DeMaw also says, "The antenna is not resonant at the operating frequency," by which he means that the slant wire is not a quarter-wave long. This is again very confusing, as it seems to indicate that the slant wire is the antenna, which it is not. Describing his on-the-air results, DeMaw confirms what we have modeled: due to the presence of high-angle radiation, it outperforms the vertical for short and medium-range contacts, while the vertical takes over at low angles for real DX contacts.

To summarize the performance of half slopers, it is worthwhile to note Belrose's comment, "If I had a single quarter-wave tower, I'd employ a full-wave delta loop, apex up, lower-corner fed, the best DX-type antenna I have modeled." Of course, a delta loop still has a baseline of approximately 100 feet (on the 80-meter band), which is not the case with the half sloper. But the half sloper, like any vertical, requires radials in order to work well. It may look like the half sloper has a space advantage over many other low-band antennas, but this is only as true as for any vertical.

Contents

CHAPTER 11

Phased Arrays

Robye Lahlum, W1MK

Pekka Ketonen, OH1TV

Greg Ordy, W8WWV

Some seven years ago I came in contact with Robye Lahlum, W1MK. At the time, I would never have thought that he would become literally my right hand for this 5th and final edition of *Low Band DXing*. Robye has been my master assistant, my most meticulous proofreader, a most creative inspirer, a much appreciated critic and a very good friend as well.

Large parts of Chapters 7 and 11 in 4th edition of this book thrive on the original ideas and the work of Robye. And it is no different for this final edition.

When Robye saw the new feed system developed by OH1TV (opposite voltage feed system), he wrote all the mathematics defining the system in one day and developed a tuning procedure for it as well. I think Robye is hard to beat in matters of antennas. He keeps surprising me. The nice thing about Robye is that he's not only very knowledgeable about the theoretical (mathematical) aspects of antennas; he also builds them and gets on the air with them. In wintertime I regularly work him on 160 or 80 meter CW, and the signal he puts into Europe shows that he knows what he's doing!

After having worked together with Robye on the previous issue of this book, I knew I wanted more joint efforts, and that's why I asked Robye to be the godfather of this chapter as well as my guide for Chapter 7 on Receiving Antennas. Please read Robye's resume on the opening page of Chapter 7.

Robye's involvement in this chapter has unveiled a number of mysteries on the subject of hybrid coupler feed systems for arrays. At one time I asked him, "Could you develop a black box model for the 90° hybrid network and write all the mathematical equations that are involved?" One week later it was done and ready for me to develop what has become a very practical spreadsheet calculation tool that tells you everything about the once-so-mysterious hybrid coupler. Now that we know how it all works, it is a small step to develop systems that optimize the 90°-hybrid-based feed systems to perform exactly as they should. Modeling tools are here for the readers to use to help build top-performing arrays. All of this is highly original and exclusive, and published for the first time ever in this book.

It's been great working together on a project like this with Robye. His technical knowledge is profound, his original ideas abundant and his engagement in a project like this is very stimulating and rewarding.

Thank you Robye for your encouragements, your support, your suggestions, your help and your friendship.

I would like to express my very sincere thanks to two eminent hams who have also been very helpful with this chapter. Pekka Ketonen, OH1TV, the designer of the 160-meter 3-element Yagi at OH8X (see Chapter 13, Section 3.11) suggested a novel way of feeding vertical arrays, using the "voltage forcing" or "opposite voltage feed system" (see Section 3.4.9). Greg Ordy,W8WWV, one of the hard working antenna gurus at K3LR's super contest station, came up with an idea to improve the performance of Four Square arrays fed with a hybrid coupler (Section 3.4.6.7). He is also the author of *LBDXView*, a software program included on this book's CD that allows you to print out (as lists, graphs or patterns) the results of your swept-frequency modeling or swept-frequency measurements done with a multiplexed VNA measurement system.

These two fine gentlemen have been willing to share with the readers some of the highly original developments in the field of array engineering that they have been responsible for. Thank you Pekka and Greg!

I also would like to say a special word of thanks and appreciation to Roy Lewallen, W7EL, who was willing to help us with producing a model of the hybrid coupler that can be included in an antenna model. This is a first, and was a very welcome help in demystifying the once-so-mysterious hybrid coupler.

If you want gain and directivity on one of the low bands and if you live in an area with good or excellent ground, an array made of vertical elements may be the answer — provided you have room for it.

Arrays made with vertical elements have the same requirements as single vertical antennas as far as ground quality is concerned. Before you decide to put one up, take the time to understand the mechanism of an array where all elements are fed.

In this chapter I cover the subject of arrays made of elements that, by themselves, have an omnidirectional horizontal radiation pattern: vertical antennas.

1. RADIATION PATTERNS

1.1. How the Pattern is Formed

In **Fig 11-1** we are looking down on two verticals from above. The array made up by the two verticals is characterized by the spacing between the verticals and by the feed current in each of the verticals. The paper represents the ground and both radiators are omnidirectional by definition. We are trying to construct the horizontal radiation pattern of the array. Rays from each element have a phase difference that depends on three factors:

1) Spacing between the elements.
2) The phase difference with which RF current is applied at the feed point of each element.
3) Angle of ray a and a' with respect to the line connecting the two elements.

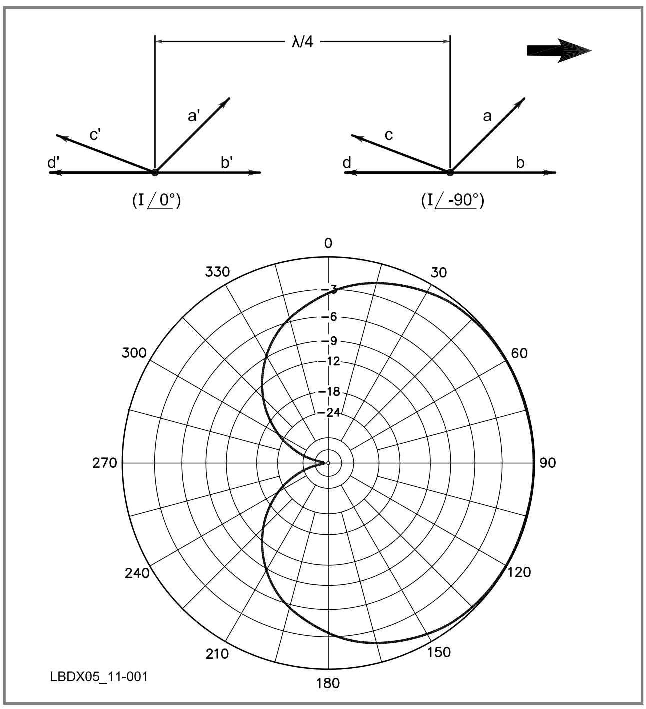


Fig 11-1 — Both array elements are fed with the same current magnitude. The front element is fed with a current lagging by 90° (I = 1∠–90°). Graphic analysis of a few rays shows that the array will radiate most power in the direction of rays b and b' where these rays, because of physical separation and phase relationship between the feed current to the elements, will show maximum reinforcement. The resultant radiation pattern is shown at a 0° wave angle over ideal ground.

Consider the specific case where the spacing is λ/4 and the phase difference is 90°, as shown in Fig 11-1. Rays b and b' are clearly in phase (λ/4 due to spacing, minus λ/4 due to phase difference of 90°). Similarly, d and d' are 180° out of phase. If the current magnitudes feeding the elements are equal, the radiation will be canceled in that direction. See also Chapter 7, Section 1.6.

1.2. Directivity Over Perfect Ground

Fig 11-2 shows a range of radiation patterns obtained by different combinations of two monopoles over perfect ground and at a theoretical zero wave angle. These directivity patterns are classics in every good antenna handbook.

1.3. Directivity Over Real Ground

Over real ground there is no radiation at zero wave angle. All the effects of real ground, which were described in detail in Chapter 9 (Verticals), apply to arrays of verticals.

1.4. Direction of Firing

The rule is simple: An array always fires in the direction of the element with the lagging feed current.

1.5. Phase Angle Sign

Phase angles are a relative thing, which means you can put your "reference" phase angle of 0° anywhere in the array. We will stick to our own convention of assigning the 0° phase angle to the "back" element of the array. This means that the feed currents in all other elements will carry the negative sign.

2. ARRAY ELEMENTS

In principle, you can use verticals of a length longer than λ/4 (electrically) for building arrays, but in that case the various feed systems described in this chapter do not apply. However, the whole range of verticals described in Chapter 9 can be used as elements for these vertical arrays, provided they are base-fed and are not longer than λ/4 electrically.

Quarter-wave elements have gained a reputation for giving a reasonable match to a 50-Ω line, which is certainly true for single vertical antennas. In this chapter we will learn the reason why quarter-wave resonant verticals do not have a resistive 36-Ω feed-point impedance when operated in arrays (even assuming a perfect ground). Quarter-wave elements still remain a good choice, since they have a reasonably high radiation resistance. This ensures good overall efficiency. On 160 meters, the elements could be top-loaded verticals, as described in Chapter 9.

The design methodology for arrays given in Section 3, as well as all the designs described in Section 4, assume that all the array elements are physically identical, with a current distribution that is the same on each element. In practice this means that only elements with a length of up to λ/4 should be used. Remember, the patterns given in Section 4 do not apply if you use elements much longer than λ/4. They certainly do not apply for elements that are λ/2 or ⅝ λ long. If you want to use long elements, you will have to model the design using the particular element lengths (Ref 959). This may be a problem if you want to use shunt-fed towers carrying HF beams as elements for an array. With their top loads, these towers are electrically often much longer than λ/4.

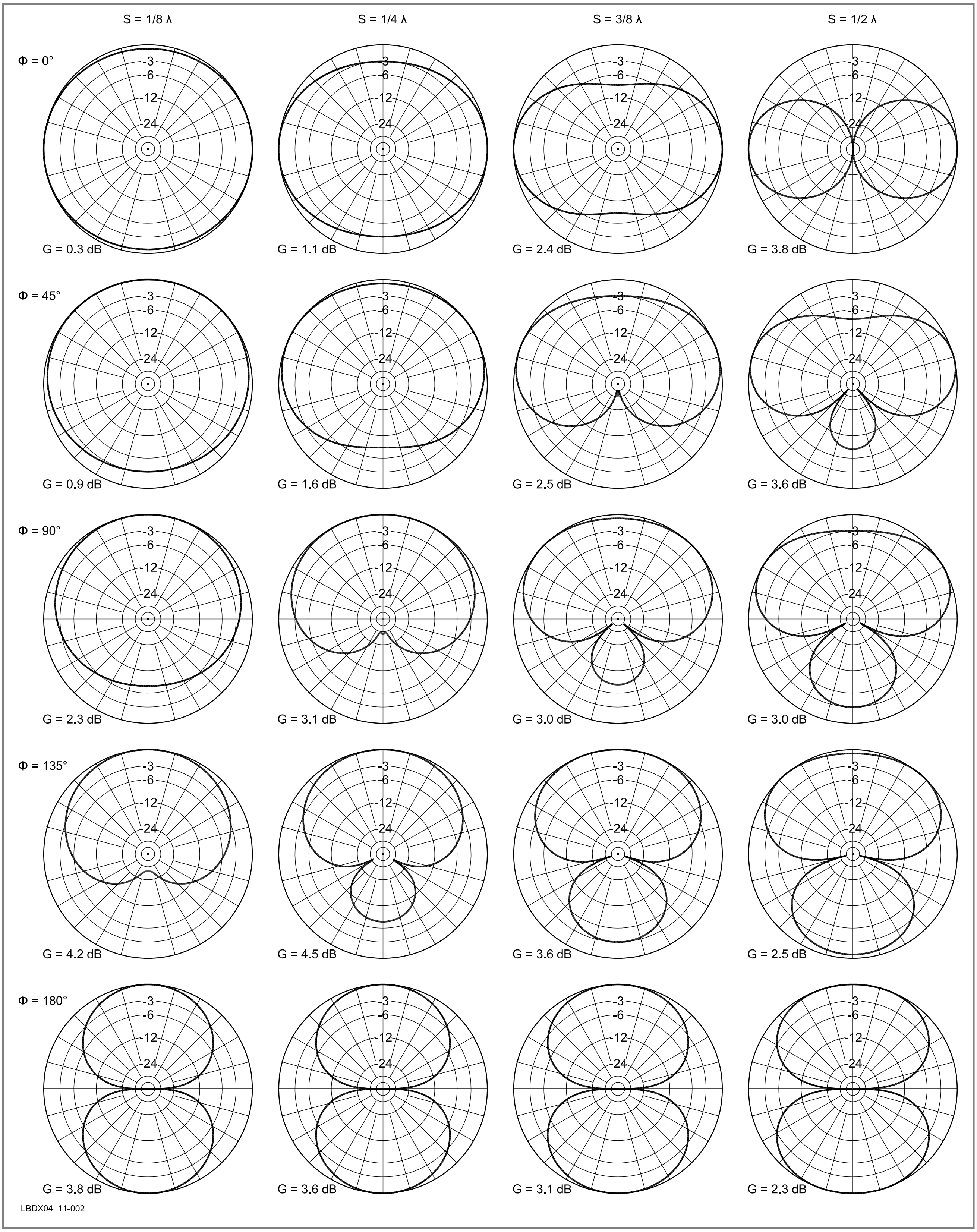


Fig 11-2 — Horizontal radiation patterns for 2-element vertical arrays (both elements fed with the same feed current magnitude). The elements are in the vertical axis, and the top element is the one with the lagging phase angle. Patterns are for 0° elevation angle over ideal ground. (Courtesy *The ARRL Antenna Book*)

3. DESIGNING AN ARRAY

The radiation patterns shown in Fig 11-2 give a good idea what can be obtained with different spacings and different phase delays for a 2-element array. For arrays with more elements there are a number of popular classic designs. Many of those are covered in detail in this chapter. A good array should meet the following specifications:

- High gain (you want your signals to be strong).
- Good directivity (F/B, forward beamwidth, RDF and DMF — see Chapter 7, Section 1.10 — especially if you will not be using separate receiving antennas).
- Ease of feeding.
- Ease of direction switching.

3.1. Modeling Arrays

We all know that it is the current at the base of the array elements (element feed current) that makes the array play. Therefore currents, rather than voltages, should be specified at the base of the vertical array elements. All modern modeling programs allow you to define "sources" as current sources or as voltage sources.

In the early days of antenna modeling we had no choice but to specify the sources at each individual radiating element. Models just comprised "radiating wires" which you could excite "locally," and loads (R, L and C) that you could insert in these wires.

Antenna modeling has come a long way though. Today antenna models not only comprise wires, but a range of other components: transmission lines, transformers and L-networks are the most important ones. All of this makes it possible to model entire *systems* instead of just a bunch of radiating wires. Today we can make a model of a multi-element array where we have only *one* source, one place where we specify the drive current or voltage. On the CD that comes with this book there are many *EZNEC* modeling files that include the feed lines to the elements or even the network(s) that take care of providing the right feed current to each one of the elements.

This has the tremendous advantage that you are now modeling an entire antenna system, and no longer just a part of the system (the radiating elements). This and the ability to sweep frequencies, makes it possible to model the antenna systems (arrays) over a given frequency range, without requiring any additional parallel circuit modeling to be done for parts of the antenna system. Such exercises make it possible to easily assess the influence of all the components together (linked to one another) on the performance of the array system.

Modern modeling programs such as *EZNEC 5.0* and *EZNEC Pro/2* use the *NEC-2* engine. Only *EZNEC Pro/4* uses the *NEC-4* engine, for which a special and rather expensive license is required. Studies on elevated radial systems and on buried radials require a program based on *NEC-4*. Some of the modeling in this book was done with a *NEC-4* engine, in which case it is mentioned,

Let me warn the readers once again: Antenna modeling is one thing; practical antenna design and construction is a very different thing. One can model complex arrays with amazing characteristics (at one frequency), but such arrays may be difficult to build and certainly are not a project for someone with little technical expertise. Always remember when you are modeling an antenna that modeling is a mathematical exercise that tries to quantify a physical process. The precision of the mathematical result is only as good as your model and as good as the precision of your input data. We must realize that most in most cases the input data are not known with a precision of more than 1%. Try to understand what the figures mean. A computer program is, in most cases, a very precise, but non-intelligent tool. The intelligence must come from the user of the computer program.

When you make a model, don't forget — at least in the final stage of your modeling effort — to include all the loss mechanisms as accurately as you can. Use real feed lines, real conductors and real ground in your models. If extreme impedances (high or low) and relatively long coax runs (eg 270° feed lines) are involved, you may be surprised at the difference between the ideal world (no losses) and the real world!

In this chapter, the influence of the loss introduced by an imperfect radial system has been included in the form of an equivalent loss resistance in series with each element feed point (most models used 5 Ω).

With this 5th edition of this book I have also made available (on the CD) the *EZNEC* modeling files of all the arrays that are used and referred to in this chapter.

3.1.1. Evaluating the Operational Bandwidth of an Array

Based on the ability to model radiating elements (wires) *plus* feed lines, networks (even hybrid couplers!) and transformers, we can easily assess the operational performance and operational bandwidth of the array over a given frequency range. Many years ago, when asked about the bandwidth of an antenna system, the answer came almost always as "the SWR is below 2:1 over a bandwidth of X MHz."

But, of course, there are more operational bandwidth determining parameters than SWR. These performance parameters are *gain*, *F/B* (or, in more general terms, *directivity*) and *SWR*. As we will see from the analysis done on different arrays with different feed systems, gain and SWR are usually not a reason for concern. Directivity is the key issue. Unless otherwise indicated, gain is calculated over Average ground (ρ = 5 mS, ε = 13) and the models include 5 Ω equivalent ground loss resistance at the base of each element.

In this book I have set a limit of F/B of 20 dB as the minimum acceptable value. We could similarly set a limit for gain (for example, not less than the gain at the design frequency minus 0.5 dB) and for SWR (eg <2:1).

I have done an in-depth analysis of the operational bandwidth of various types of commonly used arrays. The results are covered in the relevant sections of this chapter.

3.2 About Polar and Rectangular Coordinates

We will be going into detail on various issues and aspects of arrays and will be talking impedances all the time. It's a good idea to review a few basics.

- **Complex impedance**: A complex impedance is an impedance consisting of a real part (resistive part) and an imaginary part (reactive part).
- **Complex number**: A complex impedance is represented by a complex number.
- **Complex number representation**: While a real number can be represented as a point on a line, a complex number must always be represented as a point in a plane. A real

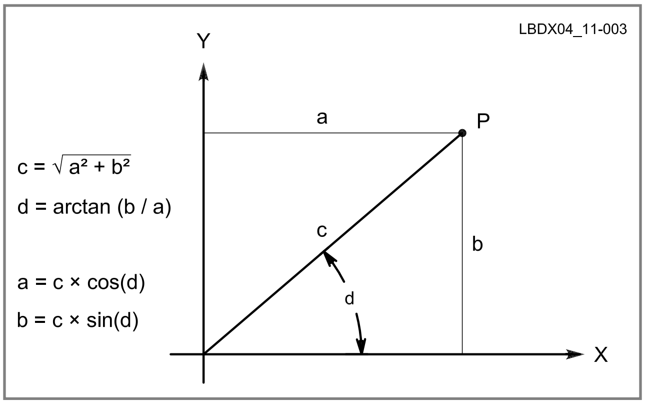


Fig 11-3 — Complex number representation.

number has one coordinate (the distance from the origin on the line) while the complex figure has two coordinates, which are necessary to unambiguously define its position in a plane.

- **Rectangular coordinates**: In a rectangular coordinate system, which in a plane consists of an X and a Y axis, the X and the Y coordinates define the complex number. If the X value is a and the Y value is b, the complex number is written as a + *j* b. The *j* indicates that the figure following is the Y coordinate, which stands for the imaginary part.
- **Polar coordinates**: In a polar coordinate system, the position of the point representing the complex number is given by its distance to the coordinate origin and the angle of the vector going from the origin to the point, the angle with respect to the X axis. The complex number in a polar coordinate system is written as c ∠d° where c = vector length and d = angle.

For some reason impedances are usually written in rectangular form as a + *j* b, while voltages and currents are most often represented in polar notation as c ∠d°. In the *New Low Band Software*, complex values of Z, I and E are always expressed in both coordinate systems. With some simple trigonometry we can always convert from one system to another. See **Fig 11-3**, where conversion formulas are included.

3.3. Getting the Right Current Magnitude and Phase

There is a world of difference between designing an array on paper or with a computer modeling program and realizing it in real life. With single-element antennas such as a single vertical or a dipole, we do not have to bother with the feed current (magnitude and phase), as there is only one feed point anyway. With phased arrays things are vastly different.

First, we must decide which array to build. Once we do this, the problem will be how to achieve the right feed currents in all the elements (magnitude and phase angle). When we analyze an array with a modeling program, we notice that the feed-point impedances of the elements change from the value for a single element. If the feed-point impedance of a single quarter-wave vertical is 36 Ω over perfect ground, it is almost always different from that value in an array because of mutual coupling.

3.3.1. The Effects of Mutual Coupling

Until about 20 years ago, few articles in Amateur Radio publications addressed the problems associated with mutual coupling in designing a phased array and in making it work as it should. Forest Gehrke, K2BT, wrote an outstanding series of articles on the design of phased arrays (Refs 921-925, 927). These are highly recommended for anyone who is considering putting up phased arrays of verticals. Another excellent article by Al Christman, K3LC (Ref 929), covers the same subject. The subject has been very well covered in the 15th and later editions of *The ARRL Antenna Book*, where Roy Lewallen, W7EL, wrote a comprehensive contribution on arrays. Today, Tom Rauch, W8JI, is a good teacher on principles and practical aspects of arrays in his excellent Web site (**www.w8ji.com**), and his advice in these matters on the Top Band reflector is much appreciated by all.

If we bring two nearly resonant circuits into the vicinity of each other, mutual coupling will occur. This is the reason that antennas with parasitic elements work as they do. Horizontally polarized antennas with parasitically excited elements are widely used on the higher bands. On the low bands the proximity of the ground limits the amount of control the designer has on the current in each of the elements. Arrays of vertical antennas, where each element is fed, overcome this limitation, and in principle the designer has an unlimited control over all the design parameters. With so-called *phased arrays*, all elements are individually and physically excited by applying power to the elements through individual feed lines. Each feed line supplies current of the correct magnitude and phase.

There is one frequently overlooked major problem with arrays. As we have made up our minds to feed all elements, we too often assume (incorrectly) there is no mutual coupling or that it is so small that we can ignore it. Taking mutual coupling into account complicates life, as we now have two sources of applied power to the elements of the array: parasitic coupling and direct feeding.

3.3.1.1. Self Impedance

If a single quarter-wave vertical is erected, we know that the feed-point impedance will be 36 + *j* 0 Ω, assuming resonance, a perfect ground system and a reasonably thin conductor diameter. In the context of our array we will call this the *self impedance* of the element. (For example, we will refer to the self impedance of element 1 as Z11.)

3.3.1.2. Coupled Impedance

If other elements are closely coupled to the original element, the impedance of the original element will change. Each of the other elements will couple energy into the original element and vice-versa. This is often termed *mutual coupling* since each element affects the other. The coupled impedance is the impedance of an element being influenced by one other element and it is significantly different from the self impedance in most cases. (For example, we will refer to the coupled impedance of element 1 with element 2 coupled as Z1,2.)

3.3.1.3. Mutual Impedance

The mutual impedance is a term that defines unambiguously the effect of mutual coupling between a set of two antenna elements. Mutual impedance is an impedance that cannot be measured. It can only be calculated. The calculated mutual impedances and driving impedances have been extensively covered by Gehrke, K2BT (Ref 923).

3.3.1.4. Drive Impedance

To design the correct feed system for an array, you must know the drive impedances of each of the elements, as well as the correct current magnitude and angle needed to feed the element(s).

3.3.2. Calculating the Drive Impedances

You cannot measure mutual impedance. It must be calculated. Mutual impedances are calculated from measured self impedances and drive impedances. Here is an example: We are constructing an array with three ¼-λ elements in a triangle, spaced ¼-λ apart. We erect the three elements and install the final ground system. Make the ground system as symmetrical as possible. Where the buried radials cross, terminate them in a bus. Then the following steps are carried out:

1) Open-circuit elements 2 and 3. Opening an element will effectively isolate it from the other elements in the case of quarter-wave elements. When using half-wave elements, the elements must be grounded for maximum isolation and open-circuited for maximum coupling.
2) Measure the self impedance of element 1 (Z11).
3) Ground element 2.
4) Measure the coupled impedance of element 1 with element 2 coupled (Z1,2).
5) Open-circuit element 2.
6) Ground element 3.
7) Measure the coupled impedance of element 1 with element 3 coupled (Z1,3).
8) Open-circuit element 3.
9) Open-circuit element 1.
10) Measure the self impedance of element 2 (Z22).
11) Ground element 3.
12) Measure the coupled impedance of element 2 with element 3 coupled (Z2,3).
13) Open-circuit element 3.
14) Ground element 1.
15) Measure the coupled impedance of element 2 with element 1 coupled (Z2,1).
16) Open-circuit element 1.
17) Open-circuit element 2.
18) Measure the self impedance of element 3 (Z33).
19) Ground element 2.
20) Measure the coupled impedance of element 3 with element 2 coupled (Z3,2).
21) Open-circuit element 2.
22) Ground element 1.
23) Measure the coupled impedance of element 3 with element 1 coupled (Z3,1).

This is the procedure for an array with three elements. The procedures for 2 and 4-element arrays can be derived from the above steps.

As you can see, measurement of coupling is done by pairs of elements. At step 15, we are measuring the effect of mutual coupling between elements 2 and 1, and it may be argued that this has already been done in step 4. It is useful, however, to make these measurements again in order to recheck the previous measurements and calculations. Calculated mutual couplings Z12 and Z21 (see below) using the Z1,2 and Z2,1 inputs should in theory be identical, and in practice should be within an ohm or so.

The self impedances and the driving impedances of the different elements should match closely if the array is to be made switchable.

The mutual impedances can be calculated as follows:

$$Z12 = \pm\sqrt{Z22 \times (Z11 - Z1,2)}$$

$$Z21 = \pm\sqrt{Z11 \times (Z22 - Z2,1)}$$

$$Z13 = \pm\sqrt{Z33 \times (Z11 - Z1,3)}$$

$$Z31 = \pm\sqrt{Z11 \times (Z33 - Z1,3)}$$

$$Z23 = \pm\sqrt{Z33 \times (Z22 - Z2,3)}$$

$$Z32 = \pm\sqrt{Z22 \times (Z33 - Z3,2)}$$

It is obvious that if Z11 = Z22 and Z1,2 = Z2,1, then Z12 = Z21.

If the array is perfectly symmetrical (such as in a 2-element array or in a 3-element array with the elements in an equilateral triangle), all self impedances will be identical (Z11 = Z22 = Z33), and all driving impedances as well (Z2,1 = Z1,2 = Z3,1 = Z1,3 = Z2,3 = Z3,2). Consequently, all mutual impedances will be identical as well (Z12 = Z21 = Z31 = Z13 = Z23 = Z32). In practice, the values of the mutual impedances will vary slightly, even when good care is taken to obtain maximum symmetry.

Because all impedances are complex values (having real and imaginary components), the mathematics involved are difficult. The *w1mk-on4un-oh1tv-arrays.xls* spreadsheet included on this book's CD will do all the calculations in seconds. No need to bother with complex algebra. Just answer the questions on the screen.

Fig 11-4 shows the mutual impedance to be expected for quarter-wave elements at spacings from 0 to 1.0 λ. The resistance and reactance values vary with element separation as a damped sine wave, starting at zero separation with both

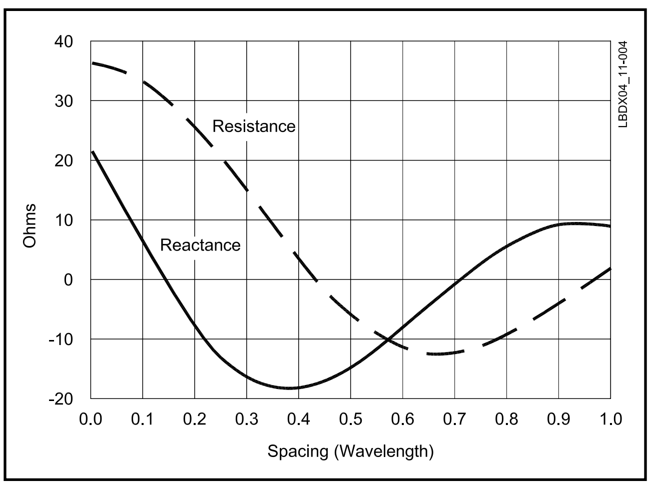


Fig 11-4 — Mutual impedance for two λ/4 elements. For shorter vertical elements (length between 0.1 and 0.25 λ), one can calculate the mutual impedance by multiplying the figures from the graph by the ratio R_{rad}/36.6 where R_{rad} = the radiation resistance of the short vertical.

signs positive. At about 0.10 to 0.15 λ spacing, the reactance sign changes from + to –. This is important to know in order to assign the correct sign to the reactive value (obtained via a square root).

Gehrke, K2BT, emphasizes that the designer should actually measure the impedances and not take them from tables. Some methods of doing this are described in Ref 923. The published tables show ballpark figures, enabling you to verify the square-root sign of your calculated results.

After calculating the mutual impedances, the drive impedances can be calculated, taking into account the drive current (amplitude and phase). The driving-point impedances are given by

$$Zn = \frac{I1}{In} \times Zn1 + \frac{I2}{In} \times Zn2 + \frac{I3}{In} \times Zn3 \ldots + \frac{In}{In} \times Znn$$

where n is the total number of elements. The number of equations is n. The above formula is for the nth element. Note also that Z12 = Z21 and Z13 = Z31, etc.

The above-mentioned programs perform the rather complex driving-point impedance calculations for arrays with up to four elements. The required inputs are:

1) The number of elements.

2) The driving current and phase for each element.

2) The mutual impedances for all element pairs.

The outputs are the driving-point impedances Z1 through Zn.

Design Example

Let us take the example of an array consisting of two ¼-λ long verticals, spaced ¼ λ apart and fed with equal magnitude currents, with the current in element 2 lagging the current in element 1 by 90°. This is the most common (though not necessarily the best) end-fire configuration with a cardioid pattern.

Self impedance: The quarter-wave long elements of such an array are assumed to have a self impedance of 36.4 Ω over a perfect ground. A nearly perfect ground system consists of at least 120 half-wave radials (see Chapter 9, Vertical Antennas). For example, a system with "only" 60 radials may (depending on the ground quality) show a self impedance on the order of 40 Ω (see Chapter 9). Let's use 41 Ω (5 Ω ground loss resistance).

Coupled Impedance: We measured 37.5 + *j* 15.2 Ω.

Mutual impedance: The mutual impedances were calculated with the above-mentioned computer program: Z12 = Z21 = 19.76 – *j* 15.18 Ω. From the mutual impedance curves in Fig 11-4 it is clear that the minus sign is the correct sign for the reactive part of the impedance.

An alternative to calculating the mutual impedance (Z12, Z21) via measuring self impedance (Z11, Z12) and coupled impedance (Z1,2, Z2,1) is to accurately model the array using *EZNEC*. In case of the 2-element array this can be done as follows:

1) Specify the current for element 1 as 1 A, and for element 2 as 0 A. Under these circumstances the impedance of element 1 (in the Source Data window) is Z11, as element 2 is ungrounded, thus uncoupled to element 1 (eg Z1 = 41 + *j* 0 Ω).

2) Next specify both feed current magnitudes as 1 A (and in phase) and note the impedance of element 1A and call it $Z1_{feed}$ (eg 56.11 – *j* 14.2 Ω).

3) Z12 = Z21 = $Z1_{feed}$ – Z11 = (60.8 – *j* 14.04) – (41 + *j* 0) = 19.8 – *j* 14.04 Ω

Note that this result is close to what we arrived at above (19.76 – *j* 15.18).

Drive impedances: The same software module calculates the drive impedances (also called feed-point impedances) of the two elements. Let us assume both elements are fed with equal magnitude feed currents and 105° phase shift.

Z1 = 50.55 + *j* 23.02 Ω for the 0° element

Z2 = 21.11 – *j* 15.16 Ω for the –105° element (the front element).

We have now calculated the impedance of each element of the array, the array being fed with the current (magnitude and phase) as set out. We have used impedances that we have measured; we are not working with theoretical impedances.

The "2 el and 4 el Vertical Arrays" module of the *New Low Band Software* is a perfect tool to guide you along the design of an array. You can enter your own values or just work

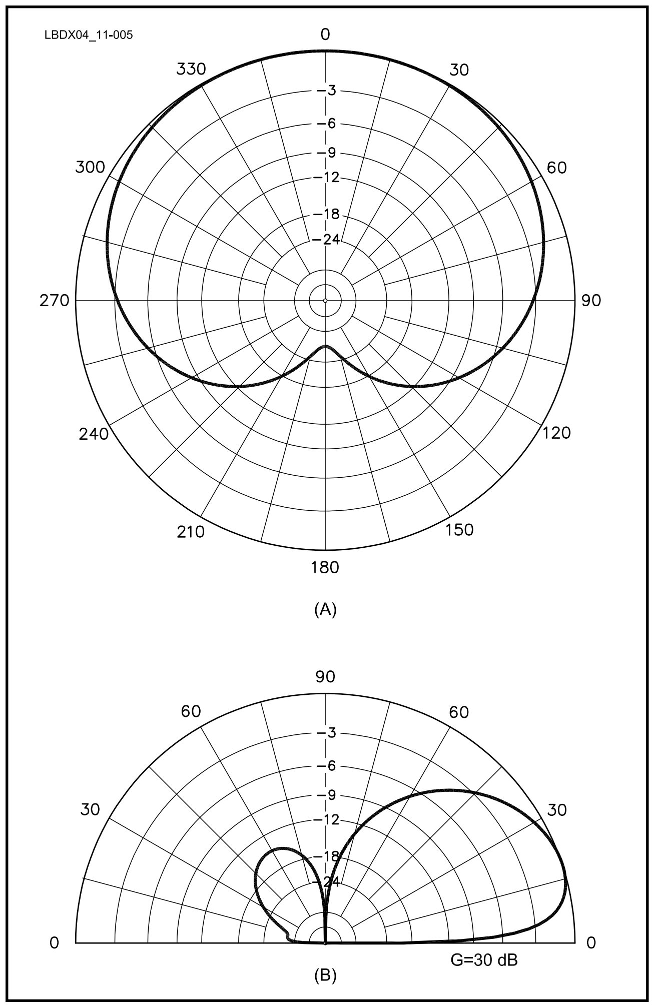


Fig 11-5 — Vertical and horizontal radiation patterns for the 2-element cardioid array, spaced 90° and fed with 90° phase difference. The pattern was calculated for very good ground with a radial system consisting of 120 radials, each 0.4 λ long (the equivalent ground resistance is 2 Ω). The gain is 3.0 dB as compared to a single vertical over the same ground and radial system. The horizontal pattern at A is for an elevation angle of 19°.

your way through using a standard set of values.

3.3.3. Modeling the Array

With the latest *NEC-4* based software, you can include the radial system, but for the design and evaluation of arrays, a *MININEC*-based modeling program, or even better a *NEC*-based program using a *MININEC* ground (such as provided in *EZNEC*) will do, as long as we realize that we must add some equivalent series resistance to account for the ground losses of the radial system. In order to simulate the effect of a radial system consisting of 60 quarter-wave radials, I inserted 4 Ω in series with the feed point of each antenna.

Modeling the cardioid antenna over *MININEC* ground with 4 Ω loss resistance included in each element, *EZNEC* comes up with the following impedances:

$Z1 = 55.0 + j\,22.7\ \Omega$ (back element)
$Z2 = 26.5 - j\,19.5\ \Omega$ (front element)

These are close to the values worked out with the software mentioned earlier, based on measured values of coupled and self impedances. The vertical and the horizontal radiation patterns for the 2-element cardioid array are shown in **Fig 11-5**.

3.4. Designing a Feed System

The challenge now is to design a feed system that will supply the right current to each of the array elements. As we now know the current requirements as well as the drive-impedance data for each element of the array, we have all the required inputs to design a feed system.

Each element will need to be supplied power through its own feed line. In a driven array each element either gets power, or it delivers power into the feed system. During calculations we will sometimes encounter a negative feed-point impedance, which means the element is actually delivering power into the feed network. If the element impedance is zero, this means that the element can be shorted to ground. It then acts as a parasitic element.

Eventually all the feed lines will be connected to a common point, which will be the common feed point for the entire array. You can only connect feed lines in parallel if the voltages on the feed lines (at that point) are identical (in magnitude and phase) — the same as with ac power!

Designing a feed system consists of calculating the feed lines (impedance and length) as well as the component values of networks used in the feed system, so that the voltages at the input ends of the lines are identical. It is as simple as that.

The ARRL has published the original (1982) work by Lewallen, W7EL, in the last five editions of *The ARRL Antenna Book*. This material is a must for every potential array builder. However, there are other feed methods than the Lewallen method. Various feed systems are covered in the following sections of this book:

- Christman method
- Using flat lines
- Crossfire principle
- Lewallen (quadrature fed arrays)
- Lewallen/Lahlum (any phase angle, any current ratio)
- The hybrid coupler (demystified)
- New optimized hybrid coupler systems
- Gehrke (broadcast approach)
- Lahlum/Gehrke (non-current-forcing, L-network)
- Ketonen opposite voltage feed system (also called voltage-forcing system)

3.4.1. The Wrong Way

In just about all cases, the drive impedance of each element will be different from the characteristic impedance of the feed line. This means that there will be standing waves on the line. This has the following consequences:

- The impedance, voltage and current will be different in each point of the feed line.
- The current and voltage phase shift is not proportional to the feed line length, except for a few special cases (eg, a half-wave-long feed line).

This means that if we feed these elements with 50-Ω coaxial cable, we cannot simply use lengths of feed line as phasing lines by making the line length in degrees equal to the desired phase delay in degrees. In the past we have seen arrays where a 90° long coax line was inserted in one of the feed lines to an element to create a 90° antenna current phase shift. Let us take the example of the 2-element cardioid array (as described above) and see what happens (see **Fig 11-6**).

We run two 90° long coax cables to a common point. Using the "Coax Transformer/Smith Chart" module of the *New Low Band Software*, we calculate the impedances at the end of those lines. (I used RG-213 with 0.35 dB/100 ft attenuation at 3.5 MHz). The array element feed impedances, including 2 Ω of equivalent ground loss resistance, are (let's use round figures):

$Z1 = 51 + j\,20\ \Omega$
$I1 = 1\ \text{A}, \angle -90°$

From E = Z/I we can calculate (don't worry the software does it for you):

$E1 = 54.8\ \text{V}, \angle -68.6°$

and

$Z2 = 21 - j\,20\ \Omega$
$I2 = 1\ \text{A}, \angle 0°$
$E2 = 29\ \text{V}, \angle -43.6°$

At the end of the 90° long RG-213 feed lines, the imped-

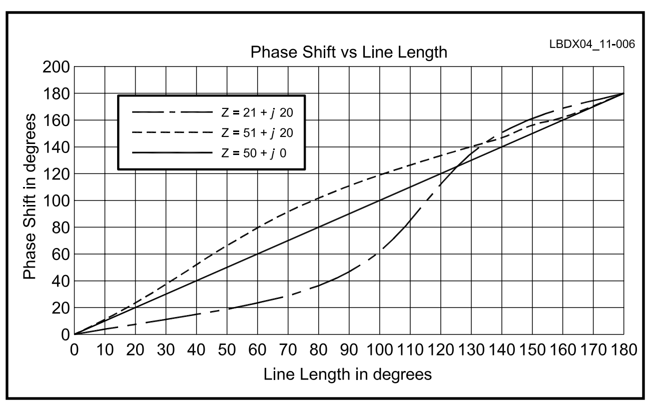


Fig 11-6 — Graph showing the current phase shift in a 50-Ω line (RG-213, on 80 meters), as a function of the load impedance. The loads shown are those for a 2-element cardioid array as used in the text. Note that the phase shift does *not* equal line length, except when the line is terminated in its own characteristic impedance!

ances (and voltages) become:

Z1' = 42.81 – j 16.18 Ω
E1' = 50.89 V, ∠0.39°
I1' = 1.11 A, ∠21.09°

and

Z2' = 63.1 + j 56.94 Ω
E2' = 50.37 V, ∠89.61°
I2' = 0.59 A, ∠47.54°

If we make the line to the lagging element 180° long (90° plus the extra 90° for obtaining an extra 90° phase shift), we end up with:

Z1" = 51.18 + j 18.64 Ω
E1" = 56.42 V, ∠110.77°
I1" = 1.04 A, ∠90.76°

Note that E2' and E1" are not identical. This means we cannot connect the lines in parallel at those points without upsetting the antenna current (magnitude and phase).

From the above voltages we see that the extra 90° line created an actual current phase difference of 90.76° – 21.09° = 68.67°, and *not* 90° as required.

The software module "Impedances, Currents and Voltages Along Feed Lines" is ideally suited for analyzing this phenomenon. Look at the values of voltage and current as you scan along the line, and remember we want the right current phase shift and we want the same voltage where we connect the feed lines in parallel.

If you have such a feed system, do not despair. Simply by shortening the phasing line from 90° to 71°, you can obtain an almost perfect feed system. (See **Fig 11-7**).

Watch out: If you want to use this system, make sure you have the same feed impedances as in the model above. How? By calculating the drive impedances as outlined in Section 3.3.2, or by carefully modeling your array, making sure you take into account all the small details!

3.4.2. Christman Method

In the Christman, K3LC method (Ref 929), we scan the feed lines to the different elements looking for points where the voltages are identical. If we find such points, we connect them together, and we are all done! It's really as simple as that. Whatever the length of the lines are, provided you have the right current magnitude and phase at the input ends of the lines, you can always connect two points with identical voltages in parallel. That's also where you feed the entire array.

Christman makes very clever use of the transformation characteristics of the feed lines. We know that on a feed line with SWR, voltage, current and impedance are different in every point of the line. The questions are now, "Are there points with identical voltage to be found on all of the feed lines?" and "Are the points located conveniently; in other words, are the feed lines long enough to be joined?" This has to be examined case by case.

It must be said that we cannot apply the Christman method in all cases. I have encountered situations where identical voltage points along the feed lines could not be found. The software module "Impedance, Current and Voltage Along Feed Lines," which is part of the *New Low Band Software*, can provide a printout of the voltages along the feed lines.

The required inputs are:

- Feed-line impedance.

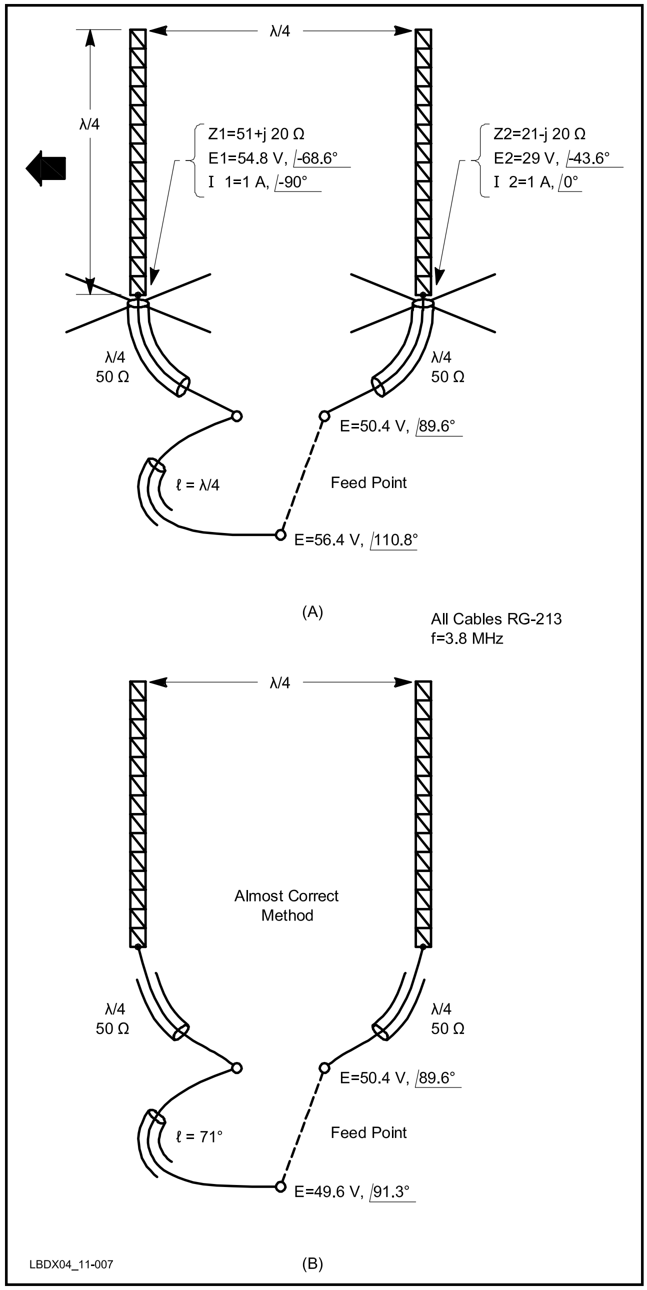


Fig 11-7 — At A, the incorrect way of feeding a 2-element cardioid array (90° phase, 90° spacing). Note that the voltages at the input ends of the two feed lines are not identical. In B we see the same system with a 70° long phasing line, which now produces almost correct voltages. The F/B ratio of existing installations will jump up by 10 or 15 dB, just by changing the line length from 90° to 70°.

- Driving-point impedances (R and X).
- Current magnitude and phase.

Continuing with the above example of a 2-element configuration (90°spacing, 90° phase difference, equal currents, cardioid pattern), we find:

E1 = (155° from the antenna element) = 47.28 ∠86.1°V
E2 = (84° from the antenna element) = 47.27 ∠85.9°V

Notice on the printout that the voltages at the 180° point on line 1 and at the 90° point on line 2 are not identical (see Section 3.4.1), which means that if you connect the lines in parallel in those points, you will not have the proper current in the antennas.

We need now to connect the two feed lines together where the voltages are identical. If you want to make the array switchable, run two 84° long feed lines to a switch box and insert a 155° – 84° = 71° long phasing line, which will give you the required 90° antenna-current phase shift. **Fig 11-8** shows the Christman feed method.

Of course the impedance at the junction of the two feed lines is not 50 Ω. Using the "Coax Transformer/Smith Chart" software module, we calculate the impedances at the input ends of the two lines we are connecting in parallel:

Z1end = 39 + j 12 Ω
Z2end = 50 + j 52 Ω

The software module "Parallel Impedances" calculates the parallel impedance as 23.8 + j 12.4 Ω. This is the feed-point impedance of the array. You can use an L network, or any other appropriate matching system to obtain a more convenient SWR on the 50-Ω feed line.

3.4.3. Using Flat Lines (SWR ~1:1) with "Length = Phase Shift"

Let's go back to Fig 11-7. The impedance at the end of the quarter-wave line going to the front element is 42.81 – j 16.18 Ω. Maybe we can turn it into a purely resistive impedance of convenient value by connecting a reactance in parallel. Using the "Shunt/Series Impedance Network" module of the *New Low Band Software*, we can easily calculate the required parallel impedance to make it a purely resistive impedance. In this case it appears that putting an inductance of +129.4 Ω in parallel at that point turns the impedance into 48.9 Ω, very close to 50 Ω. Let's do that, and now connect a quarter-wave phasing line from that point to the end of the quarter-wave line coming from the back element. As the line now operates with an SWR of very close to 1:1, phase difference equals line length, and we have exactly what we want.

Fig 11-9 shows the layout of this system. If you want more phase shift, say 120° to lift the notch off the ground (see Chapter 7) you simply make the phasing line 120° long. Note however that the element feed impedances shown are for 90° phase shift and that those are slightly different when you change the elevation angle.

It is obvious that such a method can only be applied when you are lucky to find an impedance (after tuning out the reactance by a parallel element) that matches an existing feed-line impedance. You can, of course, use parallel feed lines to obtain low impedances. You can actually connect feed lines of different impedances in parallel. For example, 25 Ω = two 50 Ω lines in parallel; 30 Ω = a 50 Ω and a 75 Ω line in parallel; and 37.7 Ω = two 75 Ω in parallel.

3.4.4. The Cross Fire (W8JI) Principle

In a "standard" array, for example as shown in Section 3.4.2 and 3.4.3, the feed line goes to the back element, and the front element is fed via a phasing line. Let us analyze what happens in such a design when we change frequency away from the nominal design frequency.

Assume we have a 2-element end-fire array, spaced exactly λ/4 (90°) and with exactly 90° phase shift (this is by far not the best arrangement!). Our notch elevation angle will be 0° (see Chapter 7). If we increase the frequency by 5%, the spacing becomes 94.9° and the phasing becomes also larger (if the lines are relatively flat, also about 5% longer). But, in order to maintain the zero notch angle at ground level, we need

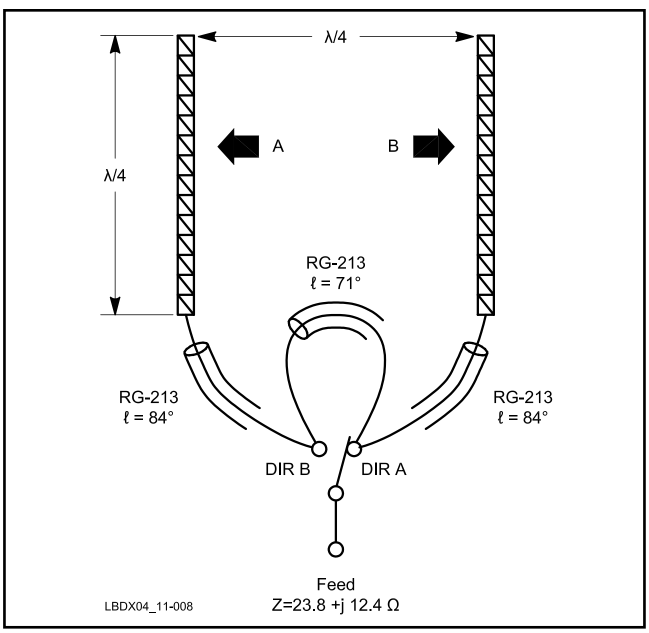


Fig 11-8 — Christman feed system for the 2-element λ/4-spaced cardioid array fed 90° out-of-phase. Note that the two feed lines are 84° long (not 90°), and that the "90° phasing line" is actually 71 electrical degrees in length. The impedance at the connection point of the two lines is 23.8 + *j* 12.4 Ω (representing an SWR of 2.3:1 for a 50-Ω line), so some form of matching network is desirable.

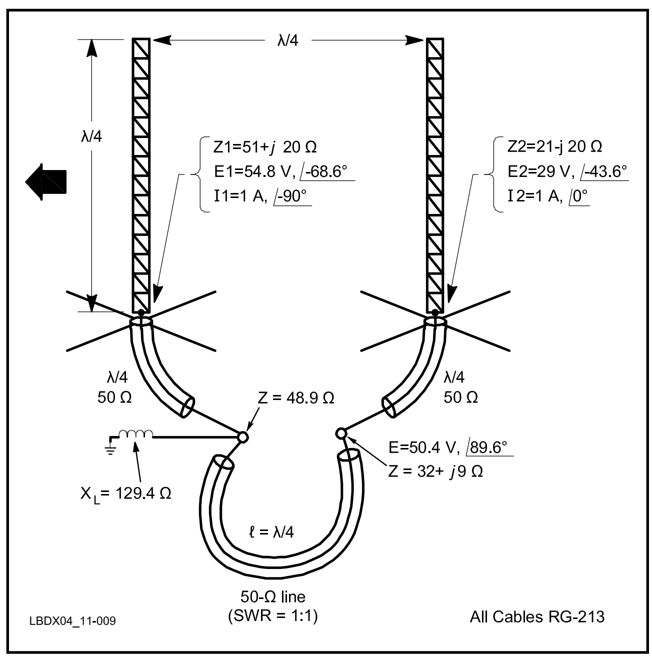


Fig 11-9 — Adding a coil with a reactance of +129.4 Ω at the end of the 1/4 λ feed line going to the front element turn the impedance at that point into 48.9 Ω, very close to 50 Ω. Now we can insert a 50-Ω delay line and know that phase shift equals line length.

the phasing line to be *shorter* by about 5%. This mechanism limits the usable bandwidth in such arrays. In simple 2-element arrays this usually is not a problem, but in more complex arrays using four or more elements it can become a key design factor.

Tom Rauch, W8JI, pointed out that we can also use the crossfire principle feed method, where we feed the array at the front element using a phase inverter (a 180° transformer) and feed the back element with a phasing line that is complementary in length to the required phasing angle (see Chapter 7). In this case the phase-shift transformer produces a 180° shift over a wide frequency range. At a frequency that is 5% higher than the design frequency, the phase shift produced by the phasing line becomes about 95° long. Subtracting this value from the 180° phase shift obtained by the transformer, the phase difference becomes 85° at the higher frequency. With this crossfire principle the tracking is achieved, which is exactly what we want.

In **Fig 11-10** we see that we will have to put the phasing line in the feed line going to the back element. The impedance at the end of the quarter wave line to the elements is $63.1 + j56.94\ \Omega$. Using the "Shunt/Parallel Impedance" section from the *New Low Band Software*, we find that a parallel capacitor with an impedance of –127 Ω will turn the impedance into 115 Ω, not exactly a common coaxial cable impedance. But what if we used a quarter-wave 75-Ω feed line for achieving a 90° phase shift? This will work but because the antenna impedance is not the same as the load impedance, the typical quarter-wave impedance transformation will occur. The impedance at the end of the line will be (75 × 75)/115 = 49 Ω.

This means that there will be a voltage transformation of 115/49 = 2.3:1. In this particular setup, we will need to use a 180° phase-shift transformer that has a transformation ratio (turns ratio) of 2.3:1 if we want to end up with equal current magnitudes at both elements.

Would you ever want to go through this procedure to achieve tracking? No, because tracking is limited anyhow by the variation in element feed impedances as you change frequency. This principle holds very well, however, when you are using elements that show little or no change in feed impedance when the frequency is changed, which is what occurs with many receiving antennas as explained in Chapter 7.

This principle can also be used with complex arrays (four elements and more) to achieve better bandwidth. Such designs are far from being "plug and play" and are explained for the reader to understand the principle rather than to serve as a building kit! For an application of this principle see Section 4.7.3.

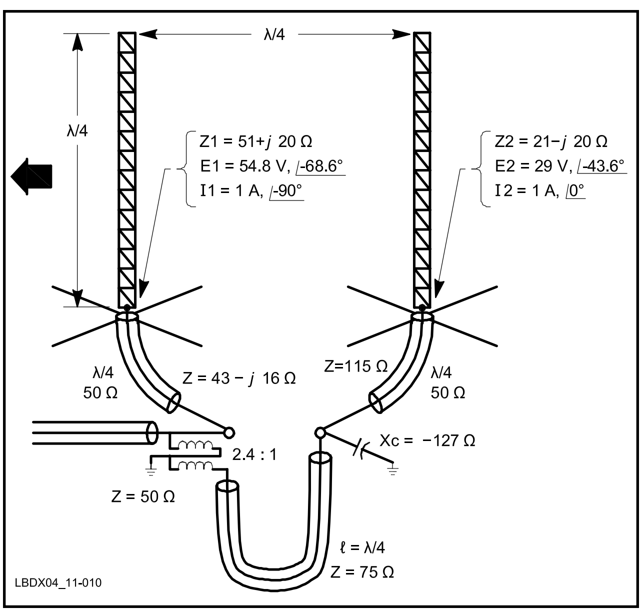


Fig 11-10 — While all other feed methods feed the back element directly and provide phase delay via coaxial cable or a network to the front element, the crossfire feeding system does the opposite. It makes use of a 180° phase-inverter transformer to achieve a feed system that guarantees phase delay to remain correct when frequency is changed. See text for details.

3.4.5. Using an L Network to Obtain the Desired Shift

3.4.5.1. Current Forcing

Roy Lewallen, W7EL, uses a method that takes advantage of the specific properties of quarter-wave feed lines (Lewallen calls it "current-forcing"). This method is covered in great detail by W7EL in recent editions of *The ARRL Antenna Book*.

A quarter-wave feed line has a wonderful property that is put to work with this particular feed method. The magnitude of the current at one end of a ¼-λ long transmission line is equal to the voltage at the other end divided by the characteristic impedance of the line, and it is independent of the load impedance. In addition, the input current lags the output voltage by 90° and is also independent of the load impedance.

3.4.5.2. Using a Simple L-Network to Obtain the Right Phase Shift

The method of using an L-network to obtain the proper phase shift has been introduced Lewallen, W7EL. The original Lewallen method is a feed method that can only be applied to antennas fed in quadrature, which means antennas where the elements are fed with phase differences that are a multiple of 90°. Later the L-network technique approach was made more flexible, and the formulas were made available where you can calculate the L-network for arrays where the L-network feeds more than one element, as well as arrays where the current magnitude is not the same in all elements.

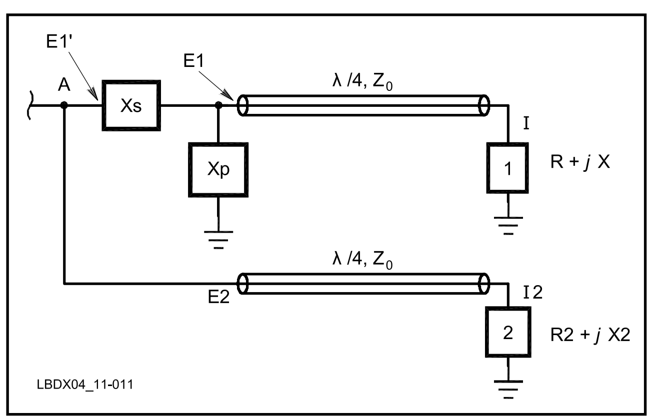


Fig 11-11 — Basic layout of the L-network phasing system developed by Roy Lewallen, W7EL, and enhanced for any phase angle by Robye Lahlum, W1MK.

It was Robye Lahlum, W1MK, who worked out the formulas that made possible the calculation of such L-networks for all of the above including any arbitrary chosen phase angle (no longer necessarily 90°).

Let's have a look at **Fig 11-11**. This is the example of a 2-element array, where element #2 is fed directly, and element #1 through the L-network. In this case the elements are fed through quarter-wave current-forcing feed lines. This is not strictly necessary, as explained in Section 3.4.8, but makes measuring and tuning easier.

Voltage E1, at the end of the feed line going to element 1, is transformed in the L-network to E1'. The transformation is:

$$E1' = k \times E1 \angle\theta° \qquad \text{(Eq 11-1)}$$

The k factor is related to the transformation's magnitude and the desired phase shift is represented by the angle θ. Obviously, we want to connect the input of the L-network (where the voltage is E1') to the input of the quarter-wave feed line going to element 2, where the voltage is E2.

We can connect those two points together, if the voltages in those points are identical. In other words if:

$$E2 = k \times E1 \angle\theta° \qquad \text{(Eq 11-2)}$$

The condition for this to apply is:

$$X_S = \frac{-\sin\theta \times Z_0^{\,2}}{n \times k \times R} \qquad \text{(Eq 11-3)}$$

$$X_P = \frac{X_S}{\left[\dfrac{n \times X \times X_S}{Z_0^{\,2}} - 1 + \dfrac{\cos\theta}{k}\right]} \qquad \text{(Eq 11-4)}$$

Theta (θ) is the desired difference between the current phase angle at the element fed through the L-network and the phase angle at the input of the network. The phase angle is responsible for a time delay, and θ must be negative. If necessary subtract 360° to obtain a negative value. Make sure you do not invert signs! Follow the examples given to understand the procedure.

The letter k is the ratio of the current supplied to the element in the branch fed through the L-network (in this case it is feed current magnitude of element 1), versus the current in the element fed directly (in this case, element 2).

The letter n is the number of identical elements (with identical feed currents) that are fed through the branch containing the L-network (see **Fig 11-12**) for a case where two elements are fed with identical current magnitudes and phases.

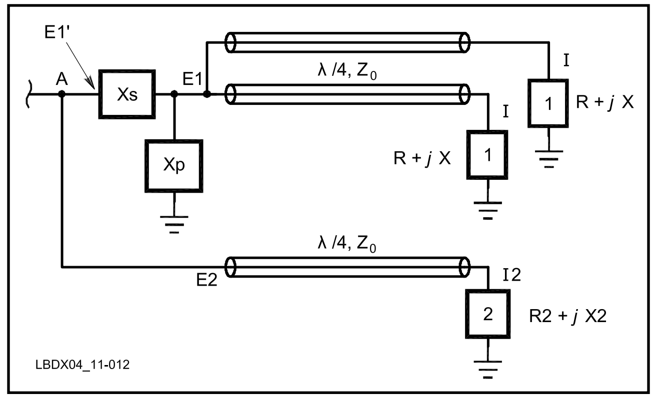


Fig 11-12 — In this particular case the L-network feeds two elements with identical feed currents. All you need to do is enter n = 2 in the *Lahlum.xls* spreadsheet.

Z_0 is the characteristic impedance of the quarter-wave (or 3λ/ 4 or 5λ/4, etc) current-forcing feed lines.

R is the real part of the feed-point impedance of one of the identical element(s).

X is the imaginary part of the feed-point impedance ($Z = X + jX$).

X_S is the impedance of the series element in the L-network.

X_P is the impedance of the parallel (shunt) element in the L-network.

These apply under all circumstances where you feed the elements via current-forcing feed lines. The impedance $R + jX$ is *not* the impedance at the end of the feed line but the feed-point impedance of an antenna element.

The equations do not work for 0° or 180°, but for 0° you do not need a phase-shifter and for 180° we have the choice between a half-wave long feed line or a 180°-phase-reversal transformer (see Section 3.4.6.3).

Note that in these equations no consideration was given to the losses in the feed lines nor in the network. Under most real-life conditions these losses are small on the low bands. We can, however, do the calculation including cable losses as well (see Section 3.4.5.4.4).

Also assuming no feed-line losses, we can easily calculate the input impedance at the input side of the L-network. The parallel input impedance components at the input of the network are:

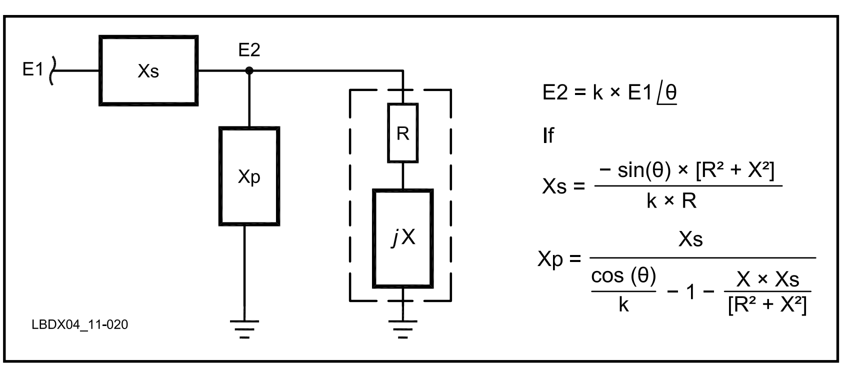


Fig 11-13 — Equations for calculating the L-network components needed to produce a desired phase shift θ, based upon the feed-point impedances (R = real part, X = reactive part). These equations do not use lossless current-forcing feed lines that are odd multiples of λ/4, although that option is available in the upper portion of the *Lahlum.xls* spreadsheet. See text for details.

$$R_{par} = \frac{Z_0^2}{K^2 \times n \times R} \qquad \text{(Eq 11-5)}$$

and

$$X_{par} = \frac{X_S}{1 - k \times \cos\theta} \qquad \text{(Eq 11-6)}$$

These values must be converted to series equivalent input impedances using the following formulas:

$$R_{ser} = \frac{R_{par} \times X_{par}^2}{R_{par}^2 + X_{par}^2} \qquad \text{(Eq 11-7)}$$

and

$$X_{ser} = \frac{R_{par}^2 \times X_{par}}{R_{par}^2 + X_{par}^2} \qquad \text{(Eq 11-8)}$$

The calculation can also be done using the using the "RC/RL Transformation" module of the *New Low Band Software* (available on the CD).

3.4.5.3. The Lahlum-Lnetwork.xls Spreadsheet Tool

I wrote an Excel spreadsheet (*Lahlum-Lnetwork.xls*) that is on the CD bundled with this book. This tool allows you to calculate the values of the L network, as well as the resulting input impedance of the branch with the L-network. Usage is simple and self-explanatory.

The spreadsheet uses the formulas shown in Section 3.4.5.2. **Fig 11-13** shows equations for calculating the L-network components needed to produce a desired phase shift based on the feed point impedances.

We will work out a number of examples of real life arrays, using the spreadsheet tool.

3.4.5.4 Two-Element End-Fire Array in Quadrature Feed

The first part of the *Lahlum-Lnetwork.xls* spreadsheet calculates without taking into account cable losses.

In this example for a two-element array, the L-network goes to one element (in a Four Square it may drive the two center elements), so enter 1 for "# of elem" (which means that 1 element requires a network to provide the required phase shift). Z_0-feed line is the characteristic impedance of the quarter-wave line going from the L-network to the element(s). R and X are the real and the imaginary values of feed point impedance of the element at the end of that line. We will work out the example for the 2-element end-fire array used in Section 3.4.1 (Fig 11-7). In this case R = 51 Ω and X = + *j* 20 Ω). For the moment always enter 1 for k (means that the current magnitude in the elements will be identical) and θ = (–90) – (0) = –90°.

The spreadsheet (**Table 11-1**) shows the calculation without cable losses. **Fig 11-14** shows the feed network for this case. Note that the difference in L-network values is very small. In most cases the "lossless" calculation will suffice. In most of the examples in this chapter we will use lossless calculations (unless otherwise mentioned).

As explained above, the formulas used in the spreadsheet assume no cable loss. If you want to calculate the L-network values and include cable loss (**Fig 11-15**), one must first calculate the impedance at the end of the current-forcing feed line, using the "Coax Transformer/Smith Chart" module of the *New Low Band Software*, and use the option "with cable losses." You can also use a transmission-line program such as ARRL's *TLW*. Once we know the impedance at the end of the feed lines, we can calculate the L-network component values using the second part of the spreadsheet, called "For system NOT USING current-forcing, or if using 'real' quarter-wave lines" (see **Table 11-2**).

For a 2-element end-fire array we normally feed the back element directly, with the exception of a feed system using the crossfire principle (see Section 3.4.4). We can, however, feed

Table 11-1
Lahlum Spreadsheet for Lossless Cable Case

INPUT DATA		
# of elemts →	1	
Enter Zo →	50.00	Ω
Enter R →	51.00	Ω
Enter X →	20.00	Ω
Enter k →	1.00	
Enter θ →	-90.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	49.02	Ω
X-Par =	-80.65	Ω
Series elem =	2.1	uh
Par elem =	519.6	pF
Rpar =	49.02	Ω
Xpar =	49.02	Ω
Rser =	24.51	Ω
Xser =	24.51	Ω

Table 11-2
Lahlum Spreadsheet for Real Cable Case Including Cable Losses

INPUT DATA		
Enter R →	42.90	Ω
Enter X →	-16.10	Ω
Enter k →	1.000	
Enter θ →	-90.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	48.9	Ω
X-Par =	-78.3	Ω
Series elem =	2.1	uH
Par elem =	534.9	pF
Rpar =	48.94	Ω
Xpar =	48.94	Ω
Rser =	24.47	Ω
Xser =	24.47	Ω

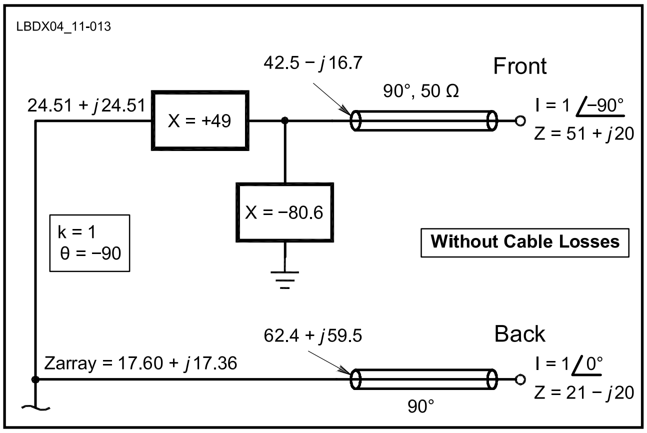


Fig 11-14 — Lewallen/Lahlum feed system for the 2-element end-fed array assuming 90° phase shift and equal feed current magnitude (k = 1). In this case calculations were done assuming zero cable losses.

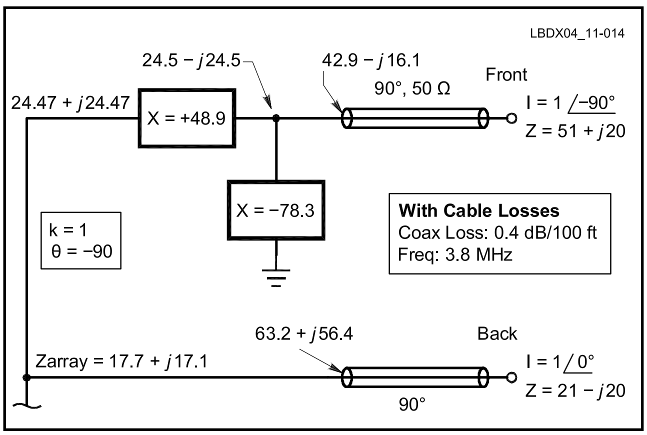


Fig 11-15 — In this case the calculation was done including cable losses. Note the minute difference between these values and those obtained using lossless cables (Fig 11-14).

the front element directly and the back element with a phase shift. In the case of quadrature feeding, this is +90°, which equals +90 – 360 = –270° (see **Table 11-3**). We can achieve the –270° phase shift by designing an L-network to do just that, or we can do this using an L-network that takes care of –90°, followed by a half-wave of feed line, for another 180° or a 180° phase inverter transformer (see Section 3.4.6.3). Use the *Lahlum-Lnetwork.xls* spreadsheet with θ = –270°, resulting in L-network component values of 352 pF and 104.7 µH (Table 11-3). The inductance required is rather high, which is not desirable. If however we replace –270° with –90°, and add a half-wave feed line or 180° phase-inversion transformer at the input of the L-network, we end up with much more attractive component values of 687.2 pF and 5.0 µH (**Table 11-4**).

In many of the phased arrays described in this chapter, the rear element has a very low feed impedance, often with a negative value for the series resistance. At the end of the λ/4 current-forcing feed line, the impedance becomes very high. If we design a feed system that includes an L-network in this branch, we will very often end up with extreme component values. If the reactances are very high, the Q will be high and

Table 11-3
Lahlum Spreadsheet for L-Network Components

L-Networks to be inserted in the back element using θ = +90°, which equals –270°

INPUT DATA		
# of elemts →	1	
Enter Zo →	50.00	Ω
Enter R →	21.00	Ω
Enter X →	-20.00	Ω
Enter k →	1.00	
Enter θ →	-270.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	-119.05	Ω
X-Par =	2498.69	Ω
Series elem =	352.0	pF
Par elem =	104.7	uH
Rpar =	119.05	Ω
Xpar =	-119.04	Ω
Rser =	59.52	Ω
Xser =	-59.52	Ω

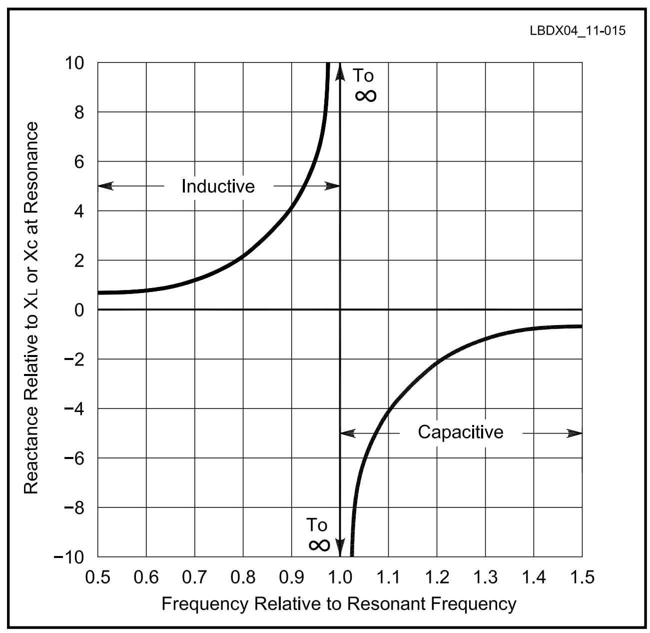


Fig 11-16 — This chart demonstrates that the change in reactance is much greater near resonance than far away.

bandwidth very low. In many cases we will see reactances change from high positive values to high negative values with just a small change in frequency. This situation must be avoided. Therefore it is always best to feed the rear element directly, and the center and front elements via L-networks.

Fig 11-16 shows how the reactance near resonance abruptly changes from inductive to capacitive, and also demonstrates that the relative change in reactance is much greater in that area than farther away. It's a good rule of thumb to design a network where the absolute value of the component

Table 11-4
Lahlum Spreadsheet for L-Network Components
L-Networks to be inserted in the back element using θ = –90°. The remaining 180° shift is obtained with a half-wave line or a 180° phase reversal transformer.

INPUT DATA		
# of elemts →	1	
Enter Zo →	50.00	Ω
Enter R →	21.00	Ω
Enter X →	-20.00	Ω
Enter k →	1.00	
Enter θ →	-90.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	119.05	Ω
X-Par =	-60.98	Ω
Series elem =	5.0	uh
Par elem =	687.2	pF
Rpar =	119.05	Ω
Xpar =	119.05	Ω
Rser =	59.52	Ω
Xser =	59.52	Ω

reactances are not larger than about 250 Ω.

It's a good idea always to work out all the alternative feed systems. You can do these exercises with 50 and with 75-Ω cable. And if there are phasing angles involved that are larger than 180°, you can use a half-wave coax cable or a 180° phase-inversion transformer to do the 180° part (see Fig 11-21 later in this chapter). For each alternative, look at the total array feed impedance and at the L-network component values.

3.4.5.4.1. Using a Different Z_0-Feed Line (Current-forcing Feed Line Impedance)

The example of Fig 11-13 results in a relatively low array feed impedance (~17.6 + *j* 17.4 Ω). We could do the same exercise by using 75-Ω feed lines, and will see a higher feed impedance. How much higher? Robye, W1MK pointed out a simple rule of thumb (not 100% correct but a close indicator):

$$Z_{(feed-75\,\Omega)} = Z_{(feed-50)} \times \frac{75^2}{50} = Z_{(feed-50)} \times 2.25 \quad \text{(Eq 11-9)}$$

In our example the estimated (lossless) impedance, according to this rule is 2.25 × (17.6 + *j* 17.4) = 39.6 + *j* 39.15 Ω. If we do the detailed calculation, the feed impedance, using 75-Ω element feed lines, turns out to be: 39.6 + *j* 39.1, which confirms the simple rule.

In this particular case it would certainly be better using 75-Ω element feed lines. Note that in almost all arrays it turns out that using 75-Ω feed lines achieves a network drive impedance closer to 50 Ω than is the case when using 50-Ω lines.

The 39.6 + *j* 39.1 Ω can be matched pretty well to either a 50 or 75-Ω feed line to the shack. Using a series capacitor with X = –39.1, the feed impedance becomes 39.1 Ω, which results in an acceptable 1.25:1 SWR. Using a parallel capacitor with an inductance of –78 Ω, which results in a feed impedance of 78 Ω, gives a good march to a 75-Ω feed line if you'd like to use that.

3.4.5.4.2. Calculation of Array Feed Impedance

There are two ways of doing this: without losses and with losses. In most cases the lossless way will suffice, but I will explain both ways.

3.4.5.4.3. Without Losses: Using the Lahlum-Lnetwork.xls Spreadsheet

See Section 3.4.5.2 for the formulas, but the top part of the spreadsheet tool does all the work. Let's do it, step by step:

$$R_{par} = \frac{Z_0}{n \times R \times k^2}$$

and

$$X_{par} = \frac{X_{ser}}{1 - K \times \cos\theta}$$

In this case K = 1 and θ = –90° so the formula becomes:

$$R_{par} = \frac{Z_0{}^2}{R}$$

and

$$X_{par} = X_{ser}$$

Using the figures from the above example we have:

$$R_{par} = \frac{50 \times 50}{51} = 49\,\Omega$$

and

$$X_{par} = 49\,\Omega$$

These values must be converted to series equivalent input impedances using the following formulas:

$$R_{ser} = \frac{R_{par} \times X_{par}{}^2}{R_{par}{}^2 + X_{par}{}^2} = \frac{49 \times 49 \times 49}{49 \times 49 \times 49 \times 49} = 24.51\,\Omega$$

$$X_{ser} = \frac{R_{par}{}^2 \times X_{par}}{R_{par}{}^2 + X_{par}{}^2} = \frac{49 \times 49 \times 49}{49 \times 49 \times 49 \times 49} = 24.51\,\Omega$$

The transformation from parallel to serial impedance (and vice versa) can also be calculated using the "RC/RL Transformation" module of the *New Low Band Software*.

Now we connect this impedance in parallel with 62.4 + *j* 59.5 Ω. The result is 17.60 + *j* 17.36 Ω (for this you can use the "Parallel Impedances" module of the *New Low Band Software*).

3.4.5.4.4. Including Losses

Z1 = 51 + *j* 20 Ω

and

Z2 = 21 – *j* 20 Ω

Using the "Coax Transformer/Smith Chart" module of the *New Low Band Software*, we calculate the transformed impedances at the end of 90° long feed lines (VF = 0.66, attenuation = 0.3 dB/100 feet, at F = 3.8 MHz):

Z1' = 42.95 – *j* 16.1 Ω

and

Z2' = 63.2 + *j* 56.4 Ω

Now – *j* 80.6 Ω in parallel with 42.5 – *j* 16.7 Ω = 24.49 – *j* 24.53 Ω. This is in series with + *j* 49 Ω, yielding 24.49 + *j* 24.53 Ω. Now, we connect this impedance in parallel with 63.2 + *j* 56.4 Ω and the result is 17.67 + *j* 17.12 Ω.

This calculation includes cable losses but not the losses from the L-network components. Note that this value is very close to what we calculated in the lossless case.

3.4.5.4.5. Operational Bandwidth of the L-Network Feed System Used on a 2-Element End-Fire Array

The principles of this assessment were covered in Section 3.1.1.

Assessment Procedure

What follows is the detailed procedure used to calculate the performance data of a 2-element end-fire fed (k = 1, θ = 90°) array using the Lewallen L-network feed system, using software and *EZNEC* modeling files available on this book's CD.

1) Run the *EZNEC* modeling file *Ch11-2el-endfire-90-90phase.ez* and note the feed impedance of the front element (this is the element where, at the end of its λ/4 feed line, the L-network will be inserted: Z2 = 53.02 + *j* 19.08 Ω
2) Run *Lahlum-Lnetwork.xls* (1st calculator, case using current forcing) to calculate the value of the L-network. Series arm: 2.1 µH, shunt arm: 592 pF for F = 3.65 MHz and cable Z_0 = 50 Ω.
3) Run the *EZNEC* modeling file *Ch11-2el-endfire-90-90-lewallen-fed.ez* on 3.65 MHz where the abovementioned L-network values have been entered under the L-networks.
4) Temporarily remove the second L-network (the V2-V3) network, which is the input impedance matching network, and change the array feeding source from V2 to V3.
5) Run the program and note the array input impedance (under Scr Dat): Z_{in} = 17.56 + *j* 17.27 Ω
6) We must now design an L-network to convert this impedance to 50 Ω. Run "L-Network Design" (part of *New Low Band Software*) and calculate L-network to match to 50 Ω: L series = 0.29 µH and C parallel = 1185 pF
7) These were the values of the L-network V2-V3 that you just removed from the L-network page. Just wanted to explain how we came to these values.
8) Run this model in sweep mode (3.5 to 3.8 MHz, 5 or 10 kHz steps) and save the data.
9) Use these data with the *LBDXView* software (the manual is on the CD that comes with this book — *LBDXView-manual.pdf*) to generate the graphs and patterns shown in **Fig 11-17** and **Fig 11-18.**

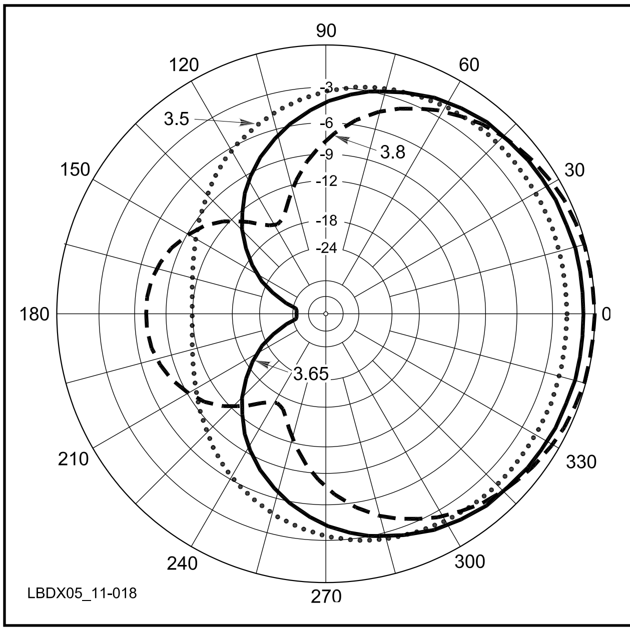
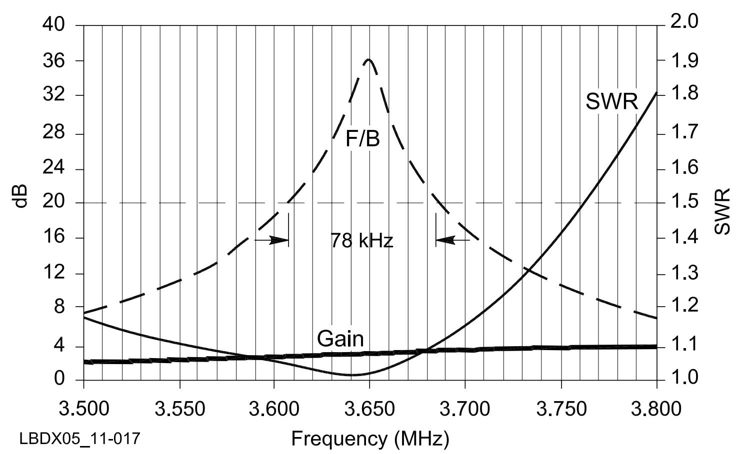


Fig 11-18 — Horizontal radiation pattern (at 20° wave angle) for the 2-element end-fire array (θ = 90°, k = 1), fed with a single L-network. (Plot generated with W8WWV's *LBDXView* software.)

Fig 11-17 — Performance characteristics of the 2-element end-fire array (θ = 90°, k = 1). The input of the array was matched to 50 Ω by an L-network (0.29 µH and 1200 pF). (Plot generated with W8WWV's *LBDXView* software.)



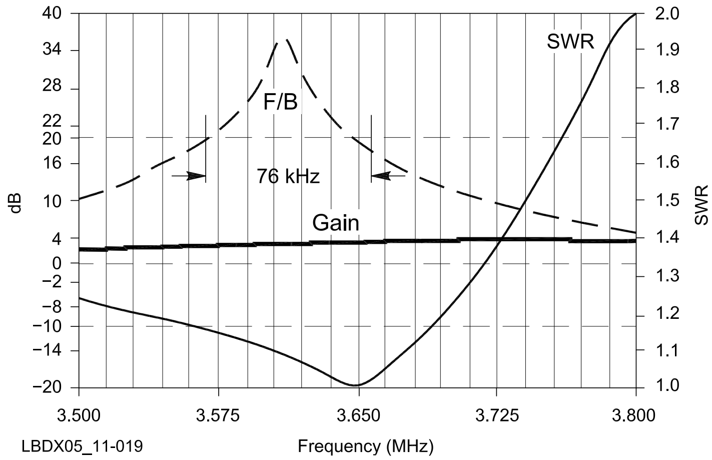


Fig 11-19 — Performance characteristics of the 2-element end-fire array (λ/4 spacing) with 105° phase shift. The input of the array was matched to 50 Ω by an L-network (0.21 µH and 1234 pF). (Plot generated with W8WWV's *LBDXView* software.)

Results

The operational bandwidth is clearly limited by the directivity, which is 20 dB or better over merely 78 kHz. In those 78 kHz the gain varies by not less than 0.8 dB, which is considerable. F/B is clearly the bandwidth limiting factor.

I also did the exercise for a 2-element end-fire array spaced 90° but fed 105° out of phase. The results are shown in **Fig 11-19** and **Fig 11-20**. The operational bandwidth is also limited to approximately 76 kHz (file: *Ch11-2el-endfire-90-105deg-lewallen-fed.ez*). Note that although we are using elements of the same length as in the above case, the F/B now peaks approximately 50 kHz below the design frequency. Also in this case the operational bandwidth was limited by its directivity.

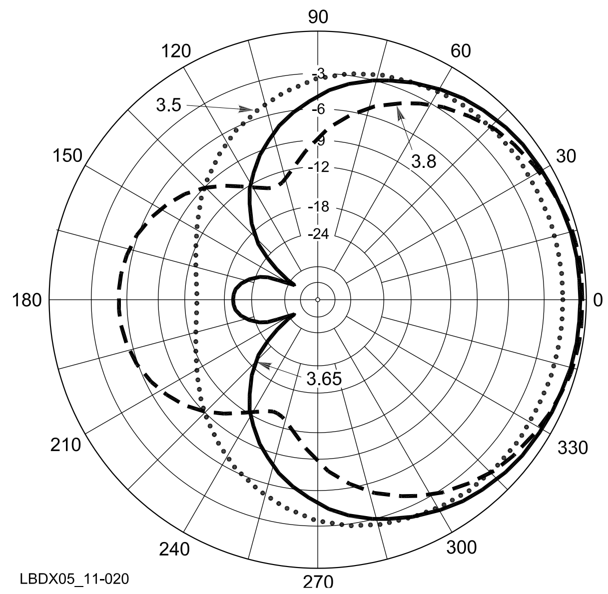


Fig 11-20 — Horizontal radiation pattern (at 20° wave angle) for the 2-element end-fire array (θ = 105°, k = 1), fed with a single L-network. (Plot generated with W8WWV's *LBDXView* software.)

3.4.5.5. Tutorial

The "2 El and 4 El Vertical Arrays" module of the *New Low Band Software* is a tutorial and engineering program that takes you step by step through the design of a 2-element cardioid type phased array (and also the famous 4-element square array, which is described later). The results as displayed in that program will be slightly different from the results shown here, since the software uses lossless feed lines.

3.4.5.6. The Quadrature Fed Four Square

Let's assume we have obtained the following feed impedance values through modeling:

Z1 = 61.7 + *j* 59.4 Ω (at the front element, fed with a –180° current phase angle)

Z2 = Z3 = 41 – *j* 19.30 Ω (the center elements, both fed with a –90° current phase angle)

Z4 = 0.4 – *j* 15.40 Ω (at the 0° element, the back element)

Note that the –0.4 Ω resistive part of the feed impedance Z4 means that the antenna is not "taking" power from the feed network, but rather delivering power to it (this is "excess" power that it has by mutual coupling to the other elements). Note also that in a lossless calculation such a negative (usually very low) value will show up as a negative (high) value at the end of the λ/4 feed line. If, however, the nominal value is low, and the cable attenuation is taken into consideration, a small negative R-value at the antenna end can turn up a high positive R-value at the other end. This is due to the effect of cable loss.

Note also that in this array, as is the case in most multi-element arrays, the SWR of the feed line going to the back element is *very* high, which normally causes a lot of additional power loss (due to SWR). But in this case, the power flow is so small into this feed line to the back element that it just does not matter much. High SWR but no power flow results in very little power being lost. If you look at the resistive part of the equivalent parallel resistance (several thousand ohms) at the end of the λ/4 feed line, any reduction in the exact value due to losses would cause very little increase in input power to get the same current to flow into the loads. Which means that you *can* use the lossless model to calculate the feed system.

As explained for the 2-element end-fire array we can design the feed system in different ways. The most common approaches are:

1) Feeding the back element directly, the front element

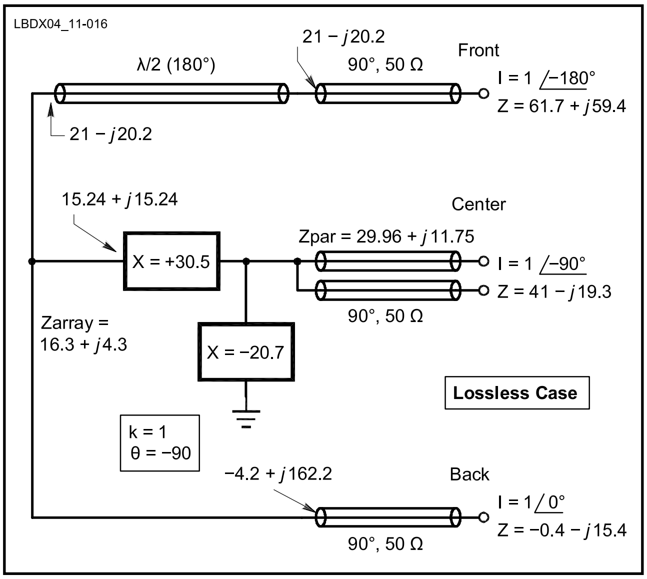


Fig 11-21 — Feed system for the quadrature-fed Four Square using 50 Ω current-forcing feed lines.

Table 11-5
Lahlum Spreadsheet for Determining the Network as shown in Fig 11-21

INPUT DATA		
# of elemts →	2	
Enter Zo →	50.00	Ω
Enter R →	41.00	Ω
Enter X →	-19.30	Ω
Enter k →	1.00	
Enter θ →	-90.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	30.49	Ω
X-Par =	-20.73	Ω
Series elem =	1.3	pF
Par elem =	2021.4	uH
Rpar =	30.49	Ω
Xpar =	30.49	Ω
Rser =	15.24	Ω
Xser =	15.24	Ω

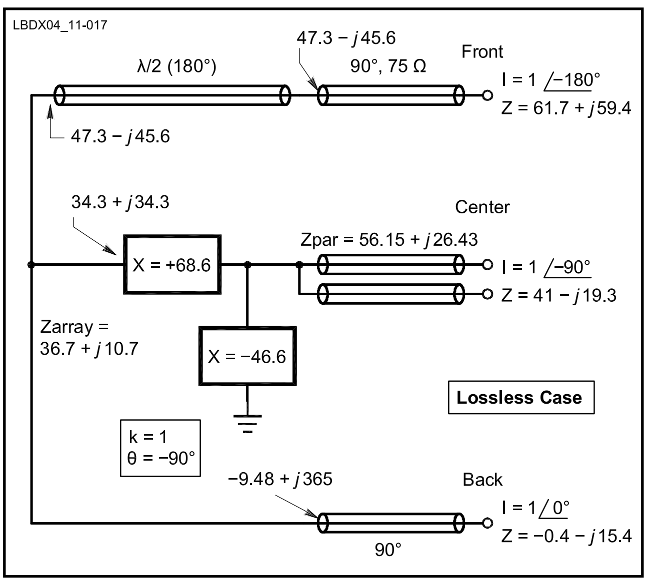


Fig 11-22 — Same feed system as in Fig 11-21 but using 75 Ω feed lines. This results in a significantly higher feed impedance and is recommended.

Table 11-6
Lahlum Spreadsheet for Determining the Network as shown in Fig 11-22

INPUT DATA		
# of elemts →	2	
Enter Zo →	75.00	Ω
Enter R →	41.00	Ω
Enter X →	-19.30	Ω
Enter k →	1.00	
Enter θ →	-90.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	68.60	Ω
X-Par =	-46.64	Ω
Series elem =	2.9	pF
Par elem =	898.4	uH
Rpar =	68.60	Ω
Xpar =	68.60	Ω
Rser =	34.30	Ω
Xser =	34.30	Ω

via a 180° phase shift line (λ/2) and the central elements via an L-network "from the back element," all of this with 50-Ω λ/4 feed lines (see **Fig 11-21**).

2) Identical to the approach above, but with 75-Ω feed lines (see **Fig 11-22**).

There is no absolute need to feed the back element directly and the middle and front via a phasing system. One could to feed the center elements directly, the front element with a –90° phasing system (L-network) and the back element with a –270° phasing system. In a third alternative you would feed the front element directly, the center elements with –270° phase shift and the back element with –180° phase shift.

Each solution will have different L-network component values and a different array input impedance and different values for the L-network components. We can then select the network with the most manageable network component values and the most attractive feed impedance (avoid values below 10 Ω).

It's a good idea always to work out all the alternative feed systems. In the case of a Four Square array you can use either the branch to the back element as the "reference" branch which is fed directly, or the branch to the center element or even the branch to the front element. You can do these exercises with 50 and 75-Ω cable. And if phasing values are larger than 180°, you can use a half-wave coax cable (or a 180° phase-inversion transformer). See Figs 11-21 and 11-22.

Table 11-5 and **Table 11-6** show the results for a 50-Ω and for a 75-Ω system impedance, if we apply R = 41, X = –19.3, n = 2 and θ = –90°. It is obvious that the 75-Ω solution is the better one as it results in a much more "convenient" array feed impedance.

3.4.5.7. The Example of a Three-In-Line End-Fire Array with Binomial Current Distribution.

If the current magnitude of the element fed through the L-network needs to be different from the magnitude of the current to the other elements in the array, the appropriate current can be achieved by specifying the correct k-value in the *Lahlum-Lnetwork.xls* spreadsheet or in the formulas from Section 3.4.5.2.

Let's work out the example of the 3-element in-line end-fire array, each spaced λ/4, fed in 90° increments, but center element fed with double current magnitude:

Front element: Z1 = 76.1 + j 51 Ω
Center element: Z2 = 26.3 – j 0.4 Ω
Back element: Z3 = 15 – j 22.6 Ω

In the Excel spreadsheet (**Table 11-7**) we enter k = 2, which means that the element(s) fed through the L-networks will have twice the current magnitude as the reference element in the array.

Fig 11-23 shows the feed system (lossless calculation) This example uses 75-Ω feed lines, and results in an array feed impedance of 20.3 + j 13 Ω. Using 50-Ω feed lines, the array impedance would be approximately 2.25 times lower (certainly not the best solution!). Hence 75 Ω is recommended.

If we included the losses, the real part of the feed impedance would have been slightly higher (you need more driving power into the feed system to get the same amount of radiated power).

Table 11-7
Lahlum Spreadsheet for the 3-Element End-Fire Array

INPUT DATA		
# of elemts →	1	
Enter Zo →	75.00	Ω
Enter R →	26.30	Ω
Enter X →	-0.40	Ω
Enter k →	2.00	
Enter θ →	-90.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	106.94	Ω
X-Par =	-106.13	Ω
Series elem =	4.5	pF
Par elem =	394.8	uH
Rpar =	53.47	Ω
Xpar =	106.94	Ω
Rser =	42.78	Ω
Xser =	21.39	Ω

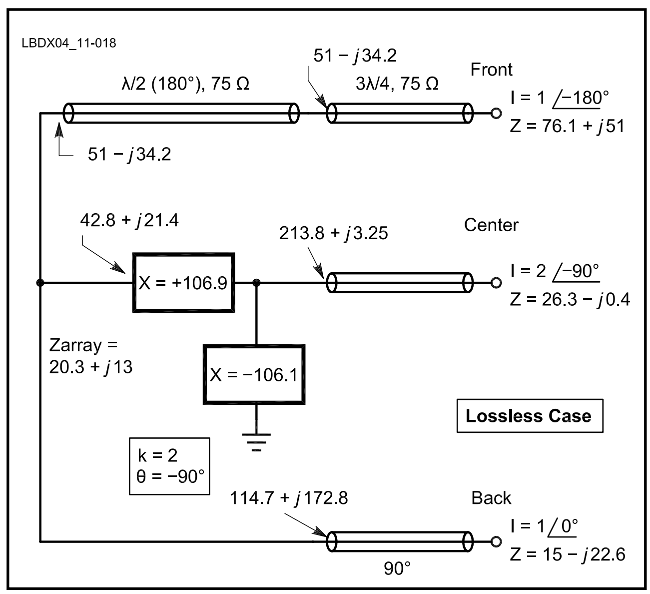


Fig 11-23 — Classic 3-in-line configuration with direct feed to the back element. Specifying k = 2 in the spreadsheet program allows us to double the feed current magnitude without having to resort to parallel feed lines.

3.4.5.7.1. Calculating the Array Input Impedance

In order to prove that the real part of the input impedance would indeed be higher, we will carry out a calculation in a "real world" environment.

Note that in order to reach the center of the array (which is necessary if you want to switch directions) you will require ¾ λ feed lines, as the element spacing is λ/4 (see Fig 11-90 later in this chapter).

Let's do some impedance calculations using the applicable modules of the *New Low Band Software*. Using the "Coax Transformer/Smith Chart" module we first calculate the impedances at the end of our ¾ λ feed lines (⅝ λ feed line to front element). As explained earlier, lossless calculation will do

Front element: Z1 = 76.1 + j 51 Ω → 79.1 + j 27.3 Ω
Center element: Z2 = 26.3 – j 0.4 Ω → 180.7 + j 2.2 Ω
Back element: Z3 = 15 – j 22.6 Ω → 128.9 + j 134.9 Ω

I used 75-Ω coax (VF = 0.8) with a loss of 0.2 dB per 100 ft for the calculation (F_{design} = 1.8 MHz). Next we calculate the parallel impedance caused by the parallel reactance X_{p1} (using the module "Parallel Impedances (T-Junction)"):

At X_{p1} we calculate – j 106 Ω in parallel with 180.7 + j 2.2 Ω which gives 46.8 – j 79.2 Ω. Adding + j 106.9 Ω in series totals: 46.8 + j 27.78 Ω.

For the back element we have an impedance of 128.9 + j 134.91 Ω at the end of the ¾ λ feed line. For the front element we have an impedance of 79.1 + j 37.3 Ω at the end of the ⅝ λ feed line.

All three in parallel: Z_{tot} = 24.5 + j 14.9 Ω. This is exactly what we expected. Compared to the value we calculated without losses (20.3 + j 13) the feed impedance gets a little higher when losses are included.

3.4.5.7.2. Input Impedance at the Input Side of the L-Network

If we neglect the effect of losses in the feed lines, we can also calculate the input impedance using the R_{ser} and

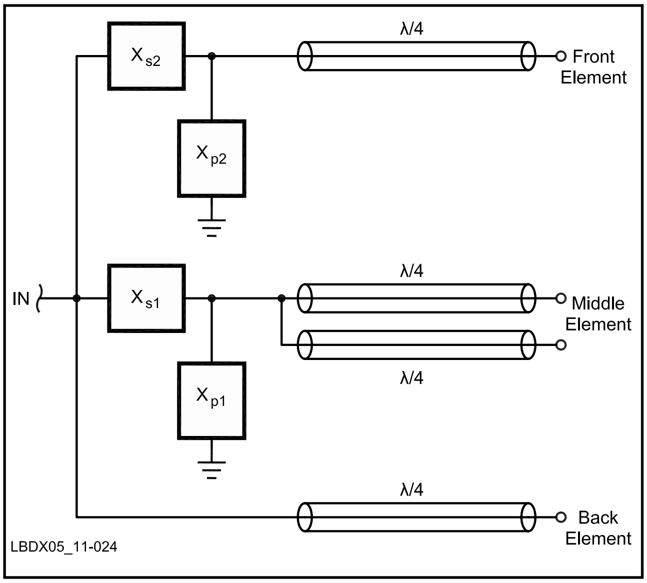


Fig 11-24 — The Lahlum/Lewallen feed method applied to a Four Square array. The back-element is the reference element (see text for details).

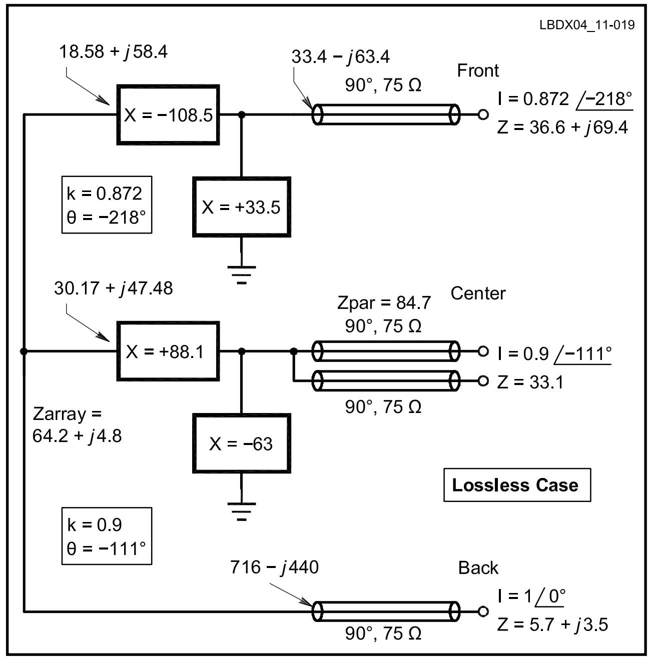


Fig 11-25 — Lahlum/Lewallen feed network for the WA3FET-type Four Square array. This version uses 75 Ω quarter-wave feed lines, and the back element is the element that is directly fed.

X_{ser} values from the *Lahlum-Lnetwork.xls* worksheet: R_{ser} = 42.78 Ω and X_{ser} = 21.39 Ω

These values are somewhat lower than those calculated considering cable losses which are 46.8 + *j* 27.78 Ω. The difference is relatively high because in this case we are using 270° feed lines, which represent substantial loss. But for all practical purposes the lossless calculations are adequate.

3.4.5.8. Using Lahlum's New Formulas for Phase Angles Other Than 90°

So far we have used θ = –90° in the generic formulas shown in Section 3.4.5. Robye Lahlum, W1MK, developed the formulas that allow us to use the L-network to obtain a phase shift other than –90° with different current magnitudes, and he decided to share them with me for publication in this book, for which I am very grateful! **Fig 11-24** shows the basic idea.

As we will see in Section 4.7.2 it appears that we can significantly improve the performance of a Four Square by not feeding the element in quadrature (in 90° steps) and with equal current magnitudes. Jim Breakall, WA3FET, developed such an optimized version of a Four Square array.

In **Fig 11-25** the back element is the reference element, with θ = 0° and k = 1. The two center elements are fed with a phase angle of –111° and a current magnitude ratio of k = 0.9, the front element with θ = –218° and k = 0.872. In this example I used the following feed impedances for a full-size quarter-wave spaced Four Square, including 2 Ω ground-loss resistance:

Z-front element: 36.6 + *j* 69.4 Ω
Z-center elements: 33.1 Ω
Z-back element: 5.7 + *j* 3.5 Ω

The component values are computed in the *Lahlum-Lnetwork.xls* spreadsheet. See **Table 11-8**, based on a 75-Ω cable impedance. If I had used 50-Ω feed lines, the array impedance would have been approximately 2.25 times lower than shown in Fig 11-21 or approximately 28 – *j* 2.2 Ω.

As explained above, the formulas used in the spreadsheet assume zero-loss transmission lines. In most cases this will give a result accurate enough to tell you what the approximate value of the components of the L-network will be. You can however also do the exercise including cable losses. In Section 3.4.5.4.3, I explained how the phasing networks can be calculated using the *Lahlum-Lnetwork.xls* calculation tool in the case we want to include cable losses.

3.4.5.8.1. Calculating the Array Impedance

From the spreadsheet we read the input impedance to both L-networks (see Table 11-8 and Fig 11-25).

Z1 (to center elements) = 30.17 + *j* 47.48 Ω

Z2 (to front element) = 18.58 – *j* 58.4 Ω

Let's now use "Coax Transformer/Smith Chart" module of the *New Low Band Software* to calculate the impedances at the end of the λ/4 feed line going to the back element (fed without phasing):

Back element: Z = 5.7 + *j* 3.5 Ω. At the other end of the current-forcing feed line: Z3 = 716 – *j* 440 Ω.

All three in parallel (calculated with the "Parallel Impedances" module of the *New Low Band Software*: Z_{tot} = 64.2 – *j* 4.8 Ω

3.4.5.9. Operational Bandwidth of the Four Square L-Network Feed System

The principles of this assessment are covered in Section 3.1.1.

Single L-Network Case: Assessment Procedure

What follows is the detailed procedure how to calculate the performance data of a quadrature fed (k = 1, θ = 90°) Four Square array using the Lewallen L-network feed system (with

Table 11-8
Lahlum Spreadsheet for Calculating the Networks to be Used in the Two Branches of the Feed System (Fig 11-25)

INPUT DATA		
# of elemts →	2	
Enter Zo →	75.00	Ω
Enter R →	33.10	Ω
Enter X →	0.00	Ω
Enter k →	0.900	
Enter θ →	-111.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	88.14	Ω
X-Par =	-63.04	Ω
Series elem =	3.7	uh
Par elem =	664.7	pF
Rpar =	104.90	Ω
Xpar =	66.65	Ω
Rser =	30.17	Ω
Xser =	47.48	Ω

INPUT DATA		
# of elemts →	1	
Enter Zo →	75.00	Ω
Enter R →	36.60	Ω
Enter X →	69.40	Ω
Enter k →	0.872	
Enter θ →	-218.00	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	-108.51	Ω
X-Par =	33.46	Ω
Series elem =	386.2	pF
Par elem =	1.4	uH
Rpar =	202.12	Ω
Xpar =	-64.31	Ω
Rser =	18.58	Ω
Xser =	-58.40	Ω

a single L-network as shown in Fig 11-21), using exclusively software and *EZNEC* modeling files available on this book's CD:

1) Run the *EZNEC* modeling file *Ch11-4SQ-quadfed-Z2-Z3legs.ez* (refer to Section 3.4.6.4.1.).
2) Note Z2 = 53.7 + *j* 22.5 (the impedance at the end of the parallel connected current-forcing feed lines going to the center elements, see Fig 11-41 later in this chapter).
3) Run *Lahlum-Lnetwork.xls* (2nd calculator, case not using current forcing) to calculate the value of the single L-network (see Fig 11-22).
4) Use F = 3.65 MHz (band center) and enter the Z2 impedance data.
5) The values of the L-network components are: Series element: 2.75 µH, parallel element = 981 pF.
6) Run the *EZNEC* modeling file *Ch11-4sq-quadfed-lewallen.ez* on 3.65 MHz and enter the above L-network values.
7) Note the array input impedance = 32.57 + *j* 10 Ω (from Scr Dat).
8) Run "L-Network Design" (part of *New Low Band Software*) and calculate L-network to match to 50 Ω: L series = 0.6 µH and C parallel = 638 pF.
9) Open the *EZNEC* modeling file *Ch11-4sq-quadfed-lewallen+Zmatch.ez* where the matching L-network is included.
10) Run this model in sweep mode (3.5 to 3.8 MHz, 5 kHz steps) and save the data.
11) Use these data with the *LBDXView* software to generate the graph and patterns shown in **Figs 11-26** and **11-27**.

We can also do the exercise for a Four Square fed with the optimized quadrature feed system (k = 0.85). In this case the center elements are fed with a current magnitude which is 85% of the current magnitude of the front and back elements (k = 0.85). The L-network values now are L = 3.24 µH and C = 877 pF. The 50-Ω matching L-network values are L = 0.44 µH and C = 453 pF. (See *Ch11-4sq-simple-opt-lewallen+Zmatch.ez*.) **Fig 11-28** shows the gain, F/B, SWR data, and **Fig 11-29** the radiation patterns.

Fully Optimized (Two L-Networks) Assessment Procedure

In this case two L-networks are used (see Fig 11-25). The array data for a WA3FET configuration are:

- Back element: I1 = 1 ∠0°
- Center elements: I2 = I3 = 0.9 ∠–111°
- Front element: I4 = 0.872 ∠–218°

Procedure:

1) Model this with *EZNEC* including four quarter-wave feed lines (75-Ω, no loss).
2) Enter the sources as voltages (of course!):
   - Back element: V1 = 100 V ∠0°
   - Middle elements: V2 = V3 = 87.2V ∠–216°
   - Front element: V4 = 90 V ∠–111°
3) *EZNEC* will give you the impedances at the end of the current-forcing feed lines:
   - Back: Z1 = 610 + *j* 179
   - Middle (parallel) = Z2 = Z3 = 78.1 + *j* 15.7
   - Front = Z4 = 45.3 – *j* 62.5
4) Use the impedances Z1 and Z2 (= Z3) to calculate the L-network values using *Lahlum-Lnetwork.xls*.
5) The networks to be included in the model are:
   - to the center elements: series element: 3.68 µH, shunt element: 832 pF
   - to the front element: series element: 470 pF, shunt element: 1.4 µH
6) Include these elements into the *EZNEC* model (see *Ch11-4sq-Lnetw-fed-WA3FET.ez*).
7) The feed system input Impedance is 57.6 – *j* 6.8 Ω (SWR = 1.2:1), and does not require an additional

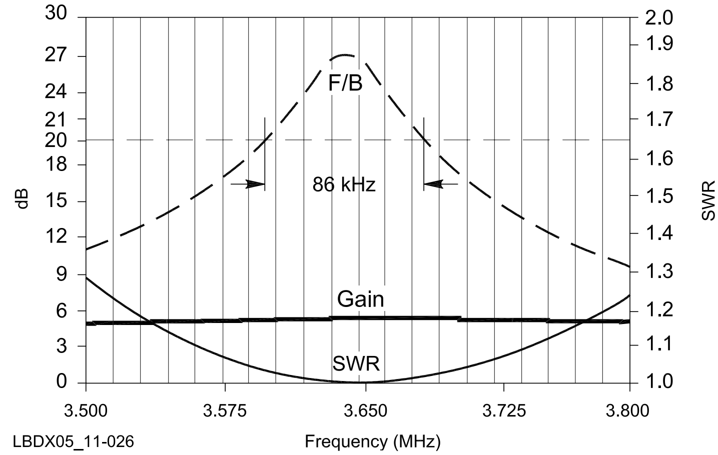


Fig 11-26 — SWR, gain and F/B results for the quadrature fed Four Square (k = 1) using a single L-network to obtain the 90° phase shift to the center elements (see Fig 11-23). A second L-network was included at the feed port to bring the SWR to 1:1 at the design frequency (3.65 MHz). Directivity-wise the bandwidth is limited to approximately 86 kHz. (Plot generated with W8WWV's *LBDXView* software.)

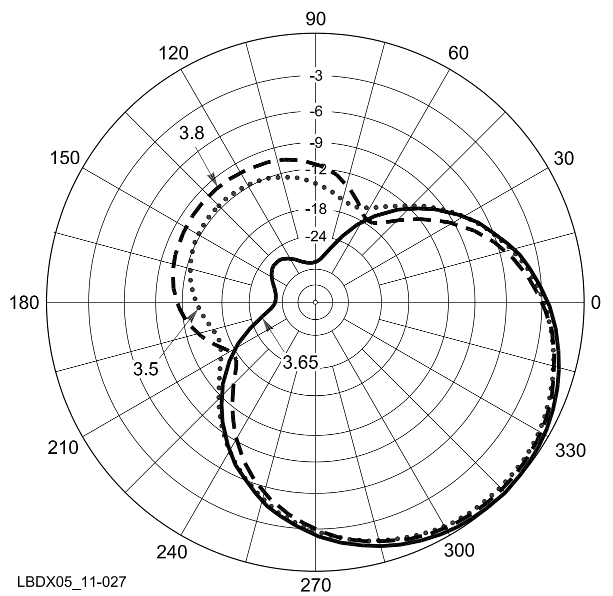


Fig 11-27 — Horizontal radiation patterns at mid-band and at the two band edges for the Four Square with θ = 90 and k = 1.0, using a single L-network. (Plot generated with W8WWV's *LBDXView* software.)

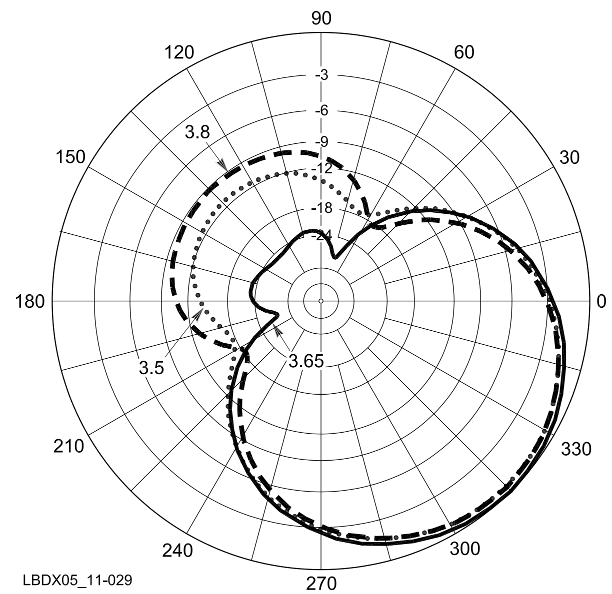


Fig 11-29 — Horizontal radiation patterns at mid band and at the two band edges for the Four Square with θ = 90 and k = 0.85, using a single L-network. (Plot generated with W8WWV's *LBDXView* software.)

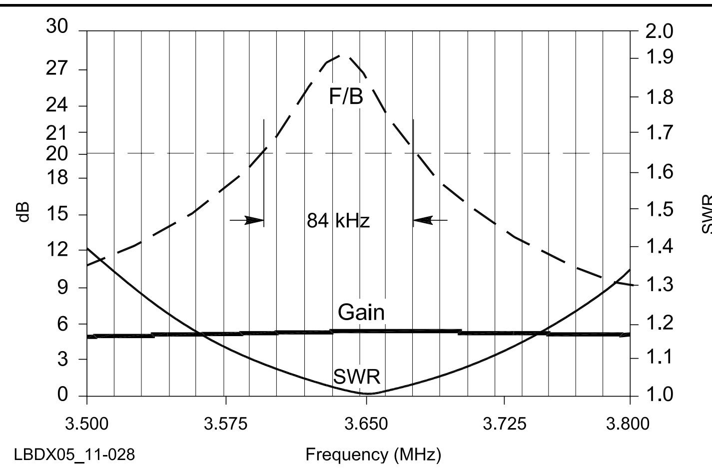


Fig 11-28 — SWR, gain and F/B results for the quadrature fed Four Square (k = 0.85) using a single L-network to obtain the 90° phase shift to the center elements (see Fig 11-22). A second L-network was included at the feed port to bring the SWR to 1:1 at the design frequency (3.65 MHz). Directivity-wise the bandwidth is limited to approximately 84 kHz. (Plot generated with W8WWV's *LBDXView* software.)

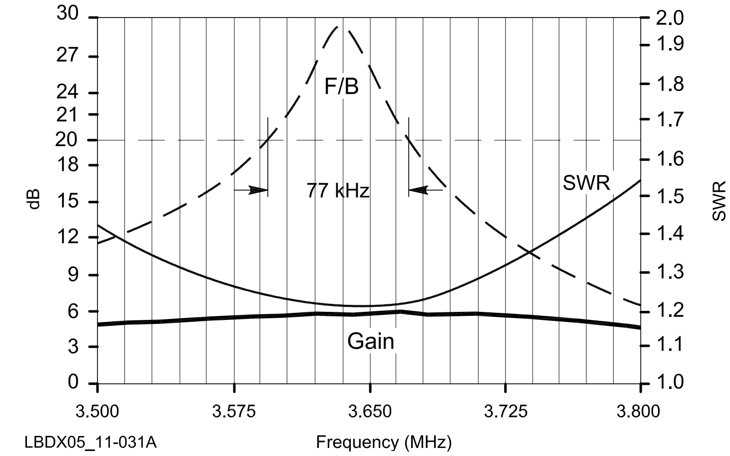


Fig 11-30 — SWR, gain and F/B results for the WA3FET optimized Four Square Directivity-wise the bandwidth is limited to approximately 77 kHz. In this case no SWR matching L-network was added. (Plot generated with W8WWV's *LBDXView* software.)

L-network for a good match to a 50 Ω feed line.

8) Run this model in sweep mode (3.5 to 3.8 MHz, 5 kHz steps) and save the data.

9) Use these data with the *LBDXView* software to generate the data shown in **Fig 11-30** and **Fig 11-31** (pattern).

Conclusion

Whatever the configuration of a Four Square, fed by the L-network (Lewallen) feed system, the operational bandwidth remains approximately 80 kHz (on 80 meters).

3.4.6. Using a 90° Lumped Constant Hybrid Coupler

Fred Collins, W1FC, developed a feed system similar to the Lewallen system in that it uses current-forcing λ/4 feed lines to the individual elements. There is one difference, however: instead of using an L network, W1FC uses a 90° *hybrid network* (also called *hybrid coupler* or *hybrid splitter* or *3 dB hybrid)*, as designed by Reed Fisher, W2CQH and shown in **Fig 11-32** (ref 993). There are many different types of 90° hybrid networks, but this particular coupler (the *lumped coupled 90° hybrid*) we work with in this section of the book is also called the "coupled line" hybrid coupler as the two coil windings on the same core provide the required cross coupling.

The big advantage of using a 90° hybrid as the heart of a feed system for quadrature fed arrays is its intrinsic high operational bandwidth. Under certain circumstances the hybrid maintains a perfect 90° phase shift (theoretically) over the entire frequency spectrum (in practice over a very large bandwidth). In a nutshell: a well designed hybrid coupler feed system can cover the entire 80-meter band (8% bandwidth) with excellent operational characteristics, while this is totally impossible with all other feed systems that have been described

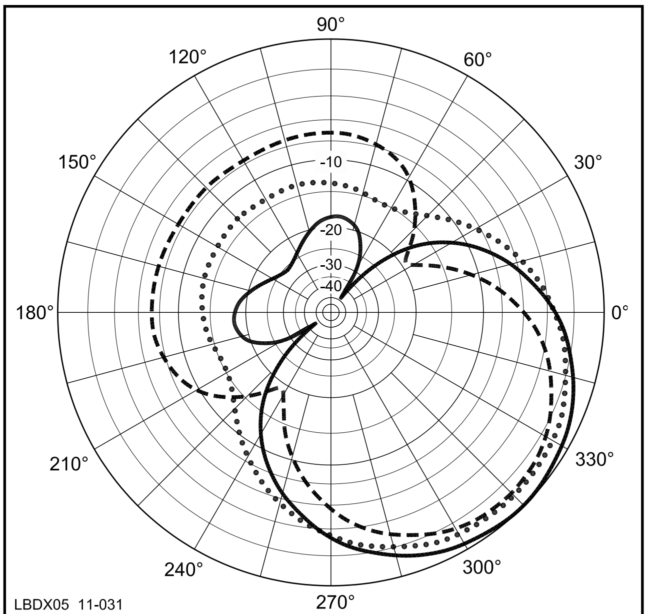


Fig 11-31 — Horizontal radiation patterns at mid-band and at the two band edges for the Four Square using the WA3FET configuration and using two L-networks (see Fig 11-25). (Plot generated with W8WWV's *LBDXView* software.)

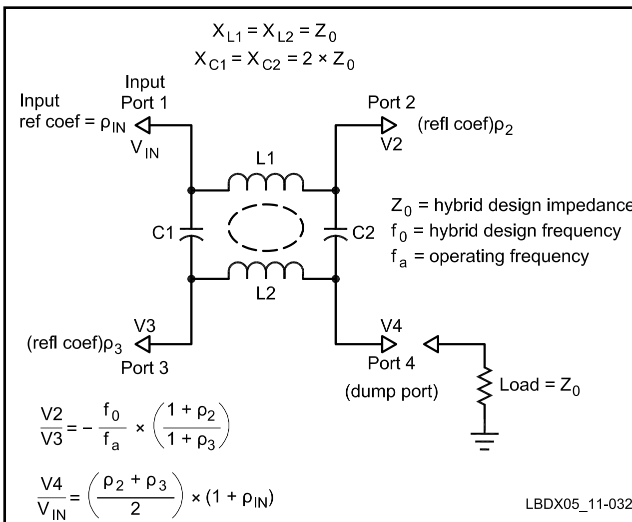


Fig 11-32 — Schematic and formulas for the so-called 3 dB hybrid coupler as described by Reed Fisher, W2CQH, in the January 1978 issue of *QST*. The formulas show the option to operate the hybrid on a frequency (f_a) different from the hybrid design frequency (f_0) as explained in the text.

so far (for example, the very flexible L-network system).

The behavior of the coupler depends 100% on the load impedances at ports 2 and 3, and port 4 to a lesser degree. These load impedances can be characterized in relation to the system design impedance (Z_0) as complex reflection coefficients ρ_2, ρ_3 and ρ_4. The complete mathematical analysis done by W1MK is based on these reflection coefficients and is available on the CD as *90 degree hybrid coupler design formulas.pdf*. The spreadsheet calculator *4sq-hyb-w1mk.xls* (also available on the CD) will calculate anything you may want to know about the behavior of this hybrid. See **Fig 11-33**.

Experimenting with the Calculator

Let's work out a few examples using the spreadsheet. **Table 11-9** shows a number of key cases. To start, enter the same frequency for f_a (operating frequency) and f_o (hybrid design frequency), and 50 Ω as hybrid design impedance.

Case 1: Both port 3 and port 4 are loaded with pure resistive 50-Ω impedances. In this (ideal) case, k = 1 (k being the ratio of the magnitude of the voltage at port 2 to the magnitude of the voltage at port 3), and the phase shift θ = –90°. At the same time no power is dumped in the port 4 load and no power is returned to port 1. Return loss at port 1 = ∞.

Case 2: Both port 2 and port 3 are loaded with equal resistive impedances. In this case k is also 1 and θ = –90° (what we wanted) but now we have some reflected and lost in port 4 (in the dummy load). The input SWR remains perfect.

Case 3: Port 2 and port 3 are loaded with unequal resistive loads. In this case the phase shift θ remains –90°, but the k-factor is no longer 1. The voltage is highest in the branch where the load resistance has the highest value.

Case 4: Same case as above, but with f_o/f_a correction (see below), which results in k = 1. Note that in this example $Z_0^2 \sim Z2 \times Z3$, which is why the port 4 dump power is very low (–34.7 dB). In general we can say that no power will be dissipated at port 4 when $Z_0 = Z2 \times Z3$. Note that Z2 and Z3 need not be real to satisfy this condition.

Case 5: Port 2 and port 3 are terminated in equal complex loads, which means ρ2 = ρ3. In this case θ = 1, k = 1 but part of the power is reflected to the port 4 load resistor. Input return loss is ∞, which means no reflection to port 1. This condition is also called "matched mismatches" (the loads are matched, but they are mismatched with respect to Z_0).

Case 6: Port 2 and port 3 are terminated with unequal complex load impedances. In this case the phase shift θ is no longer –90°, and k differs from 1, unless the reflection coefficients of Z2 and Z3 meet specific requirements.

HYBRID COUPLER DESIGN (by W1MK)			
INPUTS			
ENTER fa (antenna frequency) →	3.650	MHz	
ENTER fo (hybrid frequency) →	3.650	MHz	
ENTER Zo Design impedance hybrid →	50.00	Ω	
	Real Part	Imag part	
ENTER Impedance load PORT 2 (R2) →	52.00	27.00	Ω
ENTER Impedance load PORT 3 (R3) →	74.00	-20.00	Ω
RESULTS			
fo/fa =	1.000		
Hybrid L value (uH) =	2.18	µH	
Hybrid C value (pF) =	436	pF	
Ratio Voltage magnitude Port2/Port3 (k) =	0.910		
Phase angle Voltage port 2 vs. port 3 =	-71.43	°	
Power in Port 4 (vs. Pwr in Port 1) =	-15.8	dB	
Real part input impedance (port 1) =	71.87	Ω	
Imagainare part input impedance (port1) =	9.73	Ω	
Return loss (port 1) =	-14.2	dB	
SWR =	1.49		

Fig 11-33 — The hybrid coupler spreadsheet developed by W1MK calculates the values of L and C for the given design frequency f_0. More important, it shows the output details for port 2 and 3 (the voltage magnitude ratio and the phase angle) and the amount of power dumped in the load on port 4. It also includes the port 1 input impedance, as well as the SWR and return loss at that port.

Adjusting the k-Value (Magnitude V2/V3) by Changing f_o/f_a

With ports 2 and 3 terminated in Z_0, the frequency performance of the hybrid results in unequal voltage magnitudes as shown in **Fig 11-34**. The graph shows that, by varying the operating frequency (off the design frequency f_o) one can obtain a perfect 90° split with unequal voltage magnitude at ports 2 and 3. This is a characteristic that will be used in the design of a "compensated" hybrid splitter feed system (see Sections 3.4.6.5, 3.4.6.6 and 3.4.6.7.

You have to be careful how you interpret the graph in Fig 11-34.

- If you have a hybrid coupler designed for a given frequency, and you want to obtain a voltage magnitude split different from 1:1, you will have to redesign the hybrid for a *higher* frequency if you want the –90° port to deliver more voltage (lower if the –90° port is to deliver less voltage).
- If you have a hybrid designed for a given frequency, and you use it at a higher frequency, the –90° port will deliver

Table 11-9
Summary of Key Cases Using a 90° Hybrid Coupler

Case	*R2*	*X2*	*R3*	*X3*	Z_0	*Port 4*	*Port 1*	*k*	θ	*fa*	*fo*	*Remarks*
1	50	0	50	0	50	∞	∞	1	–90	3.65	3.65	both 50 Ω
2	68	0	68	0	50	–16.3	∞	1	–90	3.65	3.65	identical real impedances
3	54	0	43	0	50	–34.7	–25	1.155	–90	3.65	3.65	different real impedances
4	54	0	43	0	50	–34.7	–25	1	–90	3.65	3.08	as above but with f_o/f_a compensation
5	33	10	33	10	50	–12.5	∞	1	–90	3.65	3.65	identical complex imped. (matched mismatches)
6	53.7	22.5	60.7	–36	50	–17.1	–12.5	0.905	–66.8	3.65	3.65	random impedances

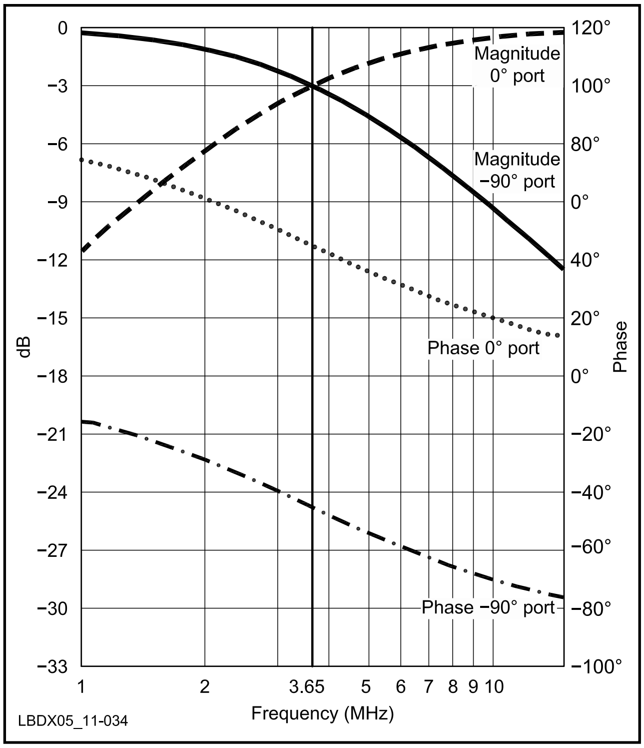


Fig 11-34 — Splitting characteristics of the hybrid coupler described in Fig 11-32, provided the split ports (2 and 4) are terminated in the design impedance (Z_0 hybrid). Note that different magnitude split radios can be obtained if the splitter is used on a frequency different from the design frequency (3.65 MHz in this example).

less voltage than the 0° port.

- The ratio of the split is equal to the ratio of the frequencies involved. We normally define that ratio as the k ratio, being the ratio of the voltage at the –90° port vs the voltage at the 0° port (the reference).

You can verify this by running the *4sq-hyb-w1mk.xls* spreadsheet, entering 50 Ω as loads Z2 and Z3 and varying f_a and/or f_o.

The hybrid spreadsheet calculator (*4sq-hyb-w1mk.xls*) will be very helpful as well when we design the two hybrid optimizing systems described in Sections 3.4.6.5, 3.4.6.6 and 3.4.6.7.

Note that most 90° hybrid couplers are used with a port 4 (dump port) dummy load with an impedance Z4 = Z_0-hybrid. The mathematical model made by W1MK makes it possible though to see what happens to the hybrid's performance if this port 4 load is different from Z_0-hybrid.

For example, using a 75-Ω dummy load on a hybrid with Z_0-hybrid = 50 Ω can, depending on the loads at port 2 and 3, cause more important changes. The *4sq-hyb-w1mk.xls* spreadsheet makes it possible to play with all the input parameters and see for yourself what happens to the required outputs.

3.4.6.1. Hybrid Construction

The values of the 90° hybrid coupler components (see Fig 11-32) are:

$X_{L1} = X_{L2} = 50\ \Omega$ (system impedance)
$X_{C1} = X_{C2} = 2 \times 50\ \Omega = 100\ \Omega$

For f_o = 3.65 MHz the component values are:

$$L1 = L2 = \frac{X_L}{2\pi f} = \frac{50}{2\pi \times 3.65} = 2.18\ \mu H \qquad \text{(Eq 11-10)}$$

$$C1 = C2 = \frac{10^6}{2\pi f X_C} = \frac{10^6}{2\pi \times 3.65 \times 100} = 436\ pF \qquad \text{(Eq 11-11)}$$

If we know the L and the C value of a hybrid, we can calculate its f_o:

$$f_o = \sqrt{\frac{10^6}{98\pi L C}} = \sqrt{\frac{12666}{L \times C}}$$

where
L is expressed in μH
C is in pF
F is in MHz

The coils should be wound on powdered-iron toroidal cores to provide the required tight coupling. The T-225A-2 cores from Amidon are a good choice for power levels well in excess of 2 kW. The larger the core, the higher the power-handling capability. Consult Table 6-2 in Chapter 6 for core data. The T-225A-2 core has a nominal A_L factor of 215 for μ = 10. The predicted number of turns is calculated as

$$N = 100\sqrt{\frac{2.18}{215}} = 10.0 \text{ turns}$$

You can also use the "Mini Ring Core Calculator" by DL5SWB (**www.dl5swb.de**), which is a very handy tool for doing calculations concerning toroidal cores. We must realize that the A_L value specified by the manufacturer is a typical value. The permeability of these cores can vary quite significantly among production lots or from one manufacturer to another. Note that moving the windings on the core can help you considerably to fine-tune the inductance of the coil. By spreading the turns from over half the core to over the entire core, I could adjust the inductance on a T-225A-2 core as follows: 7 turns: 1.75 to 2.1 μH, 8 turns: 2 to 2.3 μH and 9 turns: 2.3 to 2.6 μH. Note that in this case the coil requires 2 turns less than the formula predicted! You should always measure the inductance at the frequency of operation, using an AIM 4170 or similar quality impedance meter.

For high power applications it is advisable to wind the hybrid coupler coils with #14 AWG or #16 AWG multi-strand Teflon-covered wire. The two coils must be wound with the turns of both coils adjacent to one another (keep them nicely parallel with a strip of Scotch 27 glass cloth tape), or the two wires of the two coils can be twisted together at a rate of approximately 2 turns per cm before winding them (equally spaced) onto the core.

When selecting the capacitors to be used in the coupler, take into account the stray capacitance between the wires of the inductors L1 and L2. The correct procedure is to first wind the tightly coupled coils L1 and L2 and adjust the spacing between the pairs of turns to obtain the desired inductance. Once this is done, fix the windings so they can no longer move on the core (eg with glass cloth tape) and then measure the inter-winding capacitance and deduct half of that value from the theoretical value of C1 and C2 to determine the required capacitor value. Assume the measured stray capacitance value is 34 pF. You should always subtract half of the measured inter-winding capacitance from the calculated capacitor value: C1 and C2 must each be 436 – 34/2 = 419 pF.

A final check of the hybrid coupler can be made with a vector voltmeter or a VNA (vector network analyzer). By terminating ports 2, 3 and 4 with 50-Ω resistors, you can now check for an exact 90° phase shift between ports 2 and 3. The output voltage amplitudes should be equal at the hybrid design frequency. If the hybrid was designed for a frequency higher than the array design frequency, the voltage at the –90° port should be lower that at the 0° port (see above).

3.4.6.2. Testing the Hybrid

Rob, W1MK, who literally built dozens of 90° hybrid couplers says that the best way to evaluate these circuits is to make the following two measurements:

3.4.6.2.1. f_o Measurement

To find out the f_o (design frequency, in other words the frequency where the circuit acts as a 3 dB hybrid), simply open ports 2 and 3 while having port 4 terminated in the hybrid's design impedance. Then look at port 1 (input port) for the frequency at which the input port SWR is lowest (see **Fig 11-35A**). It is not necessary to terminate port 4 in a resistor equal to the Z_0-hybrid. For example if the coupler were designed for a Z_0 of 85 Ω, and the load resistor was 50 Ω, the input impedance with ports 2 and 3 open will be equal to the load resistor impedance (50 Ω) at f_o.

3.4.6.2.2. Z_0-hybrid Measurement

To test for Z_0, terminate port 3 in a variable load and adjust the resistance of this load carefully to find the value where the power dumped into the load resistor at port 4 is lowest (see Fig 11-35B). The RF source can be an MFJ antenna analyzer (see Section 3.5.2.1) or other similar unit. A sensitive power meter such as described in Section 3.6.1.1 can be used at port 4, or you can even use a receiver (be careful to not overdrive the receiver). The characteristic impedance of the coupler will be:

$$Z_0 = \sqrt{50 \times \mathrm{R3}} \qquad \text{(Eq 11-12)}$$

While doing this test, one should be able to achieve approximately 30 dB suppression at port 4 compared to the input at port 1.

3.4.6.3. The 180° 1:1 Transformer

In the quadrature-fed Four Square, the design calls for a 180° phase shift between the front element current and the back element current. In the original Four Square designs one usually obtained this phase shift by inserting a 180°-long line in the element with the lagging feed current (the front element). In later designs the half-wave transmission line was generally replaced by a phase inverter transformer (**Fig 11-36**).

3.4.6.3.1. Characteristics of a Half-Wave Transmission Line Transformer

One can obtain a 180° phase shift by using a coaxial feed line that is exactly λ/2 long. It is obvious that this exact phase shift only occurs at the frequency for which the cable is λ/2 long. If cut for 3.65 MHz, the phase shift is off 4° on 3.5 and 3.8 MHz. If you want to cover both ends of the band, it is best to cut the transmission line for 180° on 3.8 MHz and add a short piece of cable (approximately 2.2 meters of coax if VF = 0.66) when operating on the low end of the band. The loss depends on the quality of the cable. For RG-213 or RG-8, loss is approximately 0.4 dB if terminated in 50 Ω and 0.61 dB including attenuation due to SWR when terminated in a typical impedance of 47 – *j* 46 Ω. If you use ⅞-inch Hardline, the loss is less than 0.1 dB. And yes, 0.4 or 0.6 dB is *not* a very good value; we can obtain better than 0.05 dB with a properly designed transformer (see below).

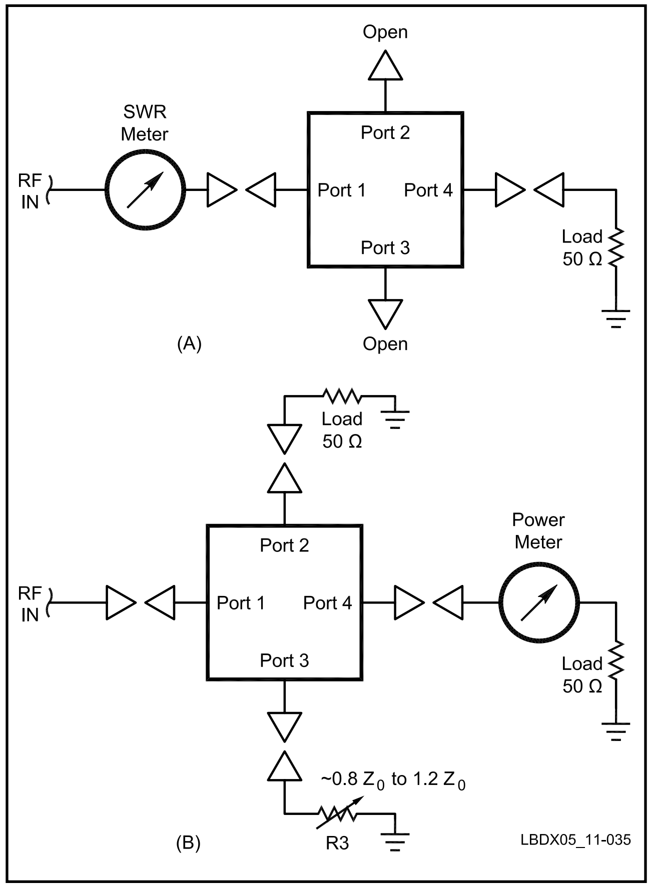


Fig 11-35 — A: Test setup for determining Z_0 of the 3-dB hybrid coupler. B: Test setup for determining the design frequency (Z_0) of the hybrid. Note that the impedance of the dummy load and the power meter connected at port 4 can have any impedance to do this test. In practice, you can use 50 or 75 Ω, it does not matter.

3.4.6.3.2. Do We Really Require Exactly 180°?

In other words: Is pure quadrature (that means "in 90° increments") with equal feed current magnitude on all elements (k = 1) the best feed solution? We know the answer is *no* (see Section 4.7). But many of us think that the quadrature 1/1/1/1 feed current is inevitable if we use a 3 dB hybrid coupler.

What if the transformer achieves a phase shift of –190° instead of –180°? What if the magnitudes of the feed currents in the elements are not all identical? It does not take long to find out the answer to that question, using *EZNEC* (and file *Ch11-4sq-0-90-180-1-1-1-1.ez*). Run this file (for an 80-meter Four Square) — change the phase angle of the feed current for the front element from –170° to –190°. You will see that the

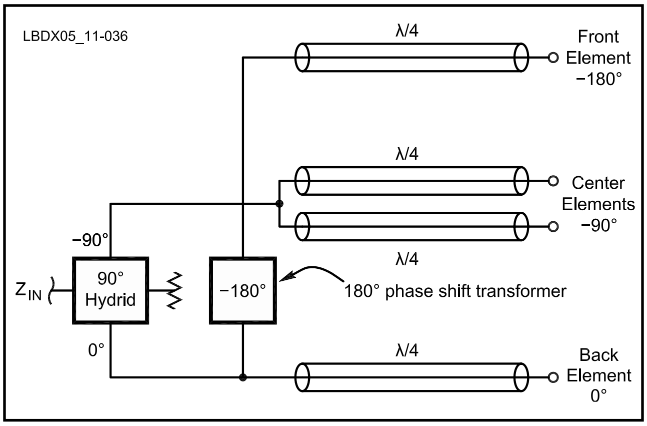


Fig 11-36 — Use of a 180° 1:1 transformer in the feed system for a quadrature-fed Four Square array.

best results (gain, F/B, RDF — see Chapter 7, Section 1.9) are obtained with a phase angle of approximately –188°, and *not* exactly 180°. This is good news, because the transformer we developed exhibits a phase shift that is *larger* than 180°, and this can be used to our advantage (see also Section 3.4.6.4.8).

We can further deviate from the "quadrature, 1/1/1/1" concept and, in addition to using –188° phase shift to the front element, we can also reduce the feed current magnitude to the center elements by approximately 15 to 20%. That further improves directivity (see Sections 3.4.6.5.10 and 3.4.6.6.7).

3.4.6.3.3. Principles of a Phase-Inversion Transformer

Compared to the half-wave transmission line creating a –180° phase shift, the advantage of the ideal 180° 1:1 transformer is that it is a wide-band component. It is expected to produce the required 180° phase shift over a wide frequency range and a perfect 1:1 voltage magnitude transformation ratio at all those frequencies. In a modeling program like *EZNEC,* one can simply specify "1:1, phase inversion" and this magic component does exactly that at any frequency, without any unwanted or uncontrollable side effects. Real transformers, however, are slightly different from "ideal" transformers. They introduce power loss, extra phase shift, and impedance transformation different from 1:1.

A phase reversal transformer is typically made in the shape of a transmission line transformer. In this type of transformer we wind a number of turns of a transmission line on a core made of high permeability material. Transmission lines used on transmission line transformers usually come in one of three shapes: parallel-wire lines, twisted-wire lines and coaxial lines. In all three cases the coupling between input and output is not done via induction, as in a regular transformer with separate primary and secondary. Rather it is achieved by the fact that the transmission line itself simply continues through the "transformer" (there is no galvanic isolation), while the two conductors of the transmission lines are flipped at one point — taking care of the phase reversal. There are no losses through imperfect coupling as the coupling is galvanic and thus 100% efficient. In other words, the core is not there to improve coupling as in most other types of transformers, but to realize a high inductive reactance Z_L (see also Chapter 7, Section 2.7.2.1).

Flipping the leads (as is commonly done with open-wire symmetric transmission lines) inverts the phase (180° phase shift), as, in each transmission line, the signals on the two conductors are always 180° out of phase. However, with *unbalanced* transmission lines there is a problem. If we would simply do that, we would short circuit the transmission line at the point where we flip the conductors, as shown in **Fig 11-37D**.

In Fig 11-37A we see a parallel transmission line (two wires, running side by side). To solve the "short circuit problem" we wind the transmission line on a high permeability core, which now introduces an inductance between A and A' as well as B and B'. While A is connected to A' and hence to ground (and B' to B, and thus also to ground), this no longer causes a short between A and B (and A' and B') because of the inductive reactance of the coil. Both the input and output are connected to ground through a high reactance coil formed by the transmission line conductors wound on the high permeability core. If we make these coils to have a high reactance ($ZL \times 10 \times Z_{COAX}$) and a very low loss resistance ($Q >> 100$), we can obtain a transformer that approaches the definition of an ideal transformer, at least as far as power losses due to the losses in the shunt coil are concerned. Fig 11-37B shows the common way of representing such a transformer.

The same can be done with a twisted pair line or with coaxial cable. The advantage of the coaxial cable is that there is no impedance discontinuity between the impedance of the parallel transmission line wound on the core and the impedance of the coaxial cables connected to the transformer. In addition,

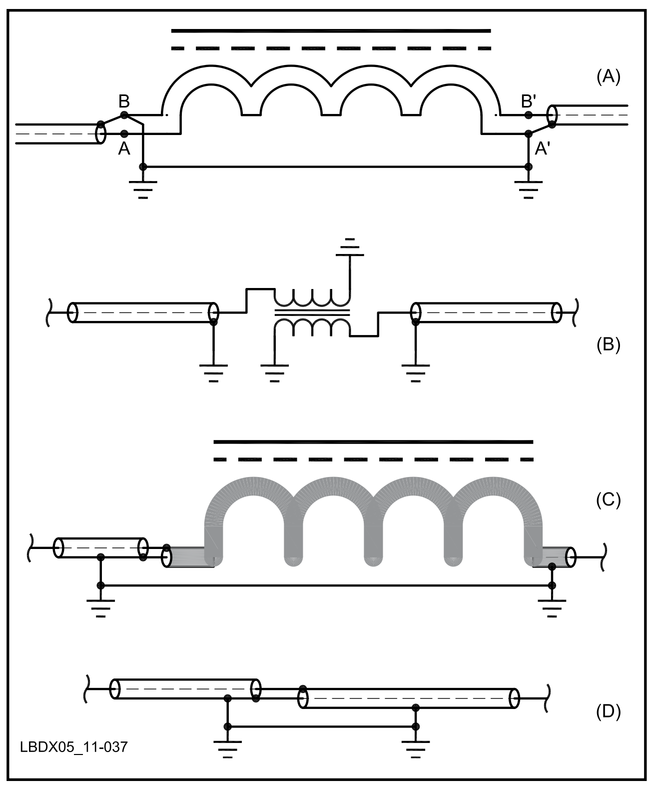


Fig 11-37 — If we merely flipped the conductors of the coax to obtain a 180° phase shift, we would also short circuit the coax (case D). Therefore we need to insert a high impedance on one side of the point where we flip the connections (A) and (C). B shows the common schematic representing 1:1 phase inversion transformer.

Fig 11-38 — The 180° flipped coax phase reversal transformer, after a design by K9DX. This unit for 80 meters uses 8 turns of RG-303 on a stack of two FT-250-61 (61 mix) cores.

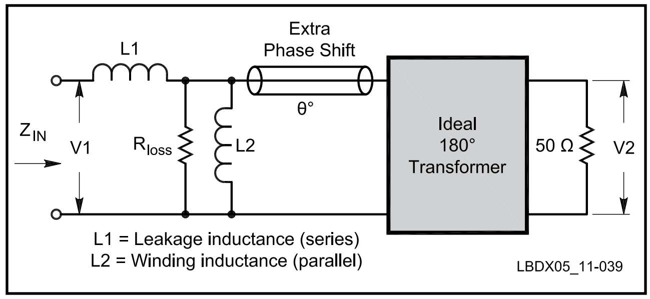


Fig 11-39 — Simplified equivalent circuit of the 180° phase shift transformer using transmission line transformer principles.

as a transmission line the coaxial cable is not influenced by the presence of the core. Seen as a fat conductor from the outside, the outside of the shield acts as conductor making up a coil wound on the toroidal core. With small Teflon insulated cable such as RG-303, the losses can be kept very small. In addition, the transmission line on the core is unbalanced as are the feed lines on both sides of the transformer. A practical example of a transformer wound with RG-303 coax is shown in **Fig 11-38**.

3.4.6.3.4 Performance Parameters

Fig 11-39 shows a simplified equivalent circuit of our 180° phase shift transformer. In addition to the ideal transformer (performs exactly 180° voltage shift at zero loss and zero voltage magnitude transformation) the real transformer has four important parameters to consider:

1) The equivalent parallel loss resistance (caused by losses in the core and in the coax used to wind the transformer).
2) The extra phase shift caused in the length of coax used to wind the transformer.
3) The parallel winding inductance L2 (the inductance of the coil made up by the coax wound on the core).
4) The series leakage inductance L1 (mainly caused by less-than-perfect coupling in the transformer).

The leakage inductance L1 is the main reason that the magnitude of V2 is smaller than the magnitude of V1. The voltage-factor is the ratio of the output voltage magnitude (V2) divided by the input voltage magnitude.

As part of a hybrid coupler feed system for a Four Square array, the important parameters are dictated by the current-forcing feed system used in the array. The "Basics of Current Forcing" say that shunt impedances at the point where the 180° transformer is located have no effect on the current in the antenna (as these change neither voltage magnitude nor angle).

- We want to have full control over the *current* at the array elements.
- This means we need to have full control of the *voltage* at the end of the λ/4 wave feed line
- We should understand that a series impedance element at the antenna element does *not* influence the feed current, a shunt element does.
- Translated to the other end of the λ/4 wave feed line, this means that any shunt element in that place does *not* influence the feed current in the antenna element.

This means that the winding inductance has no influence on the element feed current. Try this out using the modules "Shunt Impedance Network" and "Series Impedance Network," both part of the *New Low Band Software* available on the CD. It is also on this principle that the phase compensation optimization method, developed by Robye, W1MK, is based (see Section 3.4.6.6). Comtek and DX Engineering use this shunt inductance to an advantage (see Section 3.4.6.5.7).

Power Loss

By definition,

$$\text{Power Loss} = 10 \log \frac{\text{Power out}}{\text{Power in}}$$

However, if we measure the insertion loss using a VNA such as the Array Solutions VNA 2180 or the N2PK VNA (see Section 3.5.2.4), we measure S21 which is the insertion loss. This is the ratio of the voltages measured at the output of the device under test to the voltage at the input of the device under test. With these two VNAs the output load impedance is 50 Ω, and the signal generator internal impedance is 50 Ω as well.

If Z_{in} = 50 Ω and $Z_{in\text{-}real}$ = 50 Ω, power loss is equal to insertion loss. In other words, only when we have a perfect match when measuring the insertion loss will the power loss equal insertion loss.

If we want to know the power loss, we first need to measure the input impedance of the device under test when terminated in a 50-Ω load impedance. In real life this input impedance will be different from the load impedance (50 Ω) because of R_{loss}, L-shunt and L-leakage (see Fig 11-39).

Next we need to measure the insertion loss (measuring S21), and use both values (Z_{in} and insertion loss) to calculate the power loss. The formula to do this is:

$$\text{Power Loss} = \text{IL} + 10 \log \frac{(50 + Z_{in-real})^2 + Z_{in-imag}{}^2}{4 \times 50 \times Z_{in-real}} \quad \text{(Eq 11-13)}$$

where

IL = insertion loss, the loss (S21) measured on the VNA

$Z_{in\text{-}real}$ = the real part of Z_{in} with the device terminated in 50 Ω

$Z_{in\text{-}imag}$ = the imaginary part of Z_{in} with the device terminated in 50 Ω

The CD that comes with this book has a spreadsheet program that can help with these calculations (*w1mk-trf-powerloss-phaseshift.xls*). Note that the insertion loss can be quite different from the power loss! As an example, if you measure the 180° transformer from a Comtek hybrid coupler system with a VNA, you will find an insertion loss of approximately 0.5 to 0.6 dB. The power loss however is typically less than 0.03 dB.

In a well designed phase inverter transformer, power loss in the transformer is caused by two mechanisms:

- The loss caused by the transmission line used in the transformer.
- The coil loss, caused by resistive losses of the core material. In other words, we need the highest possible Q to keep these losses low. If the Q of the coils is infinite, the transformer power loss would only be determined by the loss of the transmission line used in the transformer.

Phase Shift

Flipping the conductors of our transmission line causes a voltage phase shift of exactly 180°. But there is an additional phase shift caused by the length of the transmission line used to make up the transformer! If the SWR on the transmission line is 1:1, this phase shift will be exactly the same as the length of the transmission line (both expressed in degrees), of course taking into account the velocity factor of the transmission line. This is the reason why the total phase shift of the transformer is always greater than 180°, usually between –183° and approximately –193°.

The insertion phase shift we measure with the VNA is not necessarily the voltage phase shift we are interested in. Remember, we are dealing with *voltage-forced* feed lines, in which the voltage phase shift is the image of the current phase shift at the antenna elements.

Here too, similar to what we need to do to calculate the power loss, we need to apply a two-step compensation to calculate the voltage phase shift. Starting from the insertion phase shift measured using a VNA, we can calculate the voltage phase shift using the spreadsheet *w1mk-trf-powerloss-phaseshift.xls*.

An easy way to directly measure the actual voltage phase shift is to resonate the coil on the desired operating frequencies (put a capacitor in parallel). The phase shift we now measure on our VNA is the voltage phase shift we are interested in. This value will always be –180° minus the phase delay caused in the transmission line to wind the transformer.

Leakage Inductance

In our transformer equivalent circuit shown in Fig 11-39, the *leakage inductance* is represented as an inductance in series with the transformer primary winding. This leakage inductance is generally due to imperfect coupling of the windings and creation of leakage flux that does not link with all the turns of the winding. It has been stated in the literature that transformers using ferrite core material (such as 61 mix) exhibit significantly lower leakage inductance that the powdered iron type materials (such as type 2 material). This was confirmed by some measurements indicating that the 180° phase reversal transformer wound on a T-225A-2 core exhibited a leakage inductance of 0.3 µH while a similar transformer wound on a stack of two FT-240-61 cores yielded only 0.07 µH. This leakage inductance appears to be the main reason for the difference in voltage-factor (a-factor) between transformers made with these two different materials. This voltage factor for the transformers wound on the 61 mix ferrite cores is approximately 0.999, while for the powdered iron (type 2 materials) this factor is significantly lower (0.922). Therefore the insertion loss for the transformer wound on the powdered iron core is much higher than for the transformer wound on 61 mix ferrite material, although the power loss is very similar in both cases.

This also means that the transformer wound on the powdered iron core, which has a much lower parallel inductance value, creates more impedance transformation than is the case with the transformer using the higher mu type 61 ferrite core. But the impedance transformation by itself is not an issue; voltage (magnitude and phase) are what we are concerned about. If we use such transformers at the end of one of a current-forcing feed line to an array element, we would like the voltage at the end of the current-forcing feed line *not* to be changed by the transformer in order to preserve the correct current ratio at the different array elements.

The non-ideal voltage factor of the transformer is, of course, but one of several parameters that will influence the element feed current. Another factor is the attenuation in the current-forcing feed lines, which will not be the same in the different feed lines because the SWR on the feed lines can be very different. This is especially so for the feed line going to the reflector element, where the SWR is usually very high. This is why we should use very good coax to make our current-forcing feed lines (preferably ½-inch Hardline).

This all results in the fact that the element feed currents in an array using current-forcing feed lines and a phase-inversion transformer are generally not exactly what we anticipated. Further analysis of the magnitude of the impact of the variations has however shown that the influence on the directivity and gain are rather limited (see also Section 3.4.6.4.8).

Overall the above analysis points out that, if you really want your Four Square to operate under "textbook" circumstances, you will need to check the feed currents using a method as described in Section 3.4.9.5 and shown in Fig 11-89 later in this chapter. Checking the feed voltage at the end of what should be a current-forcing feed line is not 100% accurate. The loss in the feed lines and voltage-factor (a-factor) of the phase-inversion transformer will inevitably change the voltage somewhat, which will show up as a somewhat changed feed current at the antenna. The most direct measurement method, which is measuring the feed currents using a small current transformer, is the preferred measurement method to obtain the exact feed current values.

All of this, however, does not mean that hybrid-coupler-fed Four Square does not work well "as is." It is a forgiving design that can be categorized under plug-and-play systems, unless of course you are a perfectionist. And, if you have read this far, you must be!

Note also that having the feed currents slightly different from the quadrature conditions can achieve both better directivity and (slightly) increased gain. But to achieve these results, changes should be done in a well controlled way.

3.4.6.3.5. Practical Design

Our goal was to design a phase-inversion transformer exhibiting

- high parallel inductance
- low power loss,
- a voltage transformation ratio (a) closer to 1
- a total phase shift close to 180°

We all know that the commercial hybrid couplers do not always meet these requirements. Does it hurt? We will find out.

The new transformer we designed uses a stack of 2 cores made of high-Q ferrite material (61 mix). The Q factor of coils wound on type 61 ferrite cores range from much greater than 500 on 1.8 MHz, to around 200 on 3.5 MHz, to 75-100 on 7 MHz.

The design was guided by the following principles:

- To keep the total phase shift as close as possible to 180° we needed a short transmission line on the core.
- To achieve a relatively high inductance, we required a core material with a higher permeability than is the case for iron powder type 2 material. The 61 mix ferrite material was our choice.
- To make sure we would never overheat the core when running high power (ferrite cores can more easily be permanently damaged than powdered iron cores) we decided on using a stack of two 2.4 inch cores (FT-240-61).
- We also chose to use Teflon insulated coaxial cable (RG-303) to wind the transformer, as the coupling is 100% minus, of course, the loss due to attenuation in the short RG-303 transmission line used to wind the transformer.

The development turned out to be an interesting but time-consuming exercise. I must thank Robye, W1MK, Roger, ON5WU, and Greg, W8WWV, for their guidance, help and encouragement. Roger and I built dozens of transformers on a stack of two FT-240-61 Amidon cores (µi = 125 and A_L = 171), searching for the best solutions for those ideal designs (see Fig 11-39). After thorough testing we ended up with three optimized designs, one for each of the low bands (see **Table 11-10**).

The Smoke Test

For the three transformers shown in Table 11-10, we calculated power losses between 0.022 and 0.029 dB (0.5% to 0.67% power loss). These transformers were tested by running 4 kW CW power through them into a dummy load. After 10 minutes, the transformer temperature climbed from 25 °C ambient to less than 35-40 °C (about the same as body temperature), which is very acceptable. As a matter of fact the heating seemed to come more from the RG-303 cable than the transformer ferrite core.

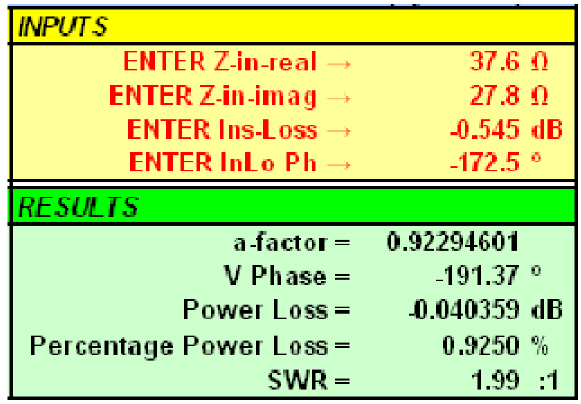


Fig 11-40 — Performance data for the 180° phase shift transformer from a Comtek hybrid coupler for 80 meters. This is a screen shot from the *W1MK-TRF-powerloss-phaseshift.xls* spreadsheet program used to calculate the phase angle and the power attenuation.

When dealing with heat dissipation you should take into account the average power over a relatively long period (depending on the mass and the area of the body being warmed up). On CW our duty cycle is 50%, and even when very busy in a contest we transmit only about half of the time. Also, don't forget that only about half of the power going into a Four Square passes through the transformer. This means that for the 80-meter design (0.53% power loss), 5.3 W are reduced to approximately 1.25 W, which is a continuous amount of power that easily can be handled by a stack of two 2.4 inch cores.

3.4.6.3.6. The Transformer in the Comtek and DX Engineering Controllers

The two most popular hybrid controllers on the market both use a 180° transformer where the inductance value of the coil is relatively low. With 10 turns on a powdered iron toroidal core these transformers achieve a shunt inductance of approximately 4 µH. Using the same number of turns on a ferrite core, the transformer we developed yields an inductance of 42 µH on 80 meters. The net result is that the transformer we developed requires fewer turns, and consequently its measured phase shift is closer to 180° than is the case with the commercial units. (The "extra" phase shift was reduced by half.)

Fig 11-40 shows the results of the analysis of the phase reversal transformer taken from an 80-meter Comtek system. The input impedance (given a 50-Ω load), the insertion loss and the voltage phase shift were measured on a VNA. (In Fig 11-104 later in this chapter we see the VNA 2180 set up to do such a measurement.) The spreadsheet calculator *w1mk-trf-powerloss-phaseshift.xls* was used to calculate the voltage phase shift and power loss.

Note that the main difference between this result and the results of the transformer we developed (data

Table 11-10
Phase Inverter Transformer Performance
Measured performance data for the final designs of phase inverter transformers as described in Section 3.4.6.3 The phase shift includes the shift created by two pigtails (in and out) of RG-303 cable, each 5 cm long.

Band	*Turns*	*Phase Shift (θ)*	*PWR Loss (dB)*	*PWR Loss (%)*	*Voltage ratio (a-factor)*	*Parallel inductance*
160	13	–184.8°	–0.026 dB	0.59%	0.993	~59 µH
80	8	–186.1°	–0.024 dB	0.55%	0.995	~27 µH
40	5	–188.8°	–0.029 dB	0.67%	0.990	~10.6 µH

shown in Table 11-10) is the a-factor, which is the ratio of the output voltage magnitude to the input voltage magnitude (Comtek: a = 0.922, the design described above: a = 0.995). The a-factor is important because the voltage at this point is the "image" of the current at the end of the current-forcing feed line.

We calculated the power loss of the transformers used by Comtek and DX Engineering to just under 1%. Although this loss is about 50 to 80% higher than what we reach with our new design, the difference is irrelevant. We know that the T-225A-2 core used in the Comtek design can take much more than 2000 W without ill consequences, and the stack of six T-157-2 cores in the shape of a binocular core (see Fig 11-99 later in this chapter) as used in the DX Engineering units will certainly perform at least as well when considering power loss and heating.

The question now is, "Is this shunt reactance causing problems?" The answer is, "No. As a matter of fact, if the shunt reactance weren't there, the hybrid coupler would perform very poorly." The shunt reactance is an essential part of the circuit. Strangely enough, I have never seen an article explaining that it *is* an essential part, nor anyone explaining why. I will explain, and that's part of the "demystification" process.

We should not forget though that the newly designed transformer yields a much higher a-ratio (voltage magnitude transformation ratio), which means that we will be supplying a more correct current magnitude to the front element than is the case with the other transformers that are currently used in commercial units.

If you analyze an "ideal" hybrid coupler, using no such "shunt reactance" across Z3, it performs very poorly. Read on, you'll find our how and why.

3.4.6.4. The Quadrature Hybrid Coupler Feed System Demystified

The hybrid has two ports to deliver power to our array: port 2 and port 3. Under ideal conditions ports 2 and 3 deliver voltages with the same magnitude (if k has been specified as 1), but 90° out of phase. Port 2 is the lagging port, the –90° port, while port 3 is the reference port, the 0° port.

Our Four Square has four elements, with four feed lines. Let us think about interfacing the four feed lines with the two ports.

3.4.6.4.1. W1MK's Four Square black box

Robye, W1MK, came up with the concept of the Four Square black box, shown in **Fig 11-41**, where he includes the Four Square radiating elements as well as the current-forcing feed lines and the 180° transformer. The black box has only two terminals, called the port 2 terminal and the port 3 terminal (equivalent to the port 2 and port 3 denominations for the hybrid network).

Z2 is the parallel impedance value at the end of the current-forcing feed lines going to the center elements. The Z3 branch incorporates the phase-inversion transformer in the leg going to the front element.

We can conceive a black box such as shown in Fig 11-41 for every quadrature-fed array (0°, –90° and –180°) where the array, the feed lines, transformers, phasing lines, etc, can be included inside the black box. The black box has only two feed ports: the port 3 (0° port) and the port 2 (–90° port) feed points.

We can distinguish between two families of such arrays:

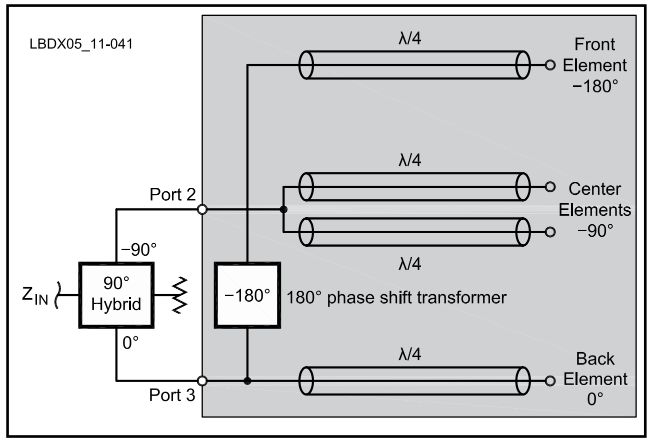


Fig 11-41 — The diagram of a hybrid coupler fed Four Square array can be separated into two blocks, having just two ports connecting these blocks. In the text we will analyze these two blocks separately. The Four Square block box contains the radiating elements, the current-forcing feed lines and the 180° phase shift transformer. The hybrid box contains the hybrid components.

The first family is arrays that are *fully symmetrical* physically as well as electrically. In such arrays the impedance at port 2 is independent of the load presented to port 3 and vice-versa. Robye, W1MK, found out that in an "ideal" quadrature fed Four Square array there is no mutual coupling between ports 2 and 3. That does not mean that there is no coupling between all the elements in the array, but only that certain couplings cancel out one another because of the physical and electrical symmetry of the array.

Examples are:

- The symmetrical quadrature-fed Four Square array (Fig 11-41).
- A 3-element symmetrical quadrature-fed end-fire array where Z11 = Z22 = Z33, and Z13 = Z31 and Z12 = Z23 = Z32 = Z21 (see Section 3.4.6.10.2).
- A fully symmetrical and quadrature-fed 6-element array (HEX array), see Section 3.4.6.10.3.

The other family is *non-symmetrical arrays*, such as the 2-element end-fire array, triangular arrays, rectangular arrays, etc.

3.4.6.4.2. Our EZNEC Four Square Model

Throughout the remainder of this chapter we will often refer to *EZNEC* antenna models, all of which are made available on this book's CD. The "standard" model that we use in this and following sections is based on:

- F_{design} = 3.65 MHz
- Element diameter: 60 mm
- Element height: 19.7 meters (which makes an individual element resonant on 3.65 MHz)
- Footprint: 20.54 by 20.54 meters (quarter wave on 3.65 MHz)
- Equivalent ground loss resistance: 5 Ω
- Ground conductivity: 5 mS/m, ε = 13
- Current-forcing feed line impedance: 75 Ω
- No feed line losses (this can of course be changed in the model, if required)
- Quadrature feed conditions: phase difference in 90° steps

- k = 1 (unless otherwise indicated) which means that all elements are fed with the same feed current magnitude
- A lossless phase reversal transformer giving exactly 180° phase shift

We use the symbol "k" as the ratio between the feed current in the center elements of the Four Square versus the feed current magnitude to the front and back element.

The file *Ch11-4sq-quadfed-Z2-Z3legs.ez* is the *EZNEC* model using W1MK's black box approach, which directly calculates the values of Z2 and Z3. The values obtained via this model (using lossless 75-Ω quarter wave feed lines) are: Z2 = 53.7 + *j* 22.5 Ω and Z3 = 60.7 – *j* 36.0 Ω.

It is most important to note that, in case of perfectly symmetrical arrays, we not only can model the array in a simple way and obtain Z2 and Z3 without extra calculations, but that in real life we can also directly measure Z2 and Z3 independently and that it is totally irrelevant which load is connected to the second branch that you are not measuring. Be careful however; if the array is not perfectly symmetrical, the impedances we measure at ports 2 and 3 are *not* Z2 and Z3. Due to mutual coupling, a third impedance is involved — the mutual impedance Z23 (see Sections 3.4.6.4.4 and 3.4.6.4.5).

Probably the major difference between modeling and measuring is that you model with an ideal transformer, while in a measuring situation you use a real transformer that has a shunt reactance, and due to the length of the transmission line on the transformer core, established a shift of 180° (+5 to 10°), which is also responsible for an additional impedance transformation. This means that even in a system where all array elements are perfectly symmetrical (electrically and physically), Z3 and Z3 would not be without some mutual coupling due to the less-than-ideal phase reversal transformer.

To avoid these problems, we can use a high quality half-wave long transmission line as a transformer. I have a piece of ⅞-inch Hardline for that purpose, resonant on 3.775 MHz (approximately 32.5 meters). By adding a 2.5 meter length of the same type of coax the line is resonant on 3.51 MHz. Such a line forms the most ideal transformer that we can make. We can use it to make perfect measurements on both ends of the band. The line loss is <0.1 dB, which is comparable to the loss of a well-made transformer (see Section 3.4.6.3.5). But even then, the perfect electrical balance will be lost because of the additional loss in the ½ λ line. Also consider the different losses in the different current-forcing feed lines due to different SWR. Losses in the feed line to the front element will generally be significantly higher because of the very high SWR on that line.

3.4.6.4.3. Checking for Symmetry

If you want to check the symmetry of your array in the real world, you should use a quality half-wave transformer as described above (it's the closest we can come to an ideal transformer). Almost as good is a transformer such as developed in Section 3.4.6.3.5, where you eliminate the shunt resistance by resonating the transformer with a parallel capacitor. Although our ideal model is perfectly symmetrical, in reality some imperfections (physical, the "imperfect transformer," the different losses in the feed lines because of different SWRs) will make the best Four Squares less than 100% symmetrical.

On a model, one way to check if there is no mutual coupling between the Z2 and Z3 legs is first to run the model with both ports 2 and 3 excited (voltage magnitude and phase). Note the impedances at this point (read them from Scr Dat). Next specify a source voltage for the Z3 leg only, and run the program again. Check the impedance Z3 (on the Scr Dat window). It the value remains the same as with both legs driven, it means that there is no mutual coupling between the legs.

If you want to check it on a real antenna, the easiest way is to measure Z2 and alternatively open/short the Z3 feed line. If there is no change in Z2 impedance, it means there is no coupling between the ports. There is of course mutual coupling between all the elements, but because of the symmetry in the design, these couplings cancel one another as explained earlier.

A much more sensitive way to assess the symmetry of your array is to use a VNA (vector network analyzer, see Sections 3.5.2.3 and 3.5.2.4) and excite port Z2 while observing the output from port Z3. If the black box were fully symmetrical, the output would be zero. If we would measure the impedance, it would be 0 Ω. Then reverse the situation: inject in Z3 and measure the level in Z2 (see **Fig 11-42**). If there is no coupling at all, the rejection should be infinite. We call this test the diagonal isolation test.

Watch out: Is the array all by itself at your station? Is there no other tower, antenna, flag pole, metallic downspout or other metallic structure within, say, 1 wavelength? At few antenna locations will this be the case. That is also the reason why in very few Amateur Radio stations will we measure an isolation that is >40 dB (which by itself is a very high figure). Any metallic structure will couple to a certain degree with your array. In such a case, even though the array itself may be fully symmetrical (both mechanically as well as electrically) a "third player" may upset the situation. The VNA measurement is a perfect measurement to help you assess the degree of isolation between the two diagonal legs of your array.

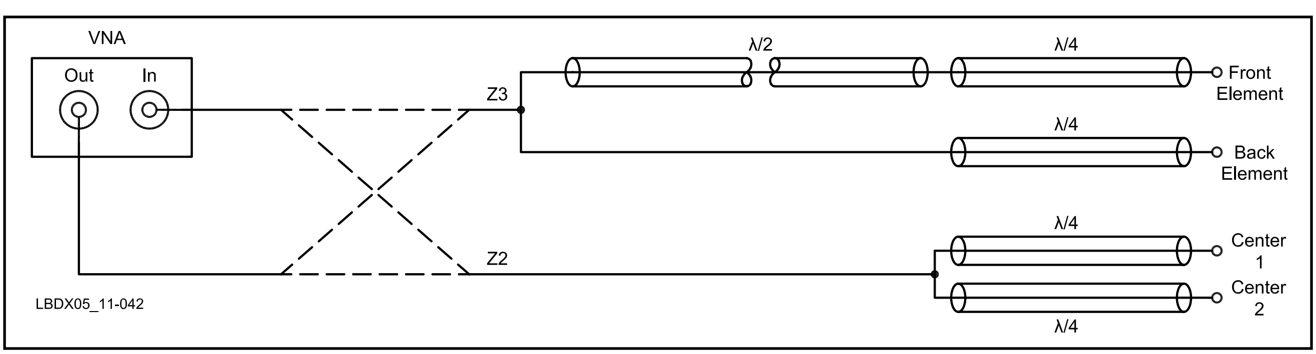


Fig 11-42 — Diagonal isolation measurement setup.

Greg, W8WWV, modeled a number of error situations, such as different radiator lengths, differences in length of the current-forcing feed lines and errors in element positioning. We can of course also have a combination of all of those. Greg calculated the diagonal isolation for a large number of isolated as well as cumulative cases. It appears that the asymmetry in the setup of the array causes the (theoretically) infinite isolation in case of perfect symmetry to drop to the 30 dB ballpark. Isolation would be lower if you misplaced one element of the array 1 meter or more (on 3.65 MHz), which is rather unlikely.

Another, much more disturbing factor is mutual coupling by conductors that are not part of the array. It appears that such factors are much more likely to upset the functioning of the array than slight asymmetry in the array itself. A resonant element at nearly ½ wavelength from the center of the array, brings the diagonal isolation down from 60 dB or better to a mere 13 dB!

When the same element is lifted from ground (now becoming resonant at approximately twice the original frequency), the isolation improved to 37 dB, which is OK, but still not as good as in a virgin case with nothing around (63 dB).

Greg concluded that despite the fact that "reasonable errors" in the symmetry of the array proper would still offer 30 to 40 dB or even better diagonal isolation, any metal, resonant or not and present within less than 1 wavelength, is likely to bring down the isolation to less than 30 dB

- 50+ dB isolation: very good
- 40+ dB isolation: good
- 30+ dB isolation: OK
- 20+ dB isolation: you could do better
- <20 dB: did something fall over? Or, it's time to detune that parasitic element!

To get the full picture, both diagonals need to be measured. This full picture may also give helpful information on where the troublemaker is located.

It is clear that this VNA measurement method can be used not only to assess the diagonal symmetry inside of your array, but also as a very sensitive tool for detuning a nearby tower or other metallic structure.

What you measure with a VNA is the insertion loss (voltage loss). The power loss takes into account the input impedance and the load impedance. The *w1mk-trf-powerloss-phaseshift.xls* spreadsheet (**Fig 11-43**) can be used to convert from insertion loss to power loss. Note that with high values of loss, the difference between insertion loss and power loss is small.

3.4.6.4.4. Measuring the Impedances at the Black Box Ports 2 and 3

After having optimized the diagonal isolation of your array (to at least 30 dB, if possible), it is time to measure the impedances at the two black box ports, port 2 and port 3. If the diagonal isolation is high enough, what we measure will be very close to the impedances Z2 and Z3 obtained by modeling (within a few ohms).

First of all, if we were using a λ/2 line as a 180° transformer while doing the diagonal isolation test, it is now time to remove it, and replace it with the real transformer (unless you want to use the line as a 180° transformer, which is perfectly possible, of course).

Using a quality impedance analyzer (such as the AIM 4170) measure the impedance values at the two black box ports. Assuming that you now are using a real transformer, these values will undoubtedly be different from what was modeled or measured with the λ/2 line in place, mainly because of the shunt reactance introduced by the transformer winding (unless that reactance was "tuned out" by resonating it on the operating frequency). What we are measuring however are *not* exactly Z2 or Z3, as there inevitably is mutual coupling under these non-ideal circumstances.

What we need are the correct values of Z2 and Z3 which we need as inputs to our optimization systems as described in Sections 3.4.6.5., 3.4.6.6 and 3.4.6.7. Don't worry, there's a way to turn the incorrect impedance values into fully correct values. A little math does wonders.

3.4.6.4.5. Calculating the Correct Z2 and Z3 Values in Case of Coupling

Fig 11-44 shows the equivalent circuit of the two-port black box, characterized by the impedances Z22, Z22 and the mutual impedance Z23. Coupling between elements of the black box causes mutual coupling. We have also learned in Section 3.3.1.3 that the degree of coupling is expressed by the mutual impedance.

In the ideal case (no coupling), Z23 will be equal to zero. In such case it is very simple to measure Z22 and Z33 independently, as the termination of the other port has no effect.

Robye, W1MK, developed the mathematics and a spreadsheet program to calculate the values of Z2 and Z3 based on the measured impedance values at ports 2 and 3. These are indeed different from Z2 and Z3 because of the mutual impedance Z23 and Z32.

Procedure: Measure Z2,3 at port 2 with the port 3 closed (shorted) or Z22 with the same port open. The same holds for port 3 where we measure Z3 or Z3,2. In principle, if the coupling arises from inside the array (not from an external element such as a tower 50 meters away from the array proper), Z2,3 should equal Z3,2. But it is best to measure both coupled impedances and use both in the spreadsheet to calculate Z23 and Z32. One can then calculate the average between Z32 and Z23 to calculate the value to be used for determining the shunt

	INPUTS	
2	ENTER Z-in-real →	58 Ω
3	ENTER Z-in-imag →	22 Ω
4	ENTER Ins-Loss →	-23 dB
5	ENTER InLo Ph →	0 °
	RESULTS	
6	a-factor =	0.06289337
7	V Phase =	-9.26 °
8	Power Loss =	-22.799532 dB
9	Percentage Power Loss =	99.4751 %
10	SWR =	1.54 :1

Fig 11-43 — Assume we measure an impedance of Z2 = 58 + *j* 22 Ω with port B terminated in 50 Ω (the input impedance of port B of the VNA). If we measure an insertion loss of 23 dB, this represents a power loss of 22.8 dB as calculated using the *W1MK-TRF-powerloss-phaseshift.xls* spreadsheet.

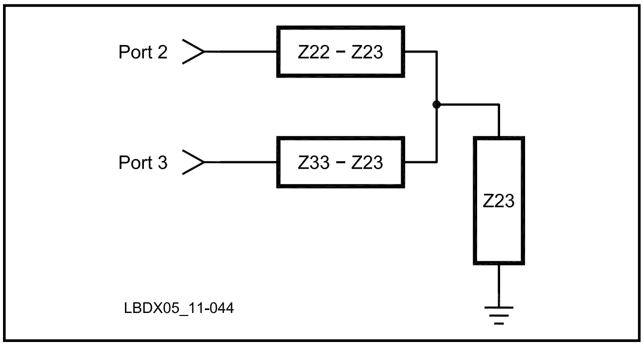


Fig 11-44 — Two port equivalent circuit of the black box shown in Fig 11-41. This equivalent circuit serves to calculate Z2 and Z2 from the measured values at port 2 and port 3 (Z22, Z33 and Z23).

BLACK BOX 2-PORT COUPLING (W1MK)			
A - INPUTS			
1	Enter k →	1	must be ≤1
2	Enter Angle θ →	-90	must be –
		real part	imag part
3	Enter Z22 →	93.93	44.26
4	Enter Z33 →	123.1	94.99
5	Enter Z2,3 →	91.05	42.04
6	Enter Zin parr Z2+Z3. →	43	21
B - CHOICE OF SIGN MUTUAL IMPEDANCE			
7	Z23 =	18.83	14.52
8	IF Zin parr =	63.58	38.01
OR			
9	Z23 =	-18.83	-14.52
10	IF Zin parr =	44.68	24.75
11	Enter Z23 →	-18.83	-14.52
C - OUTPUTS			
12	Z2 =	95.25	27.47
13	Z3 =	91.39	106.04
DIAGONAL ISOLATION			
14	Diagonal Isolation =	-21.0 dB	

Fig 11-45 — The two-port coupling spreadsheet program which allows us to calculate Z2 and Z3 from Z22, Z33, Z2,3 and Z3,2. The procedure for using the program is given in detail on the spreadsheet.

elements of the hybrid optimization system.

The inputs to the spreadsheet program are k (ratio of wanted voltage magnitude ratio of V2/V3), θ (phase shift between these voltages) and the Z22, Z33, Z23 and/or Z32 measurements. The program first calculates the mutual impedance (Z23 or Z32). As the mutual impedance is the result of a square root, we have two mathematical results, which are only different by the sign. The software has a built-in procedure that makes it possible to verify which of the two solutions is correct. Therefore one must make one extra measurement — the impedance of port 2 and port 3 in parallel. The procedure is described in detail in the spreadsheet software. Once you have selected the correct mutual impedance, the software will calculate — for the frequency where the measurements were made — what the impedances are at port 2 and port 3 when those ports are voltage fed with the specified k and θ.

The program consists of two sheets: the first sheet uses the measured values Z11, Z22 and Z2,3 while the second sheet uses the coupled impedance Z3,2. In a perfectly balanced array the results should be identical, which is seldom the case. Therefore the Z2 and Z3 impedances can be slightly different. One practical solution is to take the average of the two Z2 and the two Z3 values to define the value to be used for calculating the impedance of the shunt reactances.

The spreadsheet *two-port-coupling.xls* (see **Fig 11-45**) can be used with any array that can be reduced to a two-port black box, whether it is symmetrical (such as the Four Square) or not (such as a 2-element end-fire array). This tool can be used to look at the effects of unwanted coupling and also with systems where the coupling is intentional (as in the 2-element end-fire array).

3.4.6.4.6. Calculating the Diagonal Isolation Instead of Measuring

Section 3.4.6.4.3 introduced W8WWV's concept of *diagonal isolation* as a measure of symmetry in a Four Square array, and showed how this can be measured using a VNA (in S21 mode). The *two-port-coupling.xls* spreadsheet also calculates this isolation figure (see Fig 11-35 where in the example case the program calculates an isolation of 21 dB). This makes it possible for those hams who do not have a VNA to have access to the diagonal isolation figure. The disadvantage of this system, however, remains that you cannot constantly sweep a band of frequencies and see the isolation curve in real time on the screen of a laptop PC, as you might want when making adjustments. Also, be aware that the value you calculated is influenced by the characteristics of your 180° transformers, and the calculated value will only be correct for an ideal transformer. Such conditions can be approached by using a λ/2 long transmission line (high quality low loss transmission line) or if you resonate the parallel inductance of the transformer with a parallel capacitor.

3.4.6.4.7. Simulating the Measurements of Z22, Z33, Z2,3 and $Z_{parallel}$ with EZNEC

You can create these impedances using a model in *EZNEC*. Assume we want to calculate the values for a 2-element end-fire (1⁄4 λ spacing, quadrature fed, k = 1). Make a modeling file that includes the feed lines. Let's take 50-Ω feed lines. At the end of the current-forcing feed line to the back element we have port 2 and at the end of the current-forcing feed line to the front element we have port 3. *Ch11-2el-endfire-90-90-incl-Z2Z3.ez* is such a file.

- To calculate the self impedances Z11 we will feed (on the Source page) port 2 with 1 A and at port 3 with 0 A (and vice-versa for Z22). The Scr Dat window shows Z11 = Z22 = 50.2 – *j* 28.8 Ω.
- To calculate the coupled impedance Z2,3 we shall feed port 2 with 1 A and short circuit port 3. We can do this by feeding port 3 with a voltage of 0. Z2,3 = 62.7 – *j* 10.9 Ω.
- To calculate the parallel impedance of Z2 and Z3 we feed both ports with I = 1 and divide the impedance shown in the source window by 2: Z_{par} = 21.2 + *j* 2.9 Ω.

When we use these values in the *two-port-coupling.xls* spreadsheet, we calculate Z2 and Z3 as: Z2 = 37.9 – *j* 18.3 Ω and Z3 = 82.4 + *j* 42.6 Ω. These impedances are very close to what we obtained directly with the same *EZNEC* file when using as sources voltage sources with the values 1V∠0° and

1V∠–90°, which are Z2 = 37.9 – *j* 18.6 Ω and Z3 = 83.6 + *j* 42.8 Ω

This also proves that the *two-port-coupling.xls* spreadsheet program is correct.

3.4.6.4.8. Evaluating the Impact of a Non-Ideal Phase Reversal Transformer

So far we have either used a high-quality ½ wave line as a transformer (which is as close as we can get to the ideal transformer on one frequency) or, in our model, the ideal (nonexistent) transformer that yields a perfect lossless 180° phase shift over a wide range of frequencies.

To assess the influence of the extra phase shift caused by the length of the transmission line wound on the transformer core (an extra 4° to 12° of phase shift) we ran a simple modeling exercise where we lengthened the quarter-wave feed line in the arm that is followed by the transformer by 10° (physical length). This is not exactly 10° voltage phase shift but close enough to assess the influence. The impedance in the port 3 leg changed from 60.6 – *j* 36 to 56.3 – *j* 38, the gain changed approximately 0.2% and the F/B changed about 2%. We can conclude that we do not need to worry that these extra degrees will upset the directivity or the gain of the array to any significant degree. The change in impedance will, of course, cause a drop in diagonal isolation, and is one of the reasons the measured Z2 is not 100% identical to the calculated Z2.

In Section 3.4.6.3.6 we already said that the shunt reactance of the transformer is not a problem, as we will do a shunt-compensation (see Sections 3.4.6.5 and 3.4.6.6), and a shunt reactance does not change the voltage magnitude anyhow.

We also learned in Section 3.4.6.3.6 that an important difference between a commercial-type transformer and the transformers developed in Section 3.4.6.3.5 is the difference in voltage-factor (a-factor). As these transformers transform impedance (because of the line length), they also transform the voltage magnitude. The a-factor is the transformation ratio of the voltage magnitude. The transformers used in the commercial units yield an a-factor of approximately 0.925 while the newly developed 80-meter transformer (with 8 turns, see Table 11-10) yields a = 0.995. The 0.995 figure creates practically no change in gain and F/B, while a = 0.992 results in a change in F/B (reduction) of 10%.

Concluding, one can say that we should not be worried either by the extra degrees of phase shift (up to approximately 10°), nor by the parallel self-inductance (we need self inductance anyhow to optimize the system). It is clearly recommended to use a transformer with the highest a-factor, which is the newly developed transformer using the two FT-240-61 ferrite cores.

3.4.6.4.9. Operational Bandwidth of the Hybrid Coupler Feed System Used on a Four Square Array

To easily do a bandwidth assessment we must put all of the elements, including the hybrid coupler into one model, just like we did for the L-network matching system with the L-network(s) (see Section 3.4.5.9). Roy Lewallen, W7EL, has been very helpful in developing a hybrid that can be included in any antenna model. A separate document giving more details on this subject is available on the CD that comes with this book (file: *EZNEC model of the hybrid coupler.pdf*). We can now easily model an array while implementing the hybrid coupler as an integral part of the model.

The principles of this assessment are covered in Section 3.1.1. The procedure is similar to what's explained in Section 3.4.5.9. With 90° hybrid coupler feed systems there is an additional factor that determines the operational bandwidth: the port 4 dump power. We have set –10 dB (vs the input power at port 1) as a reasonable limit for this power. This means that, at worst — at the band limits — 10% of the power supplied to the antenna will be converted into heat in the dummy load on port 4. The dump power is not displayed as such in *EZNEC*. You need to copy the power from load 4 (the dummy load) as displayed on the load data screen, divide this value by the input power shown on the Scr Dat screen, and then express the ratio in dB. Example: load 4 power = 9.175 W, input power = 178.24 W, then the dumped power suppression equals 10 log (9.175/178.24) = –12.8 dB.

The procedure is very simple:

- Start up an *EZNEC* model including the hybrid coupler (*Ch11-4sq+hyb-3.65-k = 1-ideal transformer-no sh.ez*).
- Make sure that in the Loads window, on line 2 and line 3, in the columns L and C you put "open," which means that there is no extra shunt element connected across the

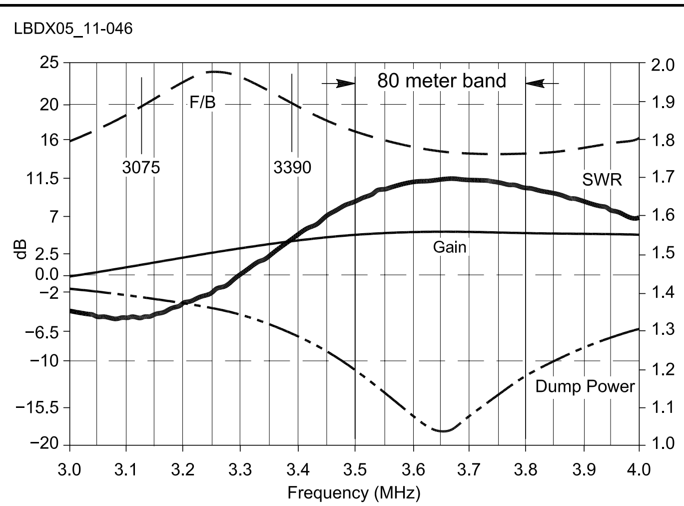


Fig 11-46 — Performance data of the quadrature-fed Four Square array fed by a "naked hybrid" with ideal components. Note the very mediocre F/B over the entire band. The F/B is calculated for a radiation angle of 20° and peaks 350 kHz below the design frequency. (Plot generated with W8WWV's *LBDXView* software.)

branches Z2 and Z3.

- Next you need to enter the value of the hybrid L and C in the Loads window. These values were calculated in Section 3.4.6.2 (2.18 µH and 436 pF for f_o = 3.65 MHz and Z_0 = 50 Ω).
- Enter 2.18 on line 5, column L, and 438 in both lines 6 and 8, in the column C of the Loads window.

The program is now ready to model your array for any frequency you enter. The fastest way to generate all the data across the band is to use the frequency sweep tool, provided in *EZNEC*.

Fig 11-46 shows the performance data of the Four Square fed by a theoretical 90° hybrid coupler feed system. Gain and dump power are good, but F/B ratio is very poor (around 14-17 dB). Such an array, which does not exist in reality but can be approached if we use a high quality λ/2 transmission line as phase inverter, would be a reasonably good transmit array but a poor performer on reception. This theoretical configuration, however, does not fulfill our minimum requirements as far as directivity (F/B) on any frequency of interest. That is, because an essential part is missing in the circuit, the shunt inductor across port 3.

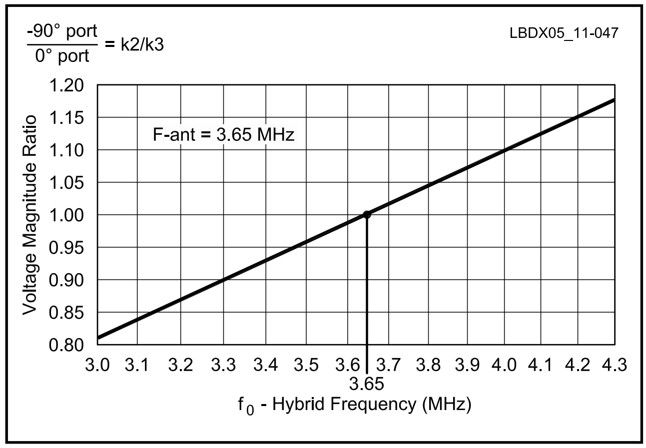


Fig 11-47 — This chart shows the hybrid design frequency (f_0-hybrid) as a function of the voltage magnitude ratio at the 0° port versus the –90° port. The chart was made for an operating frequency of 3.65 MHz. For other frequencies, just modify the horizontal scale (linear ratio).

3.4.6.5. W1MK's Single-Shunt Reactance Compensation Method

We have just seen that the performance of the Four Square using our theoretical "ideal" hybrid coupler is far from ideal. Fortunately there are ways to improve it.

3.4.6.5.1. The Principle

In Section 3.4.6 we have played with our hybrid modeling tool and described different special cases where a 90° phase voltage shift θ is obtained between port 2 and port 3. In Case 4 we showed how changing the hybrid design frequency f_o was instrumental in achieving a k-value of 1. In Fig 11-34 we showed the property of the lumped-coupled 90° hybrid that is responsible for this. **Fig 11-47** shows this effect and is based on the chart shown in Fig 11-34, but instead of showing power split, this chart shows the voltage (magnitude) ratio of the output ports (V2 and V3), centered on the operating frequency (f_a) of 3.65 MHz. These are the voltages we want to have under control, because, at the end of the current-forcing feed lines, these voltages are an image of the feed currents at the antenna elements. By changing the design frequency of the hybrid (f_o) we can adjust these voltages ratios without changing the phase shift (it remains 90°).

Table 11-11 shows how one of the complex impedances (Z3 = 60.7 – *j* 36 Ω) is transformed by connecting a shunt reactance in parallel with that port. The result is that the phase angle of the voltage in that port is changed so it has a 90° phase shift with respect to the voltage at the other port (V2/V3 = –*j*). In theory you can choose either port to do that.

The result is shown in Table 11-11 (Case 10). In this particular case, connecting a shunt impedance of +75.6 Ω across Z3 changed the value of Z3 to Z3' = 66.1 + *j* 32 Ω. This results in V3' undergoing a phase shift so that the phase difference between V2 and V3 is now –90°. However, the k value magnitude ratio V2/V3 is no longer 1, so we must bring it back to k = 1 by changing the f_a/f_o ratio, which is illustrated in Table 11-11, Case 11.

3.4.6.5.2. The Mathematics

We are interested in the ratio of V2 to V3. V2 and V3 must have equal magnitude (k = 1) and 90° out of phase. The basic formula is:

$$\frac{V2}{V3} = -j\left(\frac{1+\rho 2}{1+\rho 3}\right)\left(\frac{f_o}{f_a}\right) \qquad \text{(Eq 11-14)}$$

In addition we want to calculate the power at port 4 (dump power) and the reflected signal at port 1 (input port):

Dump Power = […] $(\rho 2+\rho 3)^2$

Reflected Power = […] $(\rho 2 - \rho 3)^2$

Table 11-11
Steps to Achieve a 90° Phase Shift (Single Shunt Method)
This table shows an example of how two steps make it possible to achieve a 90° phase shift and k = 1 for two random port 2 and port 3 load impedances.

Case	*R2*	*X2*	*R3*	*X3*	Z_0	*Port 4*	*Port 1*	*k*	θ	f_a	f_o	*Remarks*
9	53.7	22.5	60.7	–36	50	–17.1	–12.5	0.905	–66.8	3.65	3.65	
10	53.7	22.5	66.1	32	50	–11.9	–24.1	0.898	–90	3.65	3.65	add shunt of +75.6 Ω across Z3 (phase compensation)
11	53.7	22.5	66.1	32	50	–11.9	–24.1	1	–90	3.65	4.06	change fo to 4.06 MHz (k compensation)

The reflection coefficient at port 2 is ρ2 (which is a complex number) which can also be written in polar coordinates as 1 + ρ2 = M2 ∠θ2.

At port 3 we have 1 + ρ3 = M3 ∠θ3.

If we add shunt reactance across port 3 such that 1 + ρ3 = M3' ∠θ2, in other words that the phase angle becomes the same on both ports, we have:

$$\frac{V2}{V3} = -j\left(\frac{f_o}{f_a}\right) \times \left(\frac{M2}{M3}\right) \qquad \text{(Eq 11-15)}$$

where

f_o = design frequency of the hybrid

f_a = operating frequency of the array

If we now adjust f_o (change the hybrid design frequency) so that

$$f_o = f_a \times \left(\frac{M3}{M2}\right)$$

we have:

$$\frac{V2}{V3} = -j\,1$$

In other words V2 and V3 are equal in magnitude and 90° out of phase.

The same exercise can be done for a shunt element across Z2. The value of the ideal shunt reactance is computed by the following formula:

$$X_{shunt} = \frac{R^2 + X^2}{\dfrac{\left(Z_0 \times R + R^2 + X^2\right) \times \tan\theta 1}{Z_0}} - X \qquad \text{(Eq 11-16)}$$

where

R and X are the series impedances before the shunt is added

θ1 is the angle of (1 + ρ), ρ being the complex reflection coefficient of the port where the shunt is not going to be added.

3.4.6.5.3. Obtaining the Correct Values of Z2 and Z3

Before you can use the spreadsheet, you need to know the impedances Z2 and Z3 (see Figs 11-41 and 11-44). Whether or not your array is perfectly symmetrical, measure the impedances at ports 2 and 3 (see Section 3.4.6.4.4) and use these impedances in the *two-port-coupling.xls* spreadsheet to compute the correct values of Z2 and Z3 (see Section 3.4.6.4.5).

3.4.6.5.4. The Software Tool

The spreadsheet *single-shunt-hybrid-comp.xls* is based on W1MK's mathematics. It calculates the value of the required shunt inductance across Z3 (Case 1) or across Z2 (Case 2) to obtain the required 90° phase shift. It also calculates the new f_o required to obtain the voltage ratio (port 2 / port 3) as specified by the k-value.

Procedure

Enter Z2 and Z3 at the array (center) design frequency. In this example we use Z2 = 53.7 + *j* 22.5 and Z3 = 60.7 – *j* 36, which are the impedances obtained by modeling, Z_0-hybrid = 50 Ω and k = 1. The program calculates the shunt element across Z3 as a coil measuring 3.297 µH (75.6 Ω at 3.65 MHz) and specifies a new f_o of 4.063 MHz (**Fig 11-48**). It also does the calculation for a shunt on port 2, but in this particular case the port 3 compensation seems to be the better one. The software also computes the new impedance Z3' = 66.05 + *j* 32.51 Ω as well as the hybrid input impedance (port 1): Z_{in} = 50.4 + *j* 4.7 Ω, an almost perfect match for a 50 Ω feed line (lucky shot).

3.4.6.5.5. Calculating the Ideal Z_0

Note that the port 4 dump power is down only 11.9 dB vs the input power which is not very good. On the design frequency we should have at least 13 dB. We will soon find out that the port 4 power dump is the number 1 criterion in determining the operational bandwidth of the system.

The level of –11.9 dB is linked to the impedances Z2 and Z3 and the design impedance of the hybrid, Z_0. We also know that Z2 and Z3 depend on the characteristic impedance of the current-forcing (λ/4) cables we used. In practical terms: we can use either 50- or 75-Ω cable, which is best?

We can calculate that the ideal Z_0 (coupler design impedance). For minimum dump power:

$$Z_0 = \sqrt{Z2 \times Z3}$$

If we want to include the impedance of the feed lines, we come up with the following formula which calculates the ideal Z_0 (minimum dump power):

module 1	SHUNT REACTANCE CALCULATOR for single shunt element hybrid optimization		
	Real	Imag	
ENTER the impedance at port 2 : Z2 (Ω) →	53.70	22.50	Ω
ENTER the impedance at port 3 : Z3 (Ω) →	60.70	-36.00	Ω
ENTER the hybrid design impedance Zo →	50	Ω	
ENTER array design frequency fa (MHz) →	3.650	MHz	
Magnitude voltage at port 2/port 3: k→	1		
OUTPUTS FOR SHUNT ELEMENT ACROSS PORT 3			
Reactance Shunt element	75.60	Ω	
value shunt element =	3.297	µH	
new Z3 =	66.05	32.51	Ω
Fo =	4.063	MHz	
module 2a HYBRID COUPLER DESIGN (by W1MK)			
CASE 1: IMPORTED INPUT DATA FOR SHUNT ACROSS PORT 3			
(imported) fa (antenna frequency) =	3.650	MHz	
(imported) hybrid frequency →	4.063	MHz	
(imported) Zo Design impedance hybrid =	50.00	Ω	
	Real Part	Imag part	
(imported) Impedance load PORT 2 (R2) =	53.70	22.50	Ω
(imported) Impedance load PORT 3 (R3) =	66.05	32.51	Ω
RESULTS			
fo/fa =	1.113		
Hybrid L value (uH) =	1.96	µH	
Hybrid C value (pF) =	392	pF	
Ratio Voltage magnitude Port2/Port3 (k) =	1.000		
Phase angle Voltage port 2 vs. port 3 =	-90.07	°	
Power in Port 4 (vs. Pwr in Port 1) =	-11.9	dB	
Real part input impedance (port 1) =	50.35	Ω	
Imagainare part input impedance (port1) =	4.76	Ω	
Return loss (port 1) =	-26.5	dB	
SWR =	1.10		

Fig 11-48 — The spreadsheet calculates the required value of the shunt reactance across either Z2 or Z3 as well as the required new f_0 to obtain the voltage ratio k.

$$Z_0 = \sqrt{\frac{Z_{cable}{}^4}{2 \times R_{-90} \times (R_{-180} + R_0)}} \qquad \text{(Eq 11-17)}$$

where R_0, R_{-90} and R_{-180} are the real parts of the antenna impedances *at the antenna*. Take the example of our Four Square:

$R_0 = 3.4\ \Omega$
$R_{-90} = 44.6\ \Omega$
$R_{-180} = 65\ \Omega$

Putting these values in the above formula yields:

$Z_0 = 32\ \Omega$ for $Z_{cable} = 50\ \Omega$
$Z_0 = 72\ \Omega$ for $Z_{cable} = 75\ \Omega$

If we use $Z_0 = 72\ \Omega$ in our spreadsheet program, we see the dump power being reduced from –11.9 dB to –13.3 dB, which is the best that can be obtained with this single-shunt system. As a result the feed impedance will go up a little from 1.1:1 to 1.3:1 (vs 50 Ω) which is of no concern.

Warning: I found out that by going for the optimum Z_0 and thus achieving the lowest dump power on the design frequency (the middle of the band), one can slightly reduce the bandwidth over which we can achieve a level of –10 dB dump power or less. This acts like a high-Q circuit, high rejection in the middle of the curve, but narrower bandwidth.

3.4.6.5.6. In the Field

How do we determine the exact values of Z2 and Z3 in *our* array? We need to know Z2 and Z3 and use the *single-shunt-hybrid-comp.xls* spreadsheet to calculate the shunt element as well as f_o. It is almost certain that the values from our model and the real life values will be slightly different.

This is not a bad time to check the symmetry of the array by doing one of the tests described in Section 3.4.6.4.3. Make sure you measure a diagonal isolation of at least 30 dB. You may have to decouple other antennas or towers to achieve enough isolation. This test requires the use of a VNA. Try to get the isolation numbers as high as possible, with the array in all four directions.

To calculate Z2 and Z3, use the procedure outlined in Sections 3.4.6.4.4 and 3.4.6.4.5. Once you calculate Z2 and Z3 using the *two-port-coupling.xls* spreadsheet (see Fig 11-45), you can use these in the *single-shunt-hybrid-comp.xls* spreadsheet, which will calculate the shunt element and f_o (the hybrid design frequency).

Now you can make the shunt element and install it across the port it was calculated for. Next the hybrid coupler should be made based on the calculated f_o (see Section 3.4.6.1). No further alignment should be required!

As always, the proof of the pudding is in the eating. The ultimate proof of the array is in the measuring of the element drive currents, which can be done using one of the measurement procedures as described in Section 3.6. The VNA 2180 and the multiplexer, together with four current probes (Section 3.4.9.5) and the vector scope would of course be the ultimate tool (see Section 3.6.2).

3.4.6.5.7. Back to the Commercial Units

The two commercial hybrids (Comtek and DX Engineering) I measured have a little more reactance from the 180° transformer, but it was relatively close: ~95 Ω (4.1 µH on 3.65 MHz) to ~110 Ω (4.8 µH). That means that these commercial units are compensated to a certain degree. Whether or not this was pure luck or good engineering, I don't know. All that I know is that I have never seen anyone mentioning the utmost importance of this shunt reactance in relation to the correct functioning of the hybrid coupler drive system for the Four Square array.

If loaded with the impedances from our standard model (see Section 3.4.6.4.2) these commercial models perform quite well (see Figs 11-51 and 11-52 later in this chapter), and can be further optimized as shown later in Figs 11-61 and 11-62.

3.4.6.5.8. Operational Bandwidth of the Single-Shunt Compensated Hybrid Coupler Feed System

The procedure is somewhat different from what's described in Section 3.4.6.4.8 for the Four Square without compensation.

Step 1: Calculate Z2 and Z3 using the model *Ch11-4sq-quadfed-Z2-Z3legs.ez.*

Step 2: Run *single-shunt-hybrid-comp.xls* using the Z2 and Z3 values obtained in Step 1. This program gives you the single shunt value (across either Z2 or Z3) and the hybrid component values.

Step 3: Using these values, now run the *EZNEC* model that includes the hybrid and the shunt

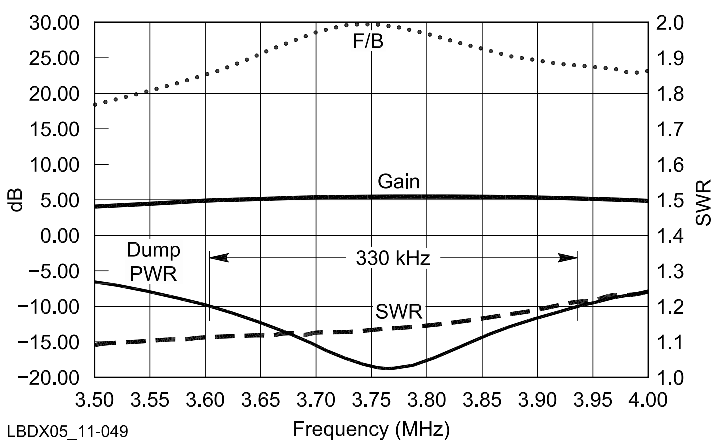


Fig 11-49 — Performance data for the array using antenna elements that are self-resonant on 3.65 MHz (19.7 meters long, 60 mm diameter). (Plot generated with W8WWV's *LBDXView* software.)

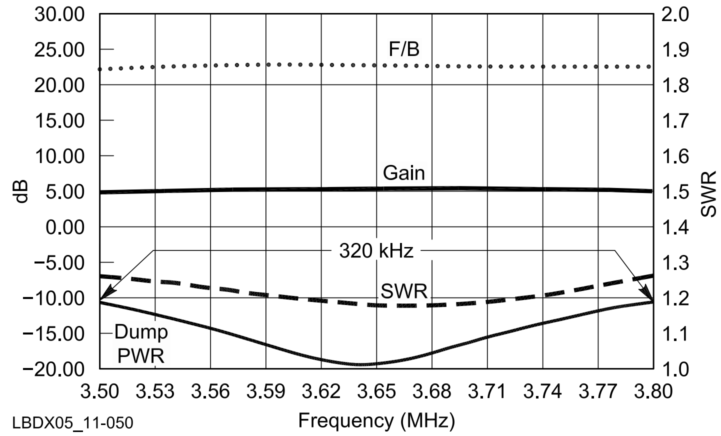


Fig 11-50 — Performance data for the array using antenna elements that are self-resonant on 3.54 MHz (20.50 meters long, 60 mm diameter). (Plot generated with W8WWV's *LBDXView* software.)

elements (*Ch11-4sq+hyb-3.65-k = 1-1sh.ez*). Run *EZNEC* in swept frequency mode.

Step 4: Use the data file produced by *EZNEC* as the import file for the *LBDXView* program, which will generate data graphs and radiation patterns.

Fig 11-49 shows the results for a shunt coil value of 3.3 µH, value which was calculated with the *single-shunt-hybrid-comp.xls* spreadsheet (see Fig 11-48). Looking only at the F/B numbers (for a given radiation angle) can be misleading. You can have a back lobe that is down 30 dB at 20° vertical angle, but down only 15 to 20 dB at 60°, which is the angle at which the stronger local stations arrive on. Therefore we should look at the vertical pattern as well.

From Fig 11-49 we learn that the F/B peak and the port 4 dump power minimum occur at approximately the same frequency (3.75 MHz), which, however, is not in the middle of the band (3.65 MHz).

To correct for this we need to change the resonant frequency of the radiating elements to a lower frequency. Some cut-and try tells us that lengthening the elements from 19.7 meters to 20.5 meters moves everything into place (**Fig 11-50**). This lengthening of the elements in order to move the dump curve right in sync with the F/B curve is the tricky part of the single-shunt compensation method

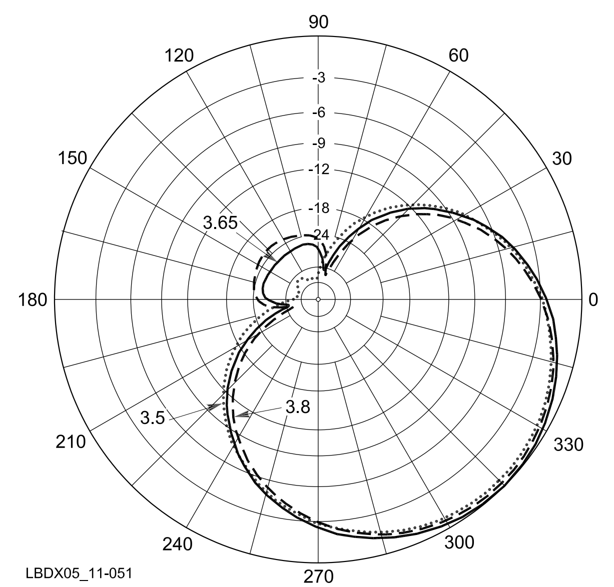


Fig 11-51 — Horizontal radiation patterns at mid-band and at the two band edges for the single-shunt compensated (4.1 µH shunt across Z3) Four Square with elements resonated on ~3.54 MHz. (Plot generated with W8WWV's *LBDXView* software.)

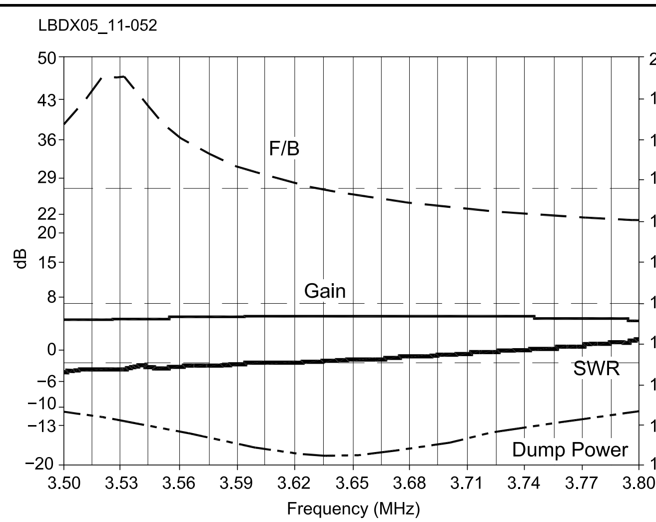


Fig 11-52 — Performance data of a commercial hybrid coupler (shunt inductance of transformer = 4.1 µH). We optimized the self resonant frequency of the elements in such a way that the lowest dump power point moved to the middle of the band. (Plot generated with W8WWV's *LBDXView* software.)

(*Ch11-4sq+hyb-3.65-k = 1-1sh-el lengthened.ez*). With two shunts (see Section 3.4.6.6) this problem does not exist. **Fig 11-51** shows the corresponding radiation patterns; note that these patterns remain almost the same over 300 kHz.

It is clear that with this system, the bandwidth limiting parameters are both the F/B and the port 4 dump power (for which we set an operational limit of >–10 dB vs the input power), which is better than –10 dB over 300 kHz! In a nutshell: this system has an impressive operational bandwidth of just over 300 kHz on 80 meters. Compare that with only 80 kHz for the L-network feed system (Section 3.4.5.9).

All of this confirms the "magic" rule that you always need to use elements in a Four Square (fed by a Comtek hybrid coupler) that are resonant below your array center design frequency. At last we know why!

In Section 3.4.6.5.7 I explained how the shunt inductance of the transformer used in the commercial hybrid coupler models achieves a certain degree of compensation. **Fig 11-52** shows you the results, which are really quite good when you compare them with those shown in Fig 11-50 (modeling file: *Ch11-4sq+hyb-3.65-commercial.ez*). For further optimization see Section 3.4.6.6.6. Keep in mind that these models were made assuming Z2 = 53.7 + *j* 22.5 Ω and Z3 = 60.7 – *j* 36.0 Ω — impedances for our standard Four Square model described in Section 3.4.6.4.2.

3.4.6.5.9. Managing the Band Shift

The best way to bring the elements to resonance on the required frequency is to make a model of the Four Square including the feed lines and 180° transformer (see the file *Ch11-4sq-quadfed-Z2-Z3legs.ez*). When you run that file you have as source two feed points that we call Z2 and Z3. All you need to do is to add a little series coil in the Loads window to make Z2 (the parallel impedance of the feed lines going to the center elements) real at the new (lower) frequency. We are putting a little loading at the bottom of the elements (maybe 2 to 3 turns of a diameter of 5 cm).

In this particular case, we need to add small coils of + *j* 18.5 Ω inductance (0.8 µH on 3.65 MHz) in series with the feed points at each element. The new Z2 is now 63.1 + *j* 0 Ω while Z3 has become 35.9 – *j* 40 Ω what you can read from the Scr Dat window. The resonant frequency of our four verticals has shifted to exactly 3.520 MHz. (Check it by modeling a single vertical with the same loading coil, identical to these used in the array.)

Now run the array with the four small loading coils and with Z2 = 63.2 Ω and Z3 = 35.9 – *j* 40 Ω on the *single-shunt-hybrid-comp.xls* worksheet. The calculator will tell you to put a coil of 3.1 µH (72.2 Ω) across Z3 and to use a hybrid with f_o = 4.034 MHz. The new Z3' becomes 80.47 + *j* 0 Ω. The port 4 dump power is suppressed by 15.1 dB and the input SWR <1.1:1.

If you use a hybrid with f_o = 3.65 (90% of 4.034 MHz), k will be 0.9 instead of 1 at 3.65 MHz, and the directional patterns will be somewhat better than if k = 1.

Conclusion: A hybrid with f_o = 3.65 MHz, and a typical Four Square, with the four verticals resonated at ~3.520 MHz, plus a single (total) shunt inductance of 3.1 µH across Z3, is about the best and simplest single-shunt optimization you can do. Remember that the 180° transformer by itself already delivers a good deal of shunt reactance (see Section 3.4.6.5.7).

3.4.6.5.10. Improving the Directivity

In Section 3.4.6 we learned that by designing the hybrid for a frequency (f_o) different from the array design frequency (f_a), we can obtain different voltage magnitudes in the 0° and the –90° ports (k ≠ 1).

By doing so we can easily adjust the magnitude of the feed currents to the two center elements of the Four Square array to be different from the current magnitudes in the front and back element. Let's analyze what this is good for.

If we reduce the feed current magnitude to the center element down to 80% of the current magnitude at the front and back elements, we obtain much better looking back lobes (see **Fig 11-53**).

The branch impedances Z2 and Z3 have *not* changed: Z2 = 53.7+ *j* 22.5 Ω and Z3 = 60.7 – *j* 36 Ω are the same as for pure quadrature. In other words: Z2 and Z3 do not change if we change the current magnitude or phase shift between the two branches (try it with the modeling file *Ch11-4sq-quadfed-Z2-Z3legs.ez*) which confirms the earlier statement that there is virtually no mutual coupling between the two branches (see also Section 3.4.6.4.1).

All we need to do is to run the *single-shunt-hybrid-comp.xls* spreadsheet and change k to 0.85 (**Fig 11-54**). As a result f_o has changed from 4.063 to 3.454 MHz. The results are quite spectacular as shown in Fig 11-53B.

Operational Bandwidth

The operational bandwidth remains approximately 300 kHz and, as in the k = 1 case, the array elements need to be designed for a frequency of approximately 150 kHz below the center of the band in this 80 meter example. The procedure

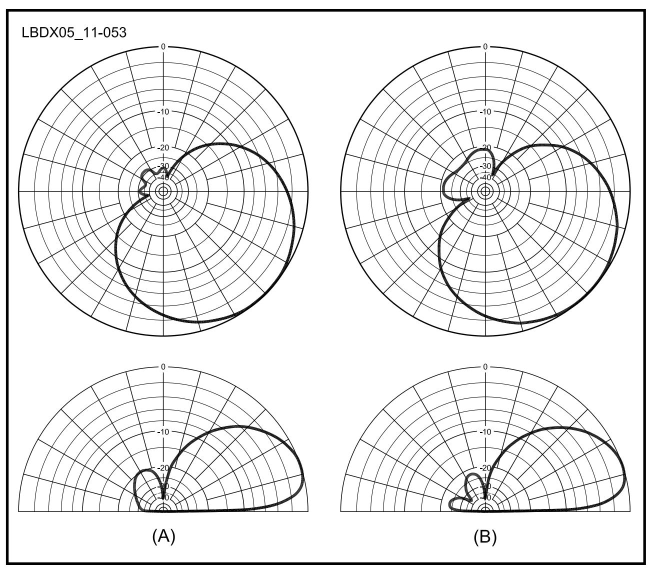


Fig 11-53 — Using the *4sq-hyb-w1mk.xls* spreadsheet one can, for example, change the hybrid design frequency (f_0) until you reach a desired voltage magnitude ratio (port 2 vs port 3). In this example we achieve a ratio of 0.85 by changing the f_0 from 4.09 MHz to 3.455 MHz. Whereas the horizontal directivity as shown in case A may look OK, we must realize that this pattern is for radiation angle of 20°. At 60 degrees the F/B would be less than 20 dB. Pattern B is the preferred pattern to attenuate nearby signals arriving at relatively high wave angles.

module 2a	HYBRID COUPLER DESIGN (by W1MK)		
CASE 1: IMPORTED INPUT DATA FOR SHUNT ACROSS PORT 3			
(imported) fa (antenna frequency) =	3.650	MHz	
(imported) hybrid frequency →	3.454	MHz	
(imported) Zo Design impedance hybrid =	50.00	Ω	
	Real Part	Imag part	
(imported) Impedance load PORT 2 (R2) =	53.70	22.50	Ω
(imported) Impedance load PORT 3 (R3) =	66.05	32.51	Ω
RESULTS			
fo/fa =	0.946		
Hybrid L value (uH) =	2.30	µH	
Hybrid C value (pF) =	461	pF	
Ratio Voltage magnitude Port2/Port3 (k) =	0.850		
Phase angle Voltage port 2 vs. port 3 =	-90.07	°	
Power in Port 4 (vs. Pwr in Port 1) =	-11.9	dB	
Real part input impedance (port 1) =	47.59	Ω	
Imagainare part input impedance (port1) =	6.70	Ω	
Return loss (port 1) =	-22.8	dB	
SWR =	1.16		

Fig 11-54 — We can simply change the k-value in the spreadsheet which will calculate the new design frequency (f_0) for the hybrid (according to the principle shown in Fig 11-47).

to evaluate the operational bandwidth is explained in Section 3.4.6.5.8.

Based on modeling file *Ch11-4sq+hyb-3.65 = k = 0.8 = 1sh = -el lengthened.ez* we developed the charts and patterns shown in **Figs 11-55** and **11-56**.

Conclusion

With the simple addition of a correctly dimensioned shunt coil across Z3, we can turn the plug-and-play hybrid coupler into a top-performing Four Square that is also a great performer on receiving.

3.4.6.6. W1MK's Two-Shunt Hybrid Optimization Approach (Phase Compensation)

Based on the results of earlier experiments with lumped parallel elements at port 2 and 3 described in "Phase Correction of the Quadrature Hybrid" published in 2005 in *NCJ* (Ref 984) and inspired by the L-network optimizing approach, and more specifically by the use of the hybrid network off its design frequency, Rob, W1MK, developed his *phase compensation design*, or *two-shunt compensation system* approach, shown in **Fig 11-57**.

3.4.6.6.1. The Principle

In this system the perfect 90° split is achieved by making the reflection coefficients ρ2 and ρ3 (at port 2 and 3) real. To achieve this, shunt reactances (L or C) will be placed across port 2 and 3, as shown in **Table 11-12**. Case 12 shows the situation we start with before doing the optimization. In this case θ = –66.8° (very bad!) and k = 0.905. After connecting the required shunt reactances across Z2 and Z3 we obtain real but different impedances (Z2 = 63.1 Ω and Z3 = 82.1 Ω). These values cause real but different reflection coefficients resulting in perfect 90° phase shift but different voltage magnitudes at ports 2 and 3 of the hybrid (Case 13 of Table 11-12).

The ratio between these two voltages depends on the

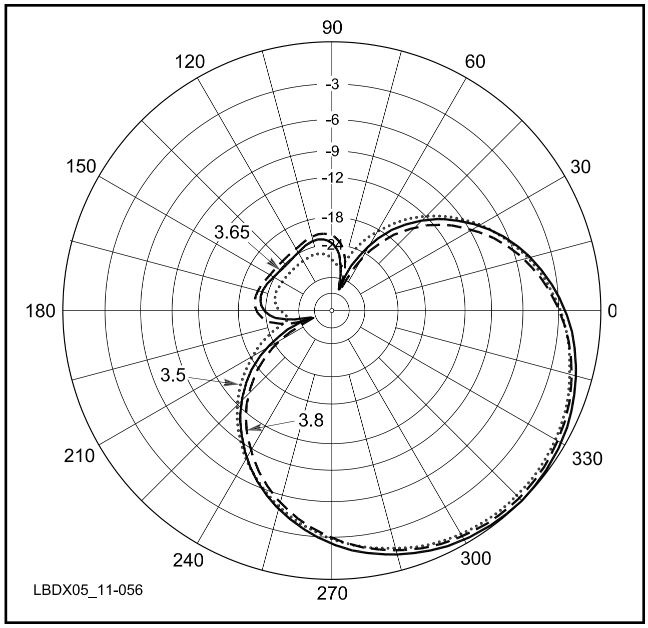


Fig 11-56 — Horizontal radiation patterns at mid-band and at the two band edges for the single shunt compensated Four Square using k = 0.85. (Plot generated with W8WWV's *LBDXView* software.)

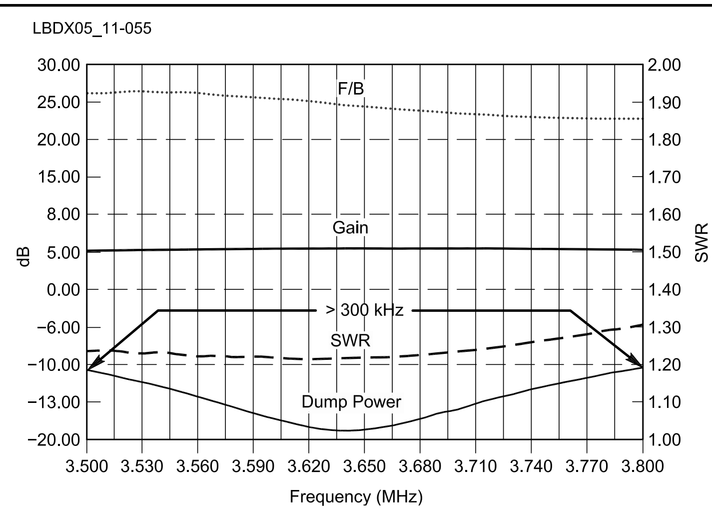


Fig 11-55 — In this case we used radiating elements measuring 20.5 meters in length (60 mm OD) and k = 0.85. (Plot generated with W8WWV's *LBDXView* software.)

Table 11-12
Steps to Achieve a 90° Phase Shift (Two Shunt Method)
This table shows an example of how two steps make it possible to achieve a 90° phase shift and k = 1 for two random port 2 and port 3 load impedances.

Case	*R2*	*X2*	*R3*	*X3*	*Z_0*	*Port 4*	*Port 1*	*k*	*θ*	*fa*	*fo*	*Remarks*
12	53.7	22.5	60.7	–36	50	–17.1	–12.5	0.905	–66.8	3.65	3.65	
13	63.1	0	82.1	0	50	–14.9	–22.7	0.898	–90	3.65	3.65	add –150 Ω across Z2 and +138 Ω across Z3
14	63.1	0	82.1	0	50	–14.9	–27	1	–90	3.65	4.06	change fo to 4.06 MHz (k compensation)

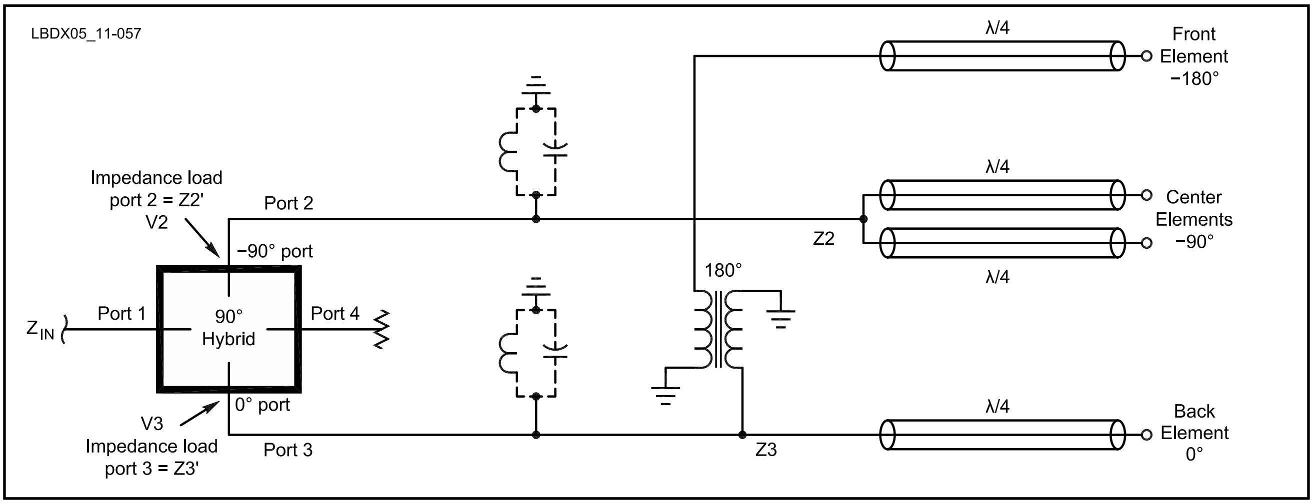


Fig 11-57 — The phase compensation method (also called "unequal resistive loads" method or "admittance compensated circuit") by Robye, W1MK. See text for details.

ratio of the (real) impedances involved. This ratio is given by

$$\frac{f_o}{f_a} = \frac{Z3}{Z2}$$

where
f_o = hybrid design frequency
f_a = array design (center) frequency

As in the case of the single-shunt optimization system, we will now change f_o so that k = 1 (Case 14).

3.4.6.6.2. The Mathematics

Rob, W1MK, developed the mathematical equations for this two-shunt compensation approach. In the hybrid coupler (designed around 50-Ω system impedance) the ratio of the voltages is given by:

$$\frac{V2}{V3} = -j\,\frac{f_o(1+\rho 2)}{f_a(1+\rho 3)} \qquad \text{(Eq 11-18)}$$

where
f_o = the design frequency for the hybrid
f_a = the design frequency of the array
ρ2, ρ3 = reflection coefficient at port 2 and 3:

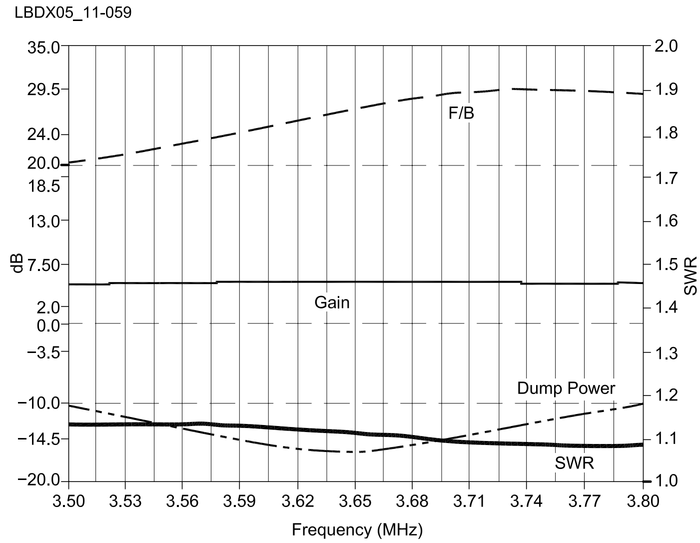
$$\rho 2 = \frac{Z2' - 50}{Z2' + 50} \text{ and } \rho 3 = \frac{Z3' - 50}{Z3' + 50} \qquad \text{(Eq 11-19)}$$

Port 3 is the 0° (reference) port and port 2 the –90° port

	HYBRID COUPLER OPTIMIZATION TWO SHUNT COMPENSATION DESIGN SYSTEM -w1mk -			
	Enter data in yellow background cells			
	CALCULATING THE SHUNT ELEMENTS			
1	Enter fa (design freq) array →	3.650	MHz	
2	Enter k-value →	1		
		Real part	Imag part	
3	(-90 ° branch) Enter Z2 →	53.70	22.50	Ω
4	Branch 2 shunt element →	-150.66	Ω	
5	→	289.42	pF	
6	Z2' (at port 2) →	63.13	Ω	
		Real part	Imag part	
7	(0° branch) Enter Z3 →	60.70	-36.00	Ω
8	Branch 3 shunt element →	138.35	Ω	
9	→	6.03	uH	
10	Z3' (at port 3) →	82.05	Ω	
	HYBRID INPUT DATA			
11	Enter Zo Hybrid →	50.0	Ω	
	OUTPUT DATA			
12	Frequency corr. factor =	1.1135		
13	New Hybrid fo =	4.064	MHz	
14	L-Hybrid =	1.96	uH	
15	C-Hybrd =	392	pF	
16	Dump port (port 4) power ratio =	-14.9	dB	
17	Real part Zin (port 1) =	49.20	Ω	
18	Imag. part Zin (port1) =	4.35	Ω	
19	Port 1 return loss (dB) =	27.1	dB	
20	Port 1 SWR =	1.09		

Fig 11-58 — The spreadsheet that calculates the values of the compensating elements as well as the f_0 of the hybrid coupler (and more). See text for details.

Fig 11-59 — Performance parameters of a quarter-wave spaced, quadrature fed Four Square, using the two-shunt optimization circuit developed by W1MK. (Plot generated with W8WWV's *LBDXView* software.)



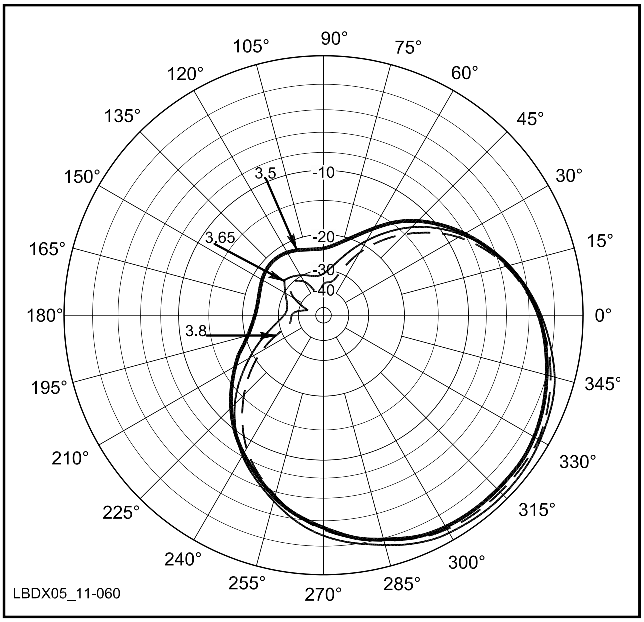


Fig 11-60 — Horizontal radiation patterns at mid-band and at the two band edges for the two-shunt optimized Four Square. (Plot generated with W8WWV's *LBDXView* software.)

We want to make V2/V3 = $-j$, so we can rewrite Eq 11-19 as:

$$f_o(1+\rho2) = f_a(1+\rho3) \text{ or } f_o = f_a \times \frac{1+\rho3}{1+\rho2} \qquad \text{(Eq 11-20)}$$

If we redesign the hybrid coupler for a new design frequency (f_o), V2 will equal V3 in magnitude (with 90° phase shift) and the 0° as well as the –90° branch will deliver equal current magnitude at the antenna elements. The recalculation of the hybrid design frequency is based on the principles shown in Figs 11-34 and 11-47.

Note that adding shunt inductance or capacitance at the end of a quarter-wave feed line is really the same as adding series reactance at the other (antenna) end of the line. At the antenna elements, we would call it "resonating the element with a series lumped impedance."

3.4.6.6.3. The Software Tool

Based on the principle explained above, W1MK developed the spreadsheet *two-shunt-hybrid-comp.xls* shown in **Fig 11-58**. Whereas the simplified formulas (above) apply for Z_0-hybrid = 50 Ω only, the spreadsheet lets you freely specify Z_0.

In Section 3.4.6.5.3 we explained how you obtain the correct values of Z2 and Z3 for the array you are building. For simplicity, let us work with the impedances Z2 and Z3 as obtained through using the *EZNEC* modeling file *Ch11-4sq-quadfed-Z2-Z3legs.ez*: Z2 = 53.7 + j 22.5 Ω and Z3 = 60.7 – j 36.0 Ω. Other data to be entered in the spreadsheet are f_a (operating frequency), k (voltage ratio port 2 vs port 3) and Z_0-hybrid.

The spreadsheet calculates the port 2 and port 3 loads (289 pF and 6.03 µH) and shows the real impedances Z2' and Z3' (Z2' = 63.13 Ω, Z3' = 82.05 Ω). Also calculated are the port 4 dump power (–14.9 dB), the hybrid network input impedance (49.29 + j 4.37 Ω), and the related return loss and the SWR vs the system impedance Z_0. The entire procedure for using the spreadsheet is explained in detail in the spreadsheet *two-shunt-hybrid-comp.xls*. We can use the results of this spreadsheet in the *4sq-hyb-w1mk.xls* spreadsheet, which will confirm the results.

For this two-shunt system you can also calculate Z_0 for which the port 4 dump power is smallest (see Section 3.4.6.5.5). You must realize that when you go for the greatest rejection in the middle of the band, the rejection at the band edges will be somewhat less than if you use a somewhat lower Z_0.

One more way to look for less port 4 wasted power is to play with the dump load impedance and change the value of R, or even introduce some parallel reactance (positive or negative). All of this can easily be done with the *4sq-hyb-w1mk.xls* spreadsheet. But if you do that, keep an eye on k and θ because they will change as well!

If you really think you need more than –10 dB rejection at the band edges, you can of course split the band in two

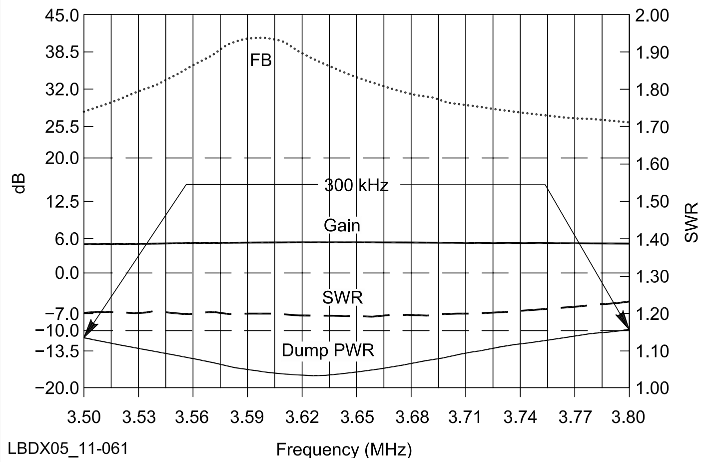


Fig 11-61 — The commercial hybrid coupler shown in Fig 11-52 with an added 100 pF capacitor across Z2, which moved the F/B peak up higher in frequency. With very little modification the off-the-shelf unit is now working "as well as it can." (Plot generated with W8WWV's *LBDXView* software.)

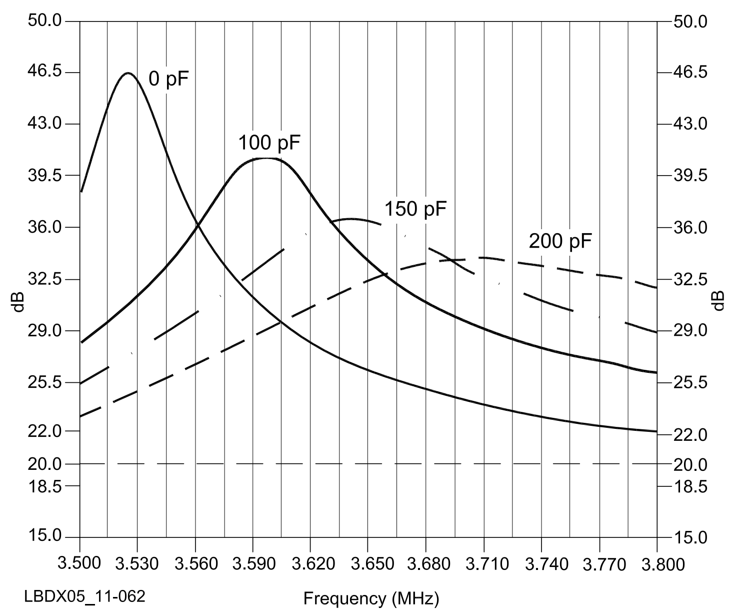


Fig 11-62 — The F/B ratio curves as a function of the value of the capacitor across Z2. (Plot generated with W8WWV's *LBDXView* software.)

sections and use two sets of shunt elements that you select by a relay.

3.4.6.6.4. Measuring and Adjusting

Let us assume the above numbers were arrived at in the design phase, and we now have to construct the feed system.

At this point in time it is good to check the symmetry of the array by doing one of the tests described in Section 3.4.6.4.3. Make sure you measure a diagonal isolation of at least 30 dB. You may have to decouple other antenna(s) or tower(s) to achieve enough isolation. This test requires the use of a VNA. Try to get the isolation numbers as high as possible with the array in all four directions.

To be able to calculate the values of the shunt elements (which is done in the *two-shunt-hybrid-comp.xls* spreadsheet), we need to know the exact value of Z2 and Z3 in *our* array, which may be a little different from the model. Follow the procedure outlined in Sections 3.4.6.4.4 and 3.4.6.4.5. Once you calculate Z2 and Z3 using the *two-port-coupling.xls* spreadsheet (see Fig 11-45), you can use these in the *two-shunt-hybrid-comp.xls* spreadsheet, which will calculate the shunt elements and f_o.

Now you can make the shunt elements and install them across port 2 and port 3. Next, the hybrid coupler should be made based on the calculated f_o (see Section 3.4.6.1). No further alignment is required!

The ultimate proof of the array is in the measuring of the element drive currents, which can be done using one of the measurement procedures as described in Section 3.6. The VNA 2180 and the multiplexer, together with four current probes (Section 3.4.9.5) and with the vector scope would of course be the ultimate tool (see Section 3.6.2).

3.4.6.6.5. Operational Bandwidth of the Quadrature Four Square Fed By W1MK's Two-Shunt Optimized Hybrid Coupler Feed System

The results for W1MK's two-shunt compensation system are shown **Figs 11-59** and **11-60**. For this model we adjusted the element lengths slightly from 19.70 to 19.85 meters in order to achieve perfect symmetry of the port 4 dump power

	HYBRID COUPLER OPTIMIZATION TWO SHUNT COMPENSATION DESIGN SYSTEM -w1mk -			
	Enter data in yellow background cells			
	CALCULATING THE SHUNT ELEMENTS			
1	Enter fa (design freq) array →	3.650	MHz	
2	Enter k-value →	0.85		
		Real part	Imag part	
3	(-90 ° branch) Enter Z2 →	53.70	22.50	Ω
4	Branch 2 shunt element →	-150.66	Ω	
5	→	289.42	pF	
6	Z2' (at port 2) →	63.13	Ω	
		Real part	Imag part	
7	(0° branch) Enter Z3 →	60.70	-36.00	Ω
8	Branch 3 shunt element →	138.35	Ω	
9	→	6.03	uH	
10	Z3' (at port 3) →	82.05	Ω	
	HYBRID INPUT DATA			
11	Enter Zo Hybrid →	50.0	Ω	
	OUTPUT DATA			
12	Frequency corr. factor =	0.9465		
13	New Hybrid fo =	3.455	MHz	
14	L-Hybrid =	2.30	uH	
15	C-Hybrd =	461	pF	
16	Dump port (port 4) power ratio =	-14.9	dB	
17	Real part Zin (port 1) =	49.83	Ω	
18	Imag. part Zin (port1) =	7.33	Ω	
19	Port 1 return loss (dB) =	22.7	dB	
20	Port 1 SWR =	1.16		

Fig 11-63 — By simply changing k from 1 to 0.85 we obtained the new value for f_0 (3.484 MHz).

with respect to the center design frequency of 3.65 MHz (file: *Ch11-4sq+hyb-3.65-k = 1-2sh.ez*). This can of course also be achieved by installing a small loading coil at the base of each element (which is easier to adjust than the element length) The 20 dB F/B (at 20° wave angle) bandwidth is better than 300 kHz, and the –10 dB dumped power bandwidth is just as wide.

3.4.6.6.6 Modifying a Commercial Unit

To see what the influence would be of using an off-the-shelf hybrid coupler with f_0 = 3.65 MHz (what we expect from a commercial unit), we added a capacitor of 100 pF across the –90° (Z2) port. In fact we now have a two-shunt compensation: the inductance of the 180° transformer across Z3 and the 100 pF capacitor across Z2. The results are shown in **Fig 11-61**. (See modeling file *Ch11-4sq+hyb-3.65-commercial+1CAP.ez*.) Compare with the data of Fig 11-52 where we used the same shunt coil (the 180° transformer) at Z3 (4.1 µH) but where the 100 pF capacitor across Z2 is missing.

Fig 11-62 shows the impact of the capacitor values across Z2. If you are a CW only operator, do not use any capacitor. If you like SSB best, use 200 pF. The ideal situation would be to use 150 pF when operating between 3.6 and 3.8 MHz (phone) and no capacitor when operating below 3.6 MHz (CW). With a simple relay to switch the capacitor, you will have better than ~30 dB front to back in both cases! Note that the dumped power curve hardly changes at all with the different C values.

Remember that this model is done on a perfectly symmetrical Four Square (see data in Section 3.4.6.4.2) where Z2 = 53.7 + *j* 22.5 Ω and Z3 = 60.7 – *j* 36.0 Ω. A different Four Square configuration with different values of Z2 and Z3 will inevitably give slightly different results. The obtained results are very good. All of this indicates that if you want to optimize your commercial unit for top notch performance, it can easily be done.

3.4.6.6.7. Improving the Directivity

In Section 3.4.6.5.10 we learned that in a quadrature-fed Four Square we can improve the directivity by reducing the current in the center elements. At the end of the current-forcing feed lines this means reducing the voltage at port 2 vs port 3. This also means changing the k-value of 1 to a lower value, such as k = 0.85. We can specify the k-value in the *two-shunt-hybrid-comp.xls* spreadsheet (or change it in the *4sq-hyb-w1mk.xls* spreadsheet).

Fig 11-63 shows the screen of the *two-shunt-hybrid-comp.xls* spreadsheet, where we changed k from 1 to 0.85 (all other data remaining the same) as shown in **Fig 11-64**. This simply changed the f_0 from 4.064 MHz to 3.484 MHz (4.064 × 0.85 = 3.484), without altering the phase shift as the impedances Z2 and Z3 and the shunt element values have not changed.

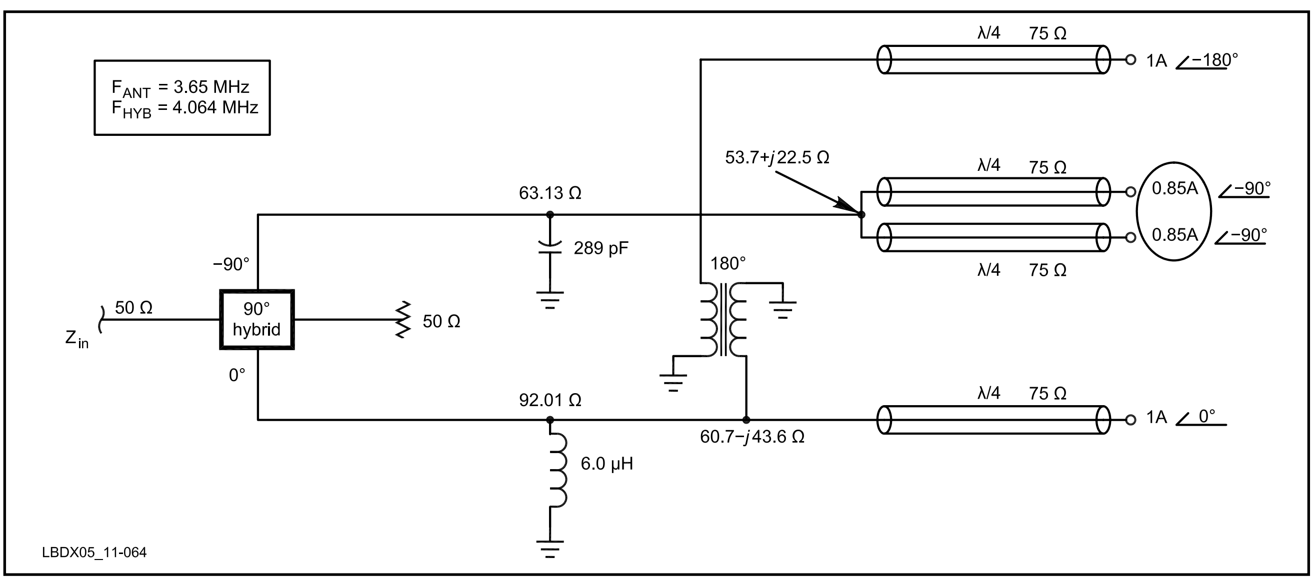


Fig 11-64 — Final hybrid coupler with system with simple current magnitude optimization for a Four Square array.

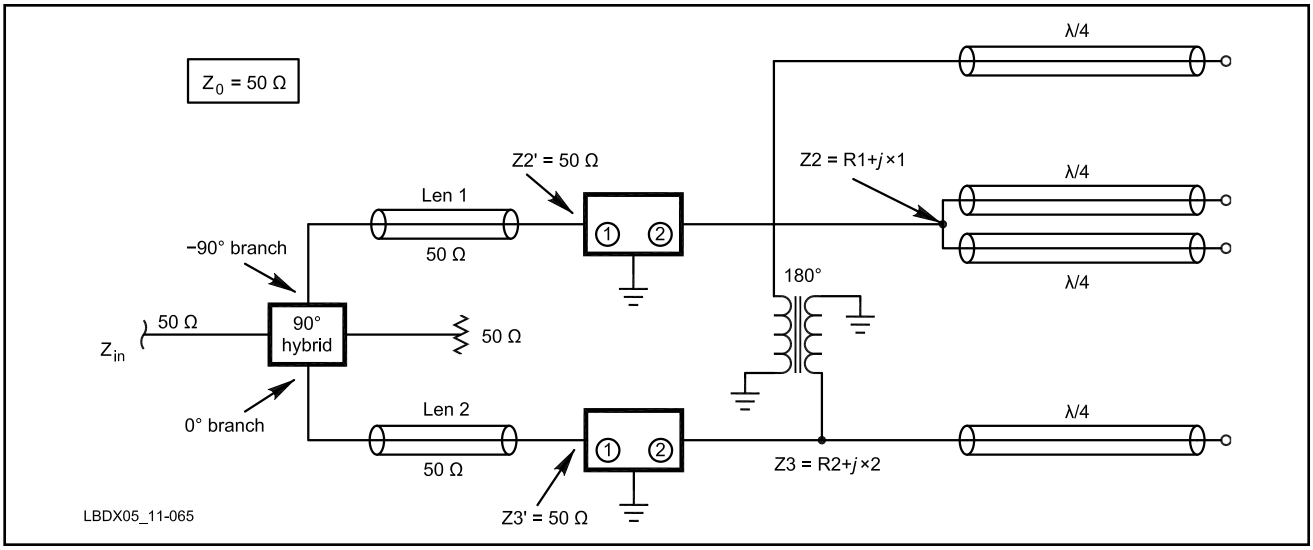


Fig 11-65 — Diagram of the "double network hybrid optimization" system as developed by W8WWV.

The directivity results are the same as obtained by reducing the k-value from 1 to 0.85 in case of the single-shunt coil optimization system. The operational bandwidth remains approximately 300 kHz.

3.4.6.7. The Double Network (W8WWV)

In a third system, shown in **Fig 11-65**, two L-networks are used between Z2 and the hybrid and between Z3 and the hybrid. The purpose of these L-networks is to match the impedances Z2 and Z3 to the 50 Ω system impedance of the hybrid, so that the hybrid provides a perfect split (90° phase shift and k = 1) between its ports 2 and 3. That's why this system is also called the "matched port" optimization system.

The introduction of these L-networks, however, has a couple of side effects: The L-networks inevitably introduce additional phase shift, which (inevitably) will be different in each branch. This can be compensated for by inserting a piece of 50-Ω coax in the leg where the phase shift is smallest. The other side effect is that the two L networks normally have different voltage transformation ratios (k-factors); these networks not only transform impedances but also voltages (and currents)! This can be taken care of by changing the hybrid design frequency to obtain identical voltage magnitudes beyond the L-networks (at the beginning of the feed lines).

The design procedure is a little more complex than for the two preceding systems. As the bandwidth performance appears to be slightly inferior to what is obtained with the two foregoing systems, this system is not covered in detail in this book. You can however find all the details in a document on this book's CD (*Double-network-hybrid-optimization-W8WWV.pdf*).

Ch11-4sq-w8wwv-com+hyb.ez is the *EZNEC* modeling file that can be used to assess the operational bandwidth using the procedure outlined in Section 3.4.6.5.8. The operational bandwidth is very similar to what we found in the other systems (300 kHz).

3.4.6.8. Making Your Own Hybrid Coupler Feed System

Let's go step-by-step through the entire procedure for developing and making a two-shunt W1MK optimized hybrid coupler feed system for a Four Square array.

1) As a very first step it's always a good idea to make an accurate model of the antenna. File *Ch11-4sq-quadfed-Z2-Z3legs.ez* is a good example. It includes the feed lines and the ideal phase reversal transformer. The elements in the model are all those shown in Fig 11-41. Let us assume we have calculated Z2 = 53.7 + *j* 22.5 and Z3 = 60.7 – *j* 36.

2) Plug these values in the *two-shunt-hybrid-comp.xls* software to calculate the value of the required shunt elements: 289 pF (–151 Ω) across Z2 and 6.03 µH (+138 Ω) across Z3.

3) To design and make such a system you require a quality antenna bridge analyzer, a 1000 pF variable capacitor and a 0-25 µH roller inductor. If you want to do the diagonal isolation test, you will need a VNA.

4) Resonate the four vertical elements on exactly 3.65 MHz (uncoupled!). Try to make them physically the same length and taper schedule.

5) Make sure your current-forcing feed lines are made with 75-Ω cable and that they are cut for exactly 3.65 MHz.

6) Make sure the four elements of your array are as identical as possible (physically and electrically). Measure their self impedance and trim the length and add radials as necessary (see Section 3.5.3.1).

7) Make a λ/2 long transmission line for the center of the band (eg 3.65 MHz on 80 meters) from the best quality coax you have. I use ⅞-inch coax for this purpose. This line will serve as an ideal phase reversal transformer, introducing no shunt inductance.

8) Connect the feed lines and the 180° transformer (the ½ wave feed line) as in Fig 11-41.

9) Using a VNA measure diagonal isolation as explained in Section 3.4.6.4.2.

10) Take the necessary steps to increase the increase the isolation to at least 30 dB (for example, decoupling other towers within 1 λ of your Four Square).

11) Measure Z2 and Z3, and check if the measurements are "independent" from one another (proof of good isolation).

(You can omit Steps 7 though 11 if you are not curious enough to know how "balanced" your array is.)

12) Build a 180° (phase reversal) transformer as described

in Section 3.4.6.3.5. Measure its shunt inductance value (on 3.65 MHz), using an instrument such as the AIM 4170 impedance analyzer or the VNA 2180.

13) Incorporate the new transformer in your array (it should replace the $\lambda/2$ wave long transmission line if you did the diagonal isolation test).

14) Using a quality antenna analyzer measure the values of Z2 and Z3.

15) Do these measurements for the four directions. If the diagonal isolation is good enough, all impedance measurements we will do will be identical, or nearly so, in all four directions.

16) Measure Z22, Z33 , Z23 and Z22 + Z33 in parallel and run the *two-port-coupling.xls* spreadsheet to calculate Z2 and Z3. Do that for each of the four directions (see Sections 3.4.6.4.4 and 3.4.6.4.5). The results should be very similar to what we calculated with our model if the array is adequately balanced.

17) Plug the Z2 and Z3 values in the *two-shunt-hybrid-comp.xls* spreadsheet to calculate the values of the two shunt elements.

18) Make the shunt elements. Apply the following formulas:

$$C = \frac{10^6}{2\,\pi\,f\,X_C} \text{ and } L = 2\,\pi\,f\,X_L \qquad \text{(Eq 11-21)}$$

where
C is in pF
L is in μH
f is in MHz

19) Copy the value of f_o (the hybrid design frequency) from the *two-shunt-hybrid-comp.xls* spreadsheet.

20) Now build the hybrid coupler for this f_o (see Section 3.4.6.1). The values of L and C are shown in the *two-shunt-hybrid-comp.xls* spreadsheet.

21) Now you can incorporate the hybrid network, the transformer and the shunt element(s) in the box containing the direction switching relays (as shown in Fig 11-136 later in this chapter).

22) The final test is to measure the element currents. This can be done indirectly with a system as described in Section 3.6, or if you have VNA and a multiplexer box, using the system described in Section 3.6.2.

3.4.6.9. Converting a Commercial Unit?

I would certainly first try the KISS system (keep it simple, stupid). Refer to Section 3.4.6.6.6. Chances are that simply adding a capacitor across the Z2 port (–90° port) you will get much better performance from your array.

With the phase-inversion transformer of the commercial unit in place, measure the port 2 and port 3 impedances as explained in Section 3.4.6.4.4 and calculate the values of Z2 and Z3 as explained in Section 3.4.6.4.5.

Now measure the f_o of the hybrid coupler (see Section 3.4.6.2.1). If you have the values of C and L, you can also calculate the hybrid Z_0 (see also Section 3.4.6.1):

$$f_{MHz} = \sqrt{\frac{12666}{L \times C}} \qquad \text{(Eq 11-22)}$$

where
L is expressed in μH
C is in pF
F is in MHz

Enter Z2 and Z3 in the *two-shunt-hybrid-comp.xls* worksheet which will calculate the required shunt elements. If the value of the shunt across Z3 is close (within 10%) to what you have measured as transformer shunt inductance, you will need to look only at the capacitor across Z3. If you are not too far off from the standard Four Square configuration, it is likely that you can get good results with just a parallel capacitor across Z2.

Run a modeling file that includes the hybrid coupler (eg *Ch11-4sq+hyb-3.65-commercial+1CAP.ez*) and enter all the elements of interest: element lengths (in the Wires window); hybrid C and L values on lines 5, 6 and 8 in the Load window; the shunt inductance of the transformer on line 3 of the Load window; and the calculated shunt capacitance on line 2 of the Load window. You can now calculate the performance across the band by running *EZNEC* model using its sweep function (eg in steps of 10 kHz).

If all of this is too difficult, you may just try various capacitor values across port 2. For 80 meters, try different values between 50 and 350 pF. If you are near where you want to be, the input SWR will be very flat across the band and measure less than 1.3:1 maximum at the band edges. You can also check the dump power ratio by measuring your input power to the hybrid and the power dumped in the dummy load. The ratio (dumped power/input power) should be at least 10 dB at the band edges. (Calculate dB from 10 log [dump power/input power]). Measuring directivity and F/B is much more difficult, but you should "hear" the difference.

There is a fair chance that by just putting a capacitor across port 2 (the port that leads to the center elements), you've turned your array a very good performing transmit antenna that works well as a receiving antenna too. Give it a try — you won't hurt anything by trying.

3.4.6.10. Using the Compensated Hybrid Feed System on Other Quadrature-Fed Arrays

In the previous sections we have covered in detail how we can improve the performance of the quadrature-fed Four Square array using a 90° hybrid coupler. It is obvious that these techniques can be used on other arrays where the elements are fed in a quadrature configuration.

3.4.6.10.1. Optimized Hybrid Coupler Fed 2-Element End-Fire Array (Quadrature Feeding)

Quadrature feeding a 2-element end-fire may not give the highest gain nor the best directivity, but it is worthwhile assessing the operational bandwidth of such an array. We know that the L-network feed method yields a bandwidth of approximately 80 kHz on 80 meters (see Section 3.4.5.4.5 and Fig 11-17).

Not Optimized

Run the modeling file *Ch11-2el-endf+hyb-80m-commercial.ez* without any shunt reactance optimization, but including the shunt inductance of the 180° phase reversal transformer (approximately 4.1 μH) that is part of commercial units such as the Comtek.

Just specify a standard frequency hybrid for 3.65 MHz (C = 436 pF, L = 2.18 μH). The result is very disappointing. With the radiating elements resonant at 3.65 MHz you will get a

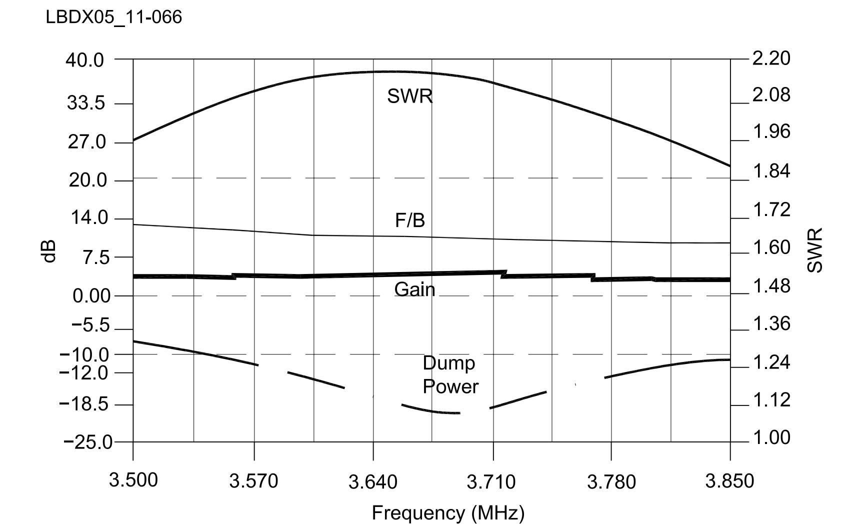


Fig 11-66 — Performance data for the quadrature fed 2-element end-fire array (quarter-wave spacing) using a commercial hybrid coupler. It has gain, but very poor F/B.

lousy F/B of between 9 and 12 dB, and that over approximately 300 kHz (see **Fig 11-66**).

Optimized

First model the antenna (file *Ch11-2el-endfire-90-90-incl-Z2Z3.ez*) to calculate Z2 (impedance front element at the end of current-forcing feed line) and Z3 (at the end of the back element). In this case 50-Ω feed lines give the most manageable impedances. Using equivalent ground loss resistance of 5 Ω in each element, these impedances are: Z2 = 36.7 – *j* 16.8 Ω and Z3 = 76.8 + *j* 35.1 Ω. Plug these values in the *two-shunt-hybrid-comp.xls* spreadsheet to calculate the required shunt elements: 4.23 µH across Z2 and 216 pF across Z3, with f_o = 5.045 MHz (L-hybrid = 1.58 µH and C-hybrid = 216 pF). See **Fig 11-67**.

Note that this application does not require a 180° transformer. If, however, you use a commercial unit, the transformer is there, and you will need to remove it! Contrary to what's the case with a Four Square, the inductive shunt will go across the Z2 arm (–90°), and the capacitive shunt across the Z3 arm (0°).

Next run the *Ch11-2el-endf+hyb-80m-commerc-2shunt comp.ez* modeling file in sweep mode to assess the operational bandwidth. **Fig 11-68** shows the reborn end-fire array which now exhibits a F/B of >25 dB over the entire 80 meter band. The operational bandwidth is almost 300 kHz. This is quite an improvement in bandwidth over the L-network feed system that yielded 80 kHz (see Fig 11-17).

Determining Z2 and Z3

It's one thing to design the array on paper, using *EZNEC* and the *two-shunt-hybrid-comp.xls* spreadsheet. It's another to make the array performance come true in the field. Instead of going only by the design data, apply the technique described in Sections 3.4.6.4.4 and 3.4.6.4.5 to calculate the exact values of Z2 and Z3. These values are derived from measurements on your 2-element array, and not just on a model. Knowing the correct values of Z2 and Z3 will allow you to insert the correct shunt elements for the optimizing system. As I say throughout this book, measuring is knowing.

Not Symmetrical?

It is obvious that this 2-element array is not a symmetrical array as is the case for the Four Square (see Section 3.4.6.4.1). This has no impact on using the above described technique which works as well on symmetrical as on nonsymmetrical arrays. All that is required is that the array can be reduced to a black box having only two terminals, Z2 and Z3.

Optimize the k-Factor?

We know that in case of a Four Square array, it is possible to improve the directivity of the array by reducing k (the ratio of V2 to V3) from 1.0 to approximately 0.8 or 0.85. This scheme does *not* work with the 2-element end-fire array. The only way to get rid of the large high angle (~60°) back lobe is to increase the phasing angle to approximately 105°. At this

	HYBRID COUPLER OPTIMIZATION TWO SHUNT COMPENSATION DESIGN SYSTEM -w1mk-			
	Enter data in yellow background cells			
	CALCULATING THE SHUNT ELEMENTS			
1	Enter fa (design freq) array →	3.650	MHz	
2	Enter k-value →	1		
		Real part	Imag part	
3	(-90 ° branch) Enter Z2 →	36.70	-16.80	Ω
4	Branch 2 shunt element →	96.97	Ω	
5	→	4.23	uH	
6	Z2' (at port 2) →	44.39	Ω	
		Real part	Imag part	
7	(0° branch) Enter Z3 →	76.80	35.10	Ω
8	Branch 3 shunt element →	-203.14	Ω	
9	→	214.66	pF	
10	Z3' (at port 3) →	92.84	Ω	
	HYBRID INPUT DATA			
11	Enter Zo Hybrid →	50.0	Ω	
	OUTPUT DATA			
12	Frequency corr. factor =	1.3821		
13	New Hybrid fo =	5.045	MHz	
14	L-Hybrid =	1.58	uH	
15	C-Hybrd =	316	pF	
16	Dump port (port 4) power ratio =	-18.3	dB	
17	Real part Zin (port 1) =	44.33	Ω	
18	Imag. part Zin (port1) =	12.88	Ω	
19	Port 1 return loss (dB) =	16.5	dB	
20	Port 1 SWR =	1.35		

Fig 11-67 — Phase compensated hybrid optimization system by W1MK applied to a 2-element end-fire array.

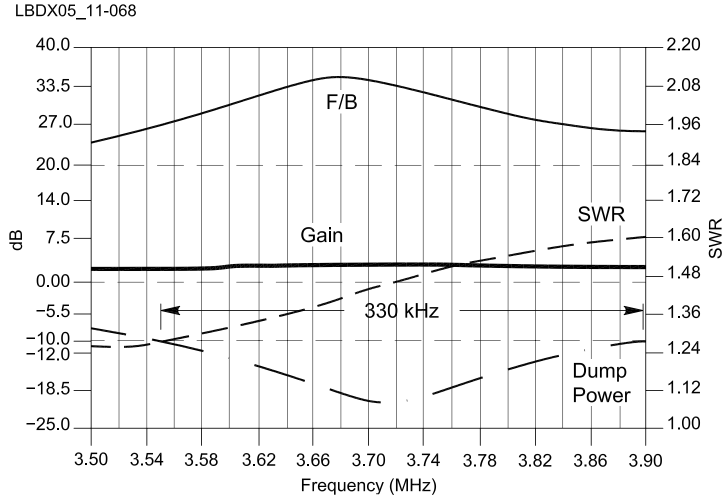


Fig 11-68 — The reborn end-fire array now using the two-shunt optimization system. Note that the hybrid f_0 also needs to be changed to 5.047 MHz.

phase angle the k ratio may change from approximately 0.9 to 1.1 without much influence on the directivity pattern. It is obvious that with θ = 105° we can no longer use the hybrid coupler feed system.

3.4.6.10.2. Optimized Hybrid Coupler Fed, Equally Spaced 3-Element In-Line End-Fire Array (Quadrature Feeding)

Fig 11-69 shows the black box configuration for the 3-in-a-line array. A model is available as file *Ch11-3el-endfire-spacing70deg-blackbox.ez*. Just as with the Four Square, we can prove that there is no mutual coupling between port 2 and 3 by feeding port 3, writing down the impedance at the feed point, and then terminating port 2 of the black box in different impedance values (open, short etc). It appears that different loads will not change the value of Z3. The same test can be done by feeding port 2 and terminating port 3, which will give the same result.

The array with elements spaced 70° is fed in quadrature with the center element fed with 1.85 times the current magnitude vs the current magnitude in the outer elements (k = 1.85). The 70° spacing is used so that one can use λ/4 feed lines (with foam dielectric cable, VF = 0.82) instead of 3⁄4 λ feed lines.

Running the modeling file gives us the following Z2 and Z3 impedances (using 50 Ω current-forcing feed lines): Z2 = 62.7 + *j* 0.9 Ω and Z3 = 25.6 – *j* 13.3 Ω. This antenna is ideally suited to be fed by one of the optimized hybrid coupler feed systems.

Alternative 1 (Single Shunt)

In this case I ran the *single-shunt-hybrid-comp.xls* spreadsheet with the Z2 and Z3 values using 50-Ω feed lines.

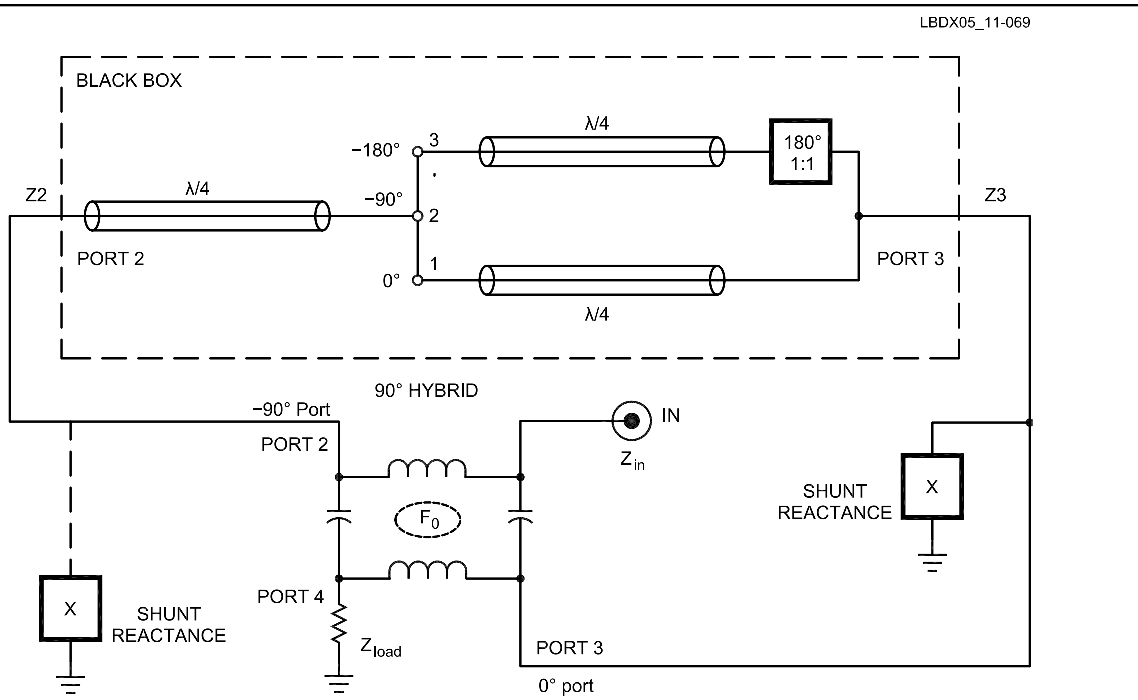


Fig 11-69 — The black box for the 3-in-line end-fire array with equal spacing and quadrature feeding.

module 1	SHUNT REACTANCE CALCULATOR for single shunt element hybrid optimization		
	Real	Imag	
ENTER the impedance at port 2 : Z2 (Ω) →	62.70	0.90	Ω
ENTER the impedance at port 3 : Z3 (Ω) →	25.60	-13.30	Ω
ENTER the hybrid design impedance Zo →	50	Ω	
ENTER array design frequency fa (MHz) →	3.650	MHz	
Magnitude voltage at port 2/port 3: k→	1.85		
OUTPUTS FOR SHUNT ELEMENT ACROSS PORT 3			
Reactance Shunt element	61.33	Ω	
value shunt element =	2.674	µH	
new Z3 =	32.51	0.34	Ω
Fo =	4.782	MHz	
module 2a	HYBRID COUPLER DESIGN (by W1MK)		
CASE 1: IMPORTED INPUT DATA FOR SHUNT ACROSS PORT 3			
(imported) fa (antenna frequency) =	3.650	MHz	
(imported) hybrid frequency →	4.782	MHz	
(imported) Zo Design impedance hybrid =	50.00	Ω	
	Real Part	Imag part	
(imported) Impedance load PORT 2 (R2) =	62.70	0.90	Ω
(imported) Impedance load PORT 3 (R3) =	32.51	0.34	Ω
RESULTS			
fo/fa =	1.310		
Hybrid L value (uH) =	1.66	µH	
Hybrid C value (pF) =	333	pF	
Ratio Voltage magnitude Port2/Port3 (k) =	1.850		
Phase angle Voltage port 2 vs. port 3 =	-90.00	°	
Power in Port 4 (vs. Pwr in Port 1) =	-25.9	dB	
Real part input impedance (port 1) =	52.08	Ω	
Imagainare part input impedance (port1) =	-15.85	Ω	
Return loss (port 1) =	-16.2	dB	
SWR =	1.37		

Fig 11-70 — Optimized hybrid coupler design for a 3-element in-line end-fire array. In this case we used the single-shunt reactance optimization method. See text for details.

Specify f_a = 3.65 MHz and k = 1.85 (the center elements need 1.85 times the current magnitude as compared to the front and back elements.) The program calculates two solutions, one with the shunt element across Z2 and another one with the shunt element across Z3. Z_0-hybrid = 50 Ω gives a port 4 dump power of –26 dB using a shunt element across Z3. This shunt element is a coil with an inductance of 2.674 µH (see **Fig 11-70**). The new hybrid design frequency is 4.782 MHz

Z_{in}-hybrid = 50.32 – *j* 15.59 Ω. We can tune out the reactive part with a series coil having an impedance of +15.59 Ω, which then results in an input SWR of 1:1 in a 50-Ω system impedance (the 50-Ω feed line impedance).

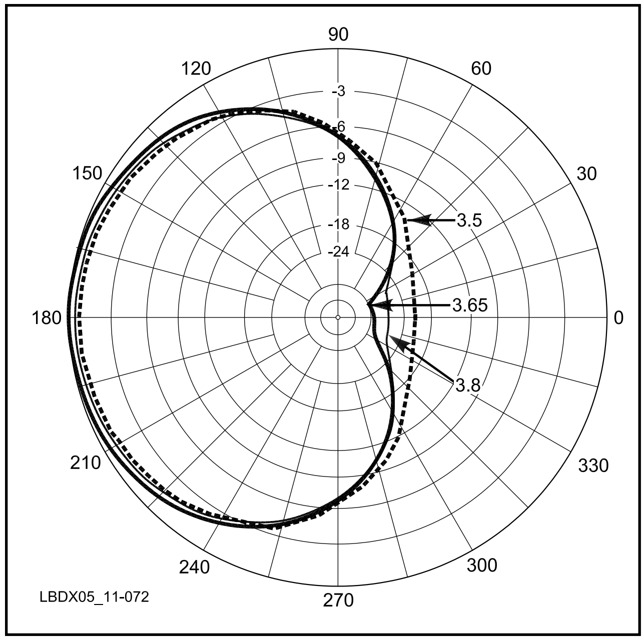


Fig 11-72 — Horizontal radiation pattern at mid-band and at the two band edges for the 3-in-line end-fire array with 70° spacing between the elements. (Plot generated with W8WWV's *LBDXView* software.)

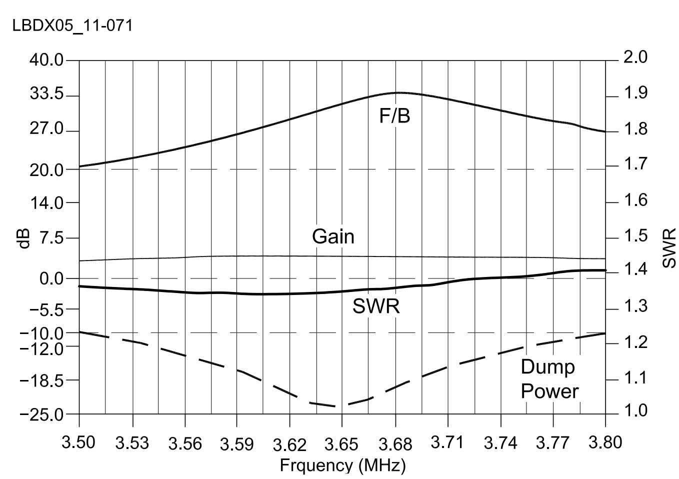
Alternative 2 (Two Shunts)

Run the *two-shunt-hybrid-comp.xls* spreadsheet. Specify k = 1.85 and use the abovementioned Z2 and Z3 values. The program calculates two shunt elements, one capacitor (10 pF, which means you can leave it off) to connect across Z2 and a shunt coil of 2.73 µH to be connected across Z3. The new values (real impedances) become: Z2' = 62.71 Ω and Z3' = 32.51 Ω.

Next determine Z_0-hybrid to the optimal value giving the lowest possible port 4 dump power:

$$Z_0 = \sqrt{Z2' \times Z3'} = \sqrt{38.41 \times 35.11} = 49\,\Omega$$

Fig 11-71 — Operational bandwidth data for the 3-in-line end-fire array with 70° spacing between elements and quadrature feeding with k = 1.85, fed by the single-shunt compensated hybrid coupler system across Z3. (Plot generated with W8WWV's *LBDXView* software.)



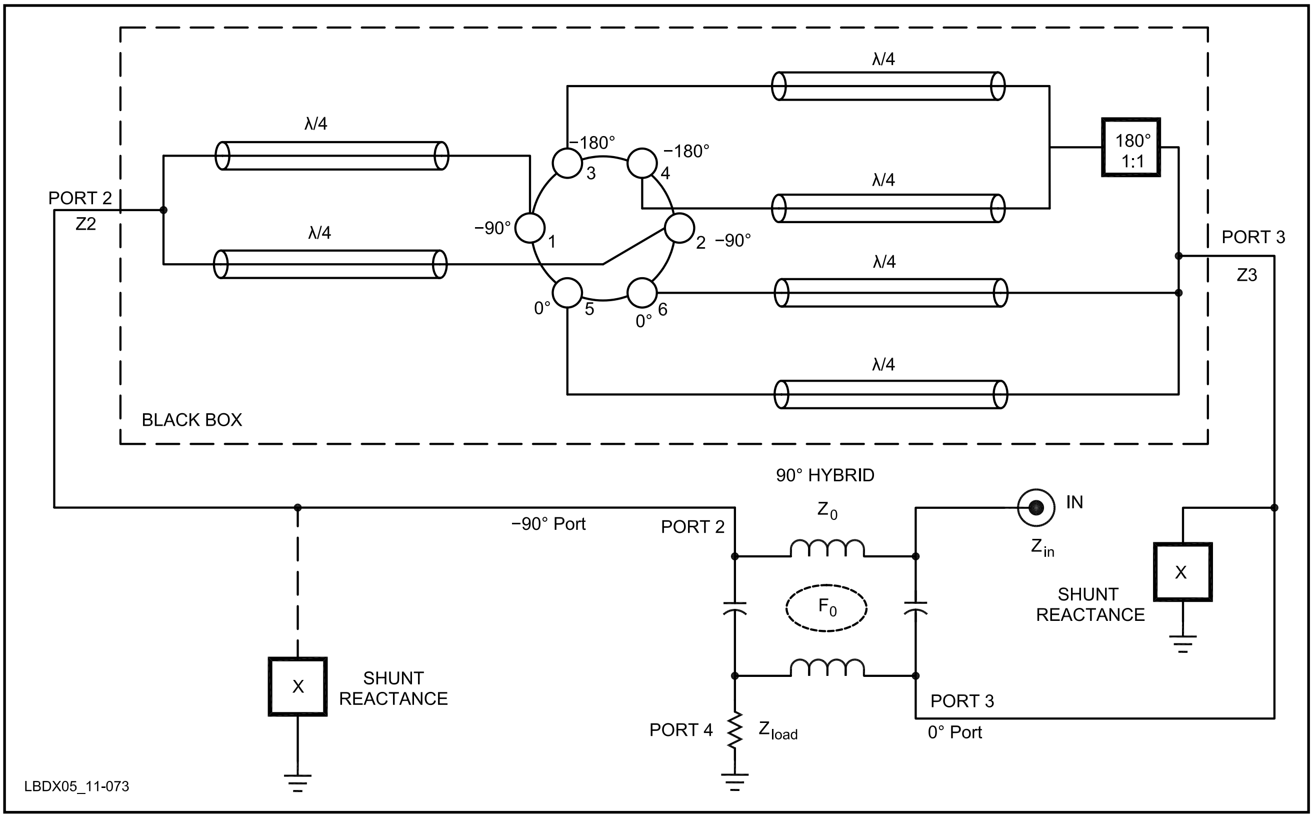


Fig 11-73 — The Six Square black box for the HEX-array, similar to the Four Square black box shown in Fig 11-41 and the 3-in-line black box circuit shown in Fig 11-69.

Let's use 50 Ω, as we did above. With this value of Z_0 we see a port 4 power which is about 26 dB down (same as above). Note that the new hybrid design frequency (f_0) is exactly the same as obtained with the single-shunt method (4.782 MHz). In fact this two-shunt system is a single-shunt system because the capacitor shunt has a negligible value of 10 pF (on 3.65 MHz) that can be ignored.

Lines 17 and 18 of the spreadsheet show the hybrid input impedance as 50.32 – *j* 15.59 Ω, the same as we found above. Note that the feed impedance found via calculation (using the spreadsheet) and via the *EZNEC* model differ slightly, but not significantly.

Measuring Z2 and Z3

As with any array that can be reduced to a black box with two ports, this array can best be built by applying the technique described in Sections 3.4.6.4.4 and 3.4.6.4.5 to calculate the exact values of Z2 and Z3 — values derived from measurements on your 2-element array and not just on a model. Knowing the correct values of Z2 and Z3 will allow you to insert the correct shunt elements for the optimizing system.

The Operational Bandwidth for the End-Fire 3-In-Line Array

Fig 11-71 shows the operational bandwidth analysis (see Sections 3.1.1. and 3.4.6.4.8) results for an array fed by the single-shunt reactance compensated hybrid network. **Fig 11-72** shows the horizontal radiation patterns at three frequencies (modeling file: *Ch11-3el-endfire+hyb-3.65-50ohm.ez*).

This array has a F/B of better than 25 dB over 200 kHz (on 80 meters), which is excellent. SWR is very flat (about 1.4:1) but can be reduced to 1.1:1 or better over the entire 300 kHz by inserting a coil with a reactance of approximately +15 Ω in series with the hybrid feed point.

3.4.6.10.3. Optimized Hybrid Coupler Fed, Fully Symmetrical Six Circle (HEX) Array (Quadrature Feeding)

Similar to what is the case with the Four Square and the symmetrical 3-in-line end-fire arrays covered above, the fully symmetrical HEX-array can be simplified to a black box containing the radiating elements, the phase reversal transformer and the current-forcing feed lines, showing only port 2 and port 3 to the outside world (**Fig 11-73**). This *EZNEC* model is available on the CD: *Ch11-6circle-HEX-small-quad-75ohm.ez*.

The quadrature fed small (60 meter diameter) version of the HEX array (see Section 4.11) can work with ¼ λ current-forcing feed lines (providing you use foam dielectric coax with VF ~ 0.82), while the larger HEX array requires ¾ λ long feed lines. For further detail see Section 4.11. When using 75-Ω current-forcing feed lines, the abovementioned modeling file (using k = 1.75) gives us Z2 = 57.6 + *j* 28.1 and Z3 = 20.2 – *j* 11.1 Ω.

Using W1MK's Two-Shunt (Phase Compensation) Hybrid Optimization System

Let's first work out the feed system according to the W1MK two-shunt phase compensation system (see Section 3.4.6.6) for f_a = 3.65 MHz. If we plug the abovementioned values of Z2 and Z3, and k = 1.75 into the *two-shunt-hybrid-*

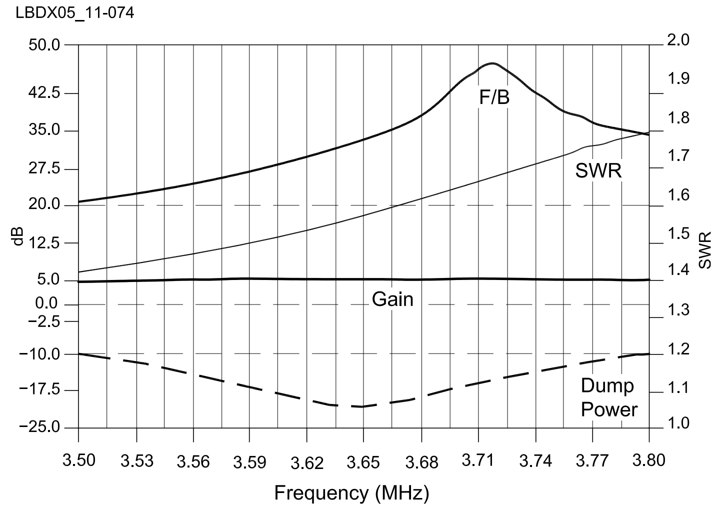


Fig 11-74 — Operational bandwidth data for the "small" (60 meter diameter) HEX- array. Similar bandwidth performance can be obtained when using the W1MK two-shunt compensation method. The SWR values listed are without series coil. With series coil the SWR ≤1.2:1 over the whole range. (Plot generated with W8WWV's *LBDXView* software.)

comp.xls spreadsheet, we calculate the required shunt element as 300 pF (on 3.65 MHz) in the –90° leg (port 2) and as 2.09 µH at port 3, resulting in Z2' = 68.71 Ω and Z3' = 26.8 Ω. Using Z_0-hybrid = 50 Ω, the spreadsheet calculates the new f_o as 3.745 MHz. The input impedance of the hybrid is 45 – *j* 23 Ω, which can be matched to 50 Ω with a series inductor of X_L = +23 Ω. The port 4 dump power is down 23.1 dB, which is excellent.

Using W1MK's Single-Shunt Reactance Optimization System

Using the same Z2 and Z3 values as above, and having specified k = 1.75, the *single-shunt-hybrid-comp.xls* spreadsheet calculates a shunt reactance of 1.341 µH across port 3 (the 0° port) of the hybrid (for Z_0-hybrid = 50 Ω). Z3 then becomes 24.53 + *j* 7.48 Ω (same value as found for the phase compensation method) with f_o = 3.745 MHz. The port 4 dump power is down 13.4 dB. The input impedance (port 1) is now 50.1 – *j* 24.4. To match this to our 50-Ω feed line we simply need a series coil with a reactance of +24 Ω.

Measuring Z2 and Z3

As described previously, apply the technique described in Sections 3.4.6.4.4 and 3.4.6.4.5 and to make measurements at the Z2 and Z3 ports so the exact values can be calculated. Knowing the correct values of Z2 and Z3 will allow you to insert the correct shunt elements for the optimizing system.

The Operational Bandwidth for the HEX Array

While the compensation systems were worked out (above) for a 160-meter array, the bandwidth assessment was done on 80 meters. **Figs 11-74** and **11-75** show the operational bandwidth analysis (see Section 3.1.1. and 3.4.6.4.9) results for the 30-meter diameter HEX-array fed by W1MK's two-shunt phase compensated hybrid feed system

This array has better than 27 dB F/B over 200 kHz (on 80 meters), which is outstanding (modeling file: *Ch11-HEX+hyb-3.65-k = 1.75-2sh-comp.ez*).

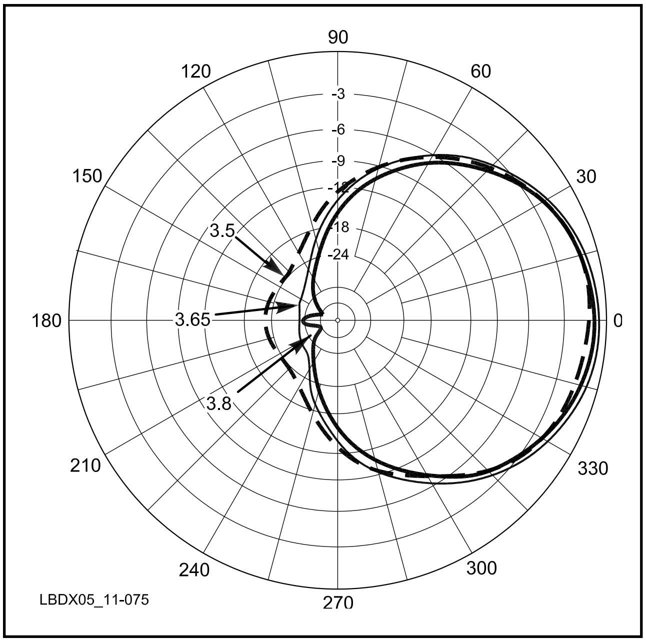


Fig 11-75 — Horizontal radiation patterns at mid band and at the two band edges for the HEX array. (Plot generated with W8WWV's *LBDXView* software.)

3.4.6.11. Bottom Line on Hybrid Coupler Feed Systems

For me too a Four Square driven by a hybrid coupler device was always something special. I always called it a "device" because I did not know the "whats, hows and how nots" of this mysterious thing.

That was until I asked my friend Robye, W1MK, "Why don't you write down all the mathematics concerning the 90° hybrid coupler." Had Robye been waiting to be asked? Anyhow, not more than a few days later, I received all the math formulas and the basic spreadsheet to do all kind of calculations (see *90 degree hybrid coupler design formulas.pdf* on the CD). Simply incredible. Now we can all understand how it works.

This is probably *the* highlight of the 5th edition of *Low Band DXing*. No more mystery. At last we know what makes the commercial units work (under most circumstances). And, what's even better, we now know what we have to do to make these things work even better — and that it is not even difficult to implement the optimization.

I have never read in any amateur or professional literature about these shunt equalization systems, or any other type of optimization systems. But the math made it clear why and how we can optimize the hybrid coupler feed system. Must we admire the guy who was the first to use the 90° hybrid with a shunt inductance across the 0° port, the shunt inductance being the self inductance of the 180° transformer, or was he just a lucky guy who hit the jackpot? If he was so smart, why did he never share his knowledge with his fellow hams? I, for one, love sharing my knowledge with other hams. Amateur Radio has been great to me; the least I can do is to do something for Amateur Radio in a return.

I am so happy I can, with the permission of Robye, W1MK, share our new knowledge on this subject with all of you, just to make it possible for our fellow hams to build better antennas and have better and more complete knowledge and understanding in this matter.

It also makes me happy to finally understand why my 80-meter Four Square with a Comtek coupler always worked well, and that now I know I can make it work even better. The message to those who have a Comtek or a DX Engineering hybrid coupler system, is: "You have a great system, and under most circumstances, if you use a classical Four Square configuration (λ/4 elements resonated on approximately 3.535 MHz, and erected on a square measuring λ/4 on the side, and being fed with 75 Ω foam current-forcing feed lines) you are close to what you can get out of it. But from reading this section of the book you understand that, in many cases, there is still room for some improvement."

When Robye, W1MK, came up with his quadrature-fed Four Square black box concept (Section 3.4.6.4.1), we put an important step forward in analyzing the behavior of the Four Square, and in the assessment of the symmetry of the array. Greg, W8WWV, came up with the terminology of *diagonal isolation* and was very helpful in evaluating different mechanisms influencing this isolation. In short, this whole demystification has been, to a large degree, a team effort.

The reason I wanted to dig in deeper in the mystery of the Four Square driven by a hybrid coupler was the fact that I knew it had to have better bandwidth than the L-network (W7EL) feed system, and that this better bandwidth was required, especially on 80 meters. For this Roy, W7EL, came to our rescue with a lumped constant model of the hybrid coupler that could be incorporated in any *EZNEC* modeling file. Finally we can now model our arrays including the hybrid coupler and do it swiftly on a range of frequencies to evaluate its operational bandwidth performance.

The thorough study of a "good" 180° (phase reversal) transformer, led us to question why the two commercially available models used a transformer with a very low shunt inductance (75 to 100 Ω). We developed one that has more like 500 Ω shunt inductance, which in a low-Z system (75 or 50 Ω) is just about negligible. The newly developed transformer (see Section 3.4.6.3.5) using type 61 material hardly loads the Z4 port, and it has the advantage of achieving a phase shift much closer to 180° than what is possible with the transformers used in the commercial units (>190°). It also has a much better a-factor (voltage magnitude transformation factor) which is important at the input of a quarter-wave current-forcing feed line to provide the right drive current magnitude to the array element at its end.

Further in-depth study of the hybrid coupler will undoubtedly lead to units that produce not just a 90° phase shift, but any phase shift. This will be ideal for driving end-fire arrays, which get much better performance if spacing wider than λ/4 is used with phase shifts in the 120°-130° ballpark.

The three optimizing methods described in this book, and the step-by-step procedure on how to modify a commercial unit (Section 3.4.6.8) should help somewhat technically oriented hams to experiment and improve their stations. The black box concept introduced by W1MK (see Fig 11-31) makes it possible to do easy on-site impedance measurements, as well as system adjustments. At the same time we have introduced, in the case of a Four Square fed by a hybrid, the facility to change the feed current ratio between branch 2 and 3. That makes it possible to reduce the big high-angle bulge in the back of the vertical radiation pattern, and to do that in a well controlled manner.

The two different shunt element optimizing methods developed by W1MK perform in a very similar way, although I must say that the two-shunt system is easier to implement as it does not require changing the resonant frequency of the radiating elements.

All mathematics and modeling were done assuming a 180° phase shift between the current in the front and the current in the back element. The commercial units now available on the market deviate by 10° to 12° from this figure (190° to 192°). Therefore the real life results, even when using one of the described optimization circuits, may be slightly different. If you are serious about building the best possible hybrid feed system and exploiting its intrinsic high performance bandwidth, take the trouble to build your own phase reversal transformer as described in Section 3.4.6.3.5. That device achieves an angle of just over 180° (183 to 185°).You can of course also use a high quality half-wave transmission line and switch in an extra length going from one end of the band to the other end.

Watch out and don't make the error of judging the operational bandwidth of the hybrid-coupler system by measuring the SWR curve at the input of the coupler. The coupler will show a very flat input SWR curve under almost *all* circumstances, over an excessively wide band. At frequencies more than approximately ±4% off the design frequency, more than 10% will be dissipated in the dummy resistor when using relatively low-Q elements (a diameter of 0.0008 λ or 6 cm minimum on 3.65 MHz).

In a high-Q array with the elements made of wire with a resonant radial system — one or just a few radials — it is not uncommon that, if the array is tuned for element resonance at 3.8 MHz, 50 to even 80% of the input power will be dissipated in the port 4 dummy load when operating at 3.5 MHz. The exact amount will depend on the Q factor of the elements. If you are using wire elements (high Q) you will have to divide the entire 80-meter band into several sections if you want to operate on both CW and phone. The trick is to readjust the elements for resonance when you switch from one section to another. Fig 11-162 later in this chapter shows how K9DX did this with his Nine Circle elements. The setup does two things (with one relay): change the resonant frequency of the antenna element, and change the length of the quarter-wave current-forcing feed line. My Four Square with wire elements and a single elevated radial requires the band to be divided in three sections, which keeps the dumped power below about 5% in each section.

It is clear that the most important bandwidth-determining parameter is the power wasted in the load resistor (port 4), and not input SWR (port 1). I have a wattmeter in the shack that measures the dumped power at all times, and sets off an alarm above a certain level.

Some hams however think that a dip in dumped power is a sign that their antenna is working, that's false too. I have seen designs where the dip in dump power does *not* coincide with the peak in F/B. The trick is to make sure they are both aligned.

To conclude, I think we have made a giant step forward in better understanding and improving the performance of our hybrid-driven quadrature-fed arrays.

On the CD that comes with this book the reader can find a *PowerPoint* presentation entitled "Demystifying the hybrid coupler" that has been used by the author to lecture on the subject (*Demystifying the hybrid coupler.ppt*).

3.4.7. Gehrke Method

Forrest Gehrke, K2BT, developed a technique that is fairly standard in the broadcast world. The elements of the array are fed with randomly selected lengths of feed line, and the required feed currents at each element are obtained by the insertion of discrete component (lumped-constant) networks in the feed system. He makes use of L networks and constant-impedance T or pi phasing networks. The detailed description of this procedure is given in Ref 924.

The Gehrke method consists of selecting equal lengths (not necessarily 90° lengths) for the feed lines running from the elements to a common point where the array switching and matching are done. With this method, the length of the feed lines can be chosen by the designer to suit any physical requirements of the particular installation. The cables should be long enough to reach a common point, such as the middle of the triangle in the case of a triangle-shaped array.

As this method is rarely used in amateur circles it is not covered in detail in this edition of the book (but it was covered in editions 1 through 3). This method however has the tremendous merit that it was the first one described in amateur literature that was technically 100% correct.

3.4.8. Lahlum/Gehrke Method

The Lahlum/Lewallen method described in Section 3.4.5 can be applied with feed lines measuring any length (not necessarily multiples of λ/4). I call this the Lahlum/Gehrke system, as it uses the mathematics developed by Robye Lahlum, W1MK, and follows more or less the principle of Gehrke's original methods, where arbitrary lengths of feed lines were used to the elements.

While the use of current forcing is a very desirable feature, there are situations where one might not care to use current forcing. For example, consider an array used on multiple bands, using the same coax feed for both bands. One example is an array covering 80 meters with wide spacing (approximately λ/4) and 160 meters with close (λ/8) spacing. The Lahlum/Lewallen method must be used in this situation.

Fig 11-76 shows the basic setup for a 2-element array. In this case we will first have to calculate the impedances at the end of the feed lines, for example using the "Coax Transformer/Smith Chart" module of the *New Low Band Software*. The formulas involved are given in Fig 11-13 where R and

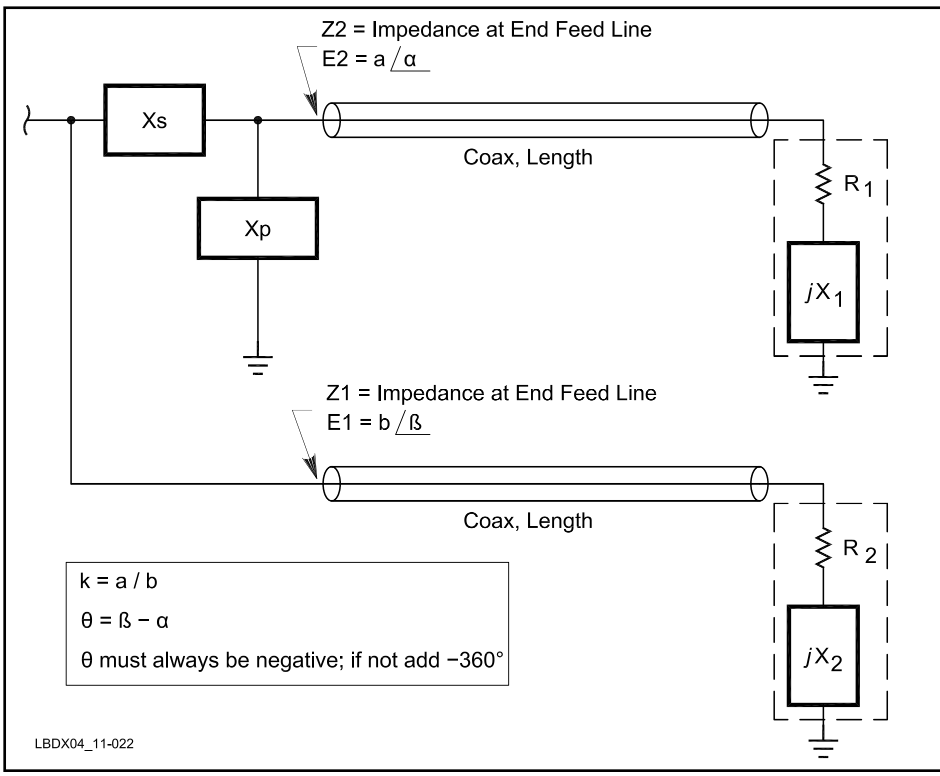


Fig 11-76 — Basic set up for a 2-element array. Starting from the feed impedances of the elements we calculate Z and E at the end of the two feed lines. The schematic clearly shows the requirements for the network: it must transform the voltage a α° to b β° with a phase shift of θ = (β°– α°), while k = a/b See text for details.

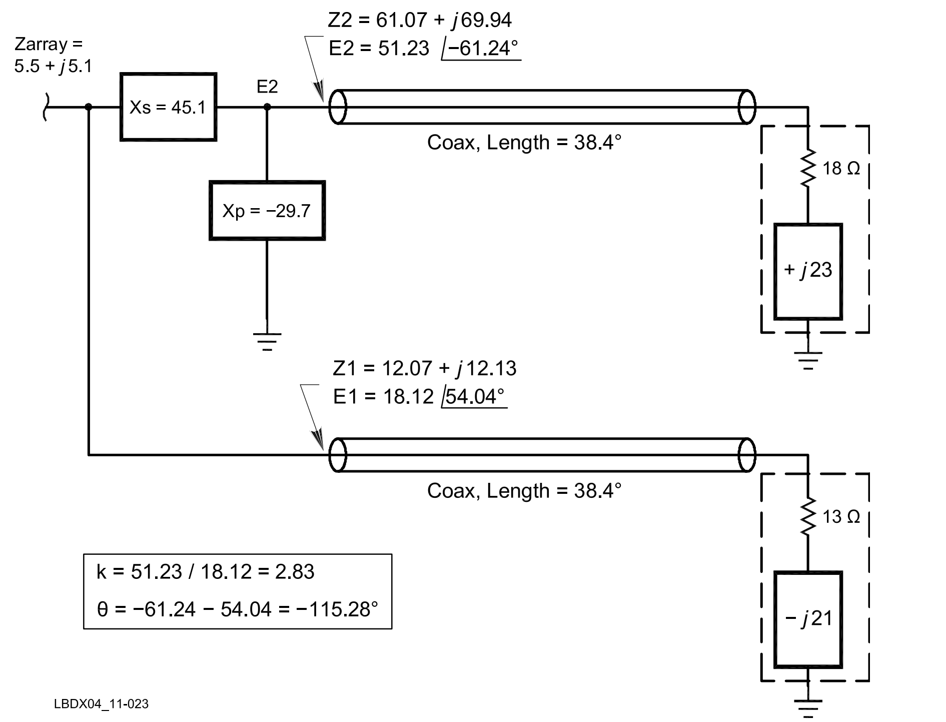


Fig 11-77 — First solution for a 2-element end-fire array (λ/8 spacing). See text for details.

j X represent the transformed impedance values of the feed impedances of the antenna elements, via the coaxial feed line. Fig 11-13 also shows the equations for X_s and X_p.

In Section 3.4.5.4 we used k (in the Lahlum spreadsheet) as the ratio of the feed currents when we use current-forcing feed lines. In this application however, k is equal to E1/E2, the ratio between the *voltages* at the end of the feed lines. The feed lines are not necessarily 90° long or odd multiples thereof, and do not even have to be of equal length.

θ is again the phasing caused by the L-network. It is the phase angle difference between the voltages at the end of the two equal-length feed lines. More precisely it is the difference between the voltage phase angle at the output of the L-network and the phase angle at the input of the network. θ *must* be negative. If necessary subtract 360° to obtain a negative value. Fig 11-76 shows the principle.

Let's work out an example for a 2-element, ⅛ wavelength spacing case where the phase shift is –135°. Through antenna modeling we obtained the following element impedance values:

Back element:
I_{back} = 1 A ∠0°
Z_{back} = 13 – *j* 21 Ω
Front element:
I_{front} = 1 A ∠–135°
Z_{front} = 18 + *j* 23 Ω

Using the "Coax Transformer/Smith Chart" module of the *New Low Band Software*, the values at the end of a 38.4° long feed line are calculated. (Note: it is *not* necessary that both feed lines be of equal length, unless of course you want to switch directions.)

I used a frequency of 1.83 MHz, using real cable (RG-213, 0.2 dB loss/100 ft). We now need to look at the voltage at the end of the feed lines as we need to connect them in parallel (equal voltages required!). The transformed values are (see also **Fig 11-77**):

At end of feed line to back element:
E1 = 18.12 ∠54.04°
Z1 = 12.07 + *j* 12.13 Ω
At end of feed line to front element:
E2 = 51.23 ∠–61.24°
Z2 = 61.07 – *j* 69.94 Ω

We need to insert an L-network in the feed line to either the front or to the back element. This L-network has to perform the followings two tasks:

- Perform the required phase shift.
- Perform the required voltage transformation so that the input voltage to the L-network is identical to the voltage at the end of the other feed line (so that we can connect them in parallel).

Solution 1: We put the L-network in the feed line going to the front element.

θ is the difference between the voltage phase angle at the output of the L-network and the phase angle at the input of the network. θ *must* be negative. If necessary subtract 360° to obtain a negative value.

θ = (–61.24) – (54.04) = –115.28°

k is the ratio of the voltage magnitudes at the end of the feed lines.

k = 51.23/18.12 = 2.83

We can use these values in the spreadsheet *Lahlum-Lnetwork.xls* available on the CD that comes with this book. This tool allows you to calculate the values of the L-network.

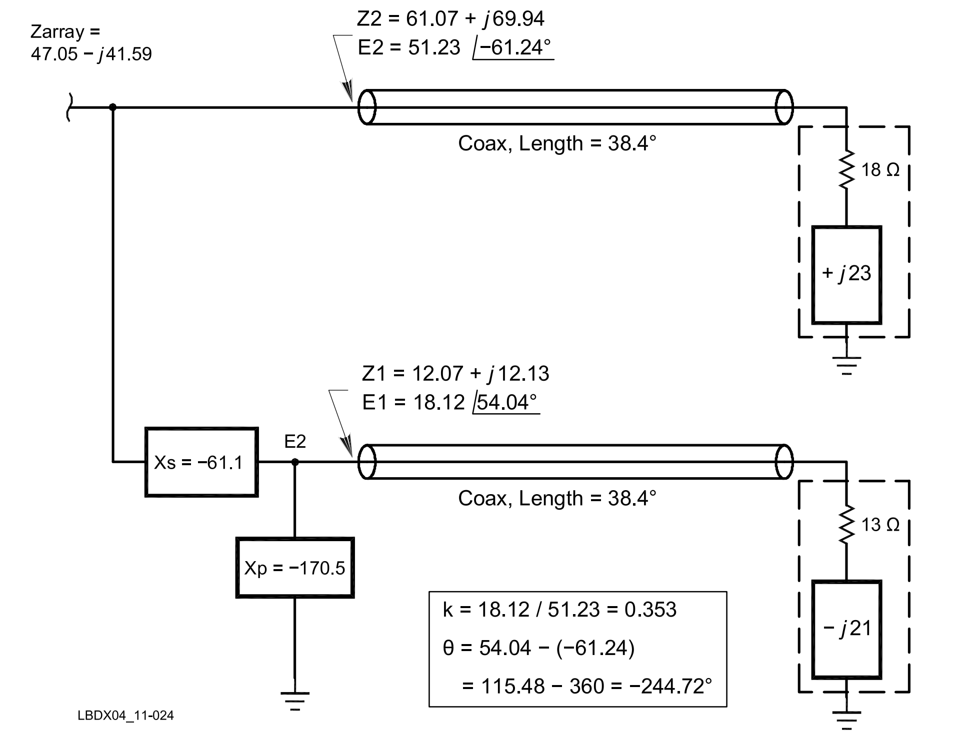


Fig 11-78 — Second (preferred) solution for a 2-element end-fire array (λ/8 spacing). See text for details.

Table 11-13
Values for Solution 1

INPUT DATA		
Enter R →	61.07	Ω
Enter X →	69.89	Ω
Enter k →	2.830	
Enter θ →	-115.28	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	45.1	Ω
X-Par =	-29.7	Ω
Series elem =	1.9	uH
Par elem =	1410.0	pF
Rpar =	17.61	Ω
Xpar =	20.41	Ω
Rser =	10.09	Ω
Xser =	8.71	Ω

Table 11-14
Values for Solution 2

INPUT DATA		
Enter R →	12.07	Ω
Enter X →	12.13	Ω
Enter k →	0.353	
Enter θ →	-244.72	°
Enter freq →	3.80	MHz
RESULTS		
X-Series =	-62.1	Ω
X-Par =	-170.5	Ω
Series elem =	674.3	pF
Par elem =	245.8	pF
Rpar =	194.69	Ω
Xpar =	-54.00	Ω
Rser =	13.91	Ω
Xser =	-50.14	Ω

For this example, X_{series} = 45.11 Ω and X_{par} = – 29.7 Ω. See **Table 11-13**.

Solution 2: The L-network is located in the feed line going to the back element (**Fig 11-78**).

$$\theta = (54.04) - (-61.24) = +115.28 = (-360 + 115.28) = -244.72°$$

$$k = 18.12/51.23 = 0.353$$

For this solution the spreadsheet program calculates X_{series} = – 62.1 Ω and X_{par} = –170.5 Ω. See **Table 11-14**.

Which is the best solution? Let's look at the input impedance of the feed system:

For Solution 1, – *j* 29.7 Ω in parallel with 61.07 + *j* 69.94 Ω gives 10.07 – *j* 36.34 Ω. Adding the series reactance of +*j* 45.11 Ω gives 10.07 + *j* 8.76 Ω (also available on the last two lines from the spreadsheet). Paralleling this impedance with 12.07 + *j* 12.13 Ω gives 5.5 + *j* 5.1 Ω, the array feed impedance (Fig 11-77).

For Solution 2, – *j* 170.5 Ω in parallel with 12.07 +

j 12.13 Ω gives 13.91 + j 12.0 Ω. Adding the series reactance of – j 262.2 Ω gives 13.91 – j 50.14 Ω (also available on the last two lines from the spreadsheet). Paralleling this impedance with 61.07 + j 69.94 Ω gives: 47.04 – j 41.59 Ω, the array feed impedance (see Fig 11-78).

Both solutions are valid; the only difference is the resulting input impedance. In Solution 1 the resulting input impedance is very low (5.5 + j 5.1 Ω). Solution 2 yields an array feed impedance much closer to 50 Ω (47 – j 41 Ω), and the use of a series inductor would give an almost perfect match.

This approach to solving the problem of obtaining the correct amplitude and phase shift using coax feeds of any length is similar to the Gehrke method, but it results in much fewer circuit elements. Solving this same above problem using the Gehrke method would result in the need for six or seven elements (see *Low Band DXing*, editions 1, 2 and 3), all of which would affect the current/phase relationships.

Using the Lahlum/Lewallen approach, four elements would be required. Two of them would be an L-network matching the array input impedance to the feed line impedance, and only two of them affect the current/phase relationship. Thus it is much easier to adjust.

3.4.8.1. Adjusting the Network Values

If you do *not* use current forcing (feed lines that are λ/4 or odd multiples thereof), you cannot use the testing and adjustment procedure as described in Section 3.6.1 (measuring voltages at the end of the current-forcing feed lines). In this case you will have to use a small current probe at the elements (see Section 3.4.9.5 and Fig 11-89 later in this chapter).

3.4.8.2 Other Applications of the Software

While the calculation procedures described in Sections 3.4.5 and 3.4.6 assume current-forcing feed lines without losses, you can use the above procedure to take the losses into account. For that you first need to calculate the impedances at the end of the current-forcing feed lines, using the "Coax Transformer/Smith Chart" module of the *New Low Band Software* (option "with cable losses") and then use these values as input date for the bottom part of the spreadsheet. This is also explained in more detail in Section 3.4.5.4.4.

3.4.9. The Opposite Voltage Feed System (by OH1TV)

Have a look at **Fig 11-79**. This is the schematic of the *opposite voltage feed system* as developed by Pekka Ketonen, OH1TV.

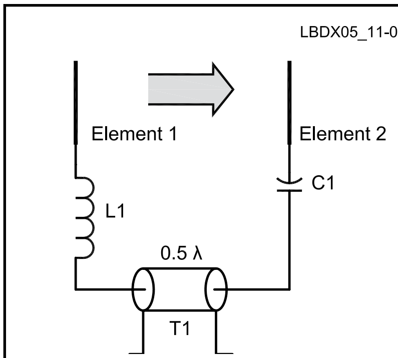


Fig 11-79 — Complete opposite voltage feed system.

So that the reader gets used to this new and never (until now) published system, let's quote some key properties:

- Both elements are voltage driven, in opposite phase (hence the expression "voltage forcing"). From the feed point of element 1 ("driven element") there is a λ/2 feed line to element 2 (called the "reflector").
- One could also feed the elements from a common point with two individual half-wave long feed lines, in which case a phase reversal transformer (or an extra half-wave feed line) would be required at either end of one feed line.
- No 180° phase reversal is required for changing direction.
- The voltage (not the current) at element 2 (the reflector) is lagging 180° out of phase with the voltage at element 1 (the driven element).
- As we all know, what is finally important is the current (magnitude and phase) in both elements. It is the current that determines the radiation.
- The current at the base of the elements depends on the "detuning" of the elements. These currents are not in phase with the voltage. The tuning is done with a loading element that is most often a coil (at the back element, the reflector) and a capacitor (at the front element, the driven element), which are both installed at the base of the element.

3.4.9.1 Calculating the Values of the Loading Elements, Method A

To understand how the array fed according to the opposite voltage system works, we have to take you through some definitions and some basic mathematics. Robye, W1MK, developed two different approaches to calculate the value of the loading elements j A and j B.

3.4.9.1.1. The Mathematics

See **Fig 11-80**. To start with, each element in an array has its own individual characteristics (self or "naked" impedance):
Z11 = self impedance of element 1
Z22 = self impedance of element 2

It also has a characteristic that defines its behavior in the group of elements. The presence of other elements in the group changes the impedance of the element. This impedance is what we call the coupled impedance. Z1,2 = Z2,1 = the impedance of one element 1 with the other element grounded (assuming identical elements). See Section 3.3.1 for more details.

From Z11, Z22, Z1,2 and Z2,1 we can calculate the mutual impedance Z12 and Z21 using the *w1mk-on4un-oh1tv-arrays.xls* spreadsheet available on the CD (sheet *MutZ, DriveZ*).

As elements of an array, it is the current (magnitude and phase) at the base of the elements that will determine the radiation pattern of the array:

I1 = 1 ∠θ° — antenna base current in element 1 (the reference element with 1 A current at 0°).

I2 = k ∠θ° — antenna base current in element 2 (k A magnitude and θ° phase shift).

(θ is the phase indicator and k the magnitude indicator.)

V1 and V2 are the *voltages* at the feed points of the two elements. As these are connected through a λ/2 long feed line, they are 180° out of phase, and we can write:

$$V1 = -V2 \qquad \text{(Eq 11-23)}$$

The voltages at the base of the elements are:

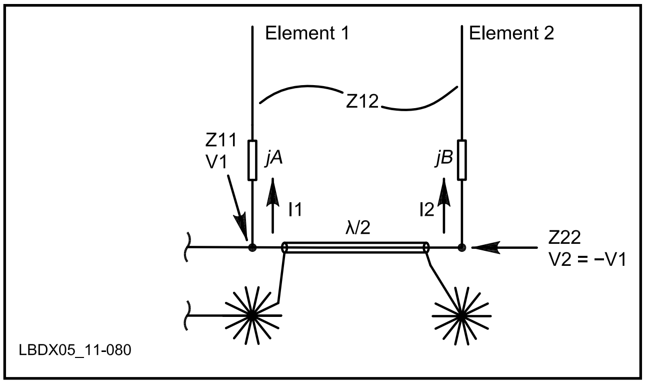


Fig 11-80 — Schematic representation of the equal voltage feed method. The λ/2 feed line between the feed points of the two elements ensures that the voltages at these points are equal in magnitude and 180° out of phase (V2 = –V1).

$$V1 = [(jA + Z11) \times I1] + (Z12 \times I2)$$

and

$$V2 = -V1 = [(jB + Z22) \times I2] + (Z12 \times I1)$$

From this we can calculate:

$$B = \frac{R1 + R2 \times k \times \cos\theta}{k \times \sin\theta} - X2 \qquad \text{(Eq 11-24)}$$

and

$$A = -X1 - (X2 \times k \times \cos\theta) - (R1 \times k \times \sin\theta) - (B \times k \times \cos\theta) \qquad \text{(Eq 11-25)}$$

where $R1 + j\,X1 = Z11 + Z12$ and $R2 + j\,X2 = Z22 + Z12$.

Note that for the simple case of two identical verticals in a perfectly symmetrical world, and for θ = 90° (quadrature), we have R = R1 = R2, X = X1 = X2, k = 1 and θ = 90°, which yield the following values for A and B:

$$A = -(X + R) \text{ and } B = R - X \qquad \text{(Eq 11-26)}$$

In practice this additional reactance (j A and j B) can be achieved by either shortening the driven element and lengthening the "reflector" element, or (more practical) by inserting a coil or capacitor at the base of the elements.

3.4.9.1.2. Calculating the Loading Devices: A Simple Spreadsheet Program

The mathematics involved in following the math steps as explained above are "complex" and somewhat tedious to do by hand (even with a calculator). This is why I developed a spreadsheet to help us with that task. The *w1mk-on4un-oppvolt.xls* spreadsheet on this book's CD will help you calculate the values of the loading elements. The first method to do this is the "*Mutual Impedance Method*" is shown in a red frame.

Let's work out the example of a 2-element end-fire array.

Step 1: Measuring the Elements

We first need to measure the impedance of each element (on its own, uncoupled from the other element) and next when it is mutually coupled to the second element. *Uncoupled* means that we leave the vertical element floating (not connected to ground). *Coupled* means that the vertical element is shorted to the ground system.

For the 2-element 80-meter array, with two identical elements, and using a ground system that is characterized by a 5-Ω equivalent ground loss resistance, the elements are trimmed in length to be resonant on exactly 3.65 MHz. The measured impedances are Z11 = Z22 = 5 + 36 = 41 + j 0 Ω. We also need to measure the coupled impedances (see Section 3.3.1.2): Z1,2 = Z2,1 = 37.5 + j 15.2 Ω.

Step 2: Calculating the Mutual Impsedances

Starting from the self impedance and the coupled impedances we can calculate the mutual impedance using the *w1mk-on4un-oh1tv-arrays.xls* spreadsheet available on the CD (see Section 3.3.2.). Use the calculator "Calculating Mutual Impedance" shown in a green frame. **Fig 11-81** shows a screen shot of the calculator, where Z12 (= Z21) was calculated as 19.79 – j 15.75 Ω.

One can also use a modeling program to "calculate" the mutual impedance, as explained in Section 3.3.2.

Step 3: Calculating the Loading Elements

Now switch to the spreadsheet *w1mk-on4un-oppvolt.xls.* The input data for this spreadsheet are Z11, Z22, Z12, k, θ and f_o (see **Fig 11-82**). The values must be entered into the cells with a yellow background and framed with a fat border. The

CALCULATING MUTUAL IMPEDANCE			
ACTION	real	imag	
ENTER Z11	41	0	self-Z EL1
ENTER Z22	41	0	self-Z EL2
ENTER Z1,2 = Z2,1	37.5	15.2	coupled Z
RESULT: Z12 =	19.79	-15.75	= mutual Z

Fig 11-81 — Spreadsheet for calculating the mutual impedance between two elements of an array.

action		real part	imag part
ENTER Z11 (EL1) →	in Ω	41	0
ENTER Z12 →	in Ω	19.79	-15.75
SUM Z11 + Z12	R1=	60.79	-15.75
ENTER Z22 (EL2) →	in Ω	41	0
Sum Z22+ Z21	R2=	60.79	-15.75
ENTER k →		1	
ENTER θ →	in °	-105	
	θ in rads	-1.8326	
front element (EL1)		-30.9	Ω
back element (EL2)		62.4	Ω
ENTER frequency	Fo =	3.65	MHz
front element (EL1)		1411.33	pF
back element (EL2)		2.72	uH

Fig 11-82 — Spreadsheet for calculating the values of the loading elements *j*A and *j*B. This method is called the "mutual impedance" method because it requires the user to enter the mutual impedance into the spreadsheet.

other cells are protected to prevent accidentally overwriting code.

If the two elements in the array are not identical (Z11 ≠ Z22), we must run the spreadsheet a second time, for the array shooting in the opposite direction. This makes it possible to optimize such an array for each direction, which is difficult to do with any other feed system.

3.4.9.2. Calculating the Values of the Loading Elements Via the "Black Box" Method

This method is different from the one described in Section 3.4.9.1 and has the distinct advantage that it can be used also for arrays with more than two elements.

3.4.9.2.1. The Mathematics

Section 3.3.2 explained in great detail self impedance and coupled impedance and how mutual impedance can be calculated from the measured values of self and coupled impedance using the "calculating mutual impedance" section of the *w1mk-on4un-oh1tv.xls* spreadsheet software.

We will use the "calculating array drive impedances" section of the *w1mk-on4un-oh1tv.xls* spreadsheet to calculate the drive impedance of the two elements.

The following example is for two quarter-wave element end-fire array with λ/4 element spacing:

- We measured Z11 = Z11 = 41 Ω (36 Ω R_{rad} and 5 Ω R_{loss}) and Z1,2 = Z2,1 = 37.5 + *j* 15.2 Ω
- Use the section called "calculating mutual impedance" to calculate the mutual Z12 as being 19.79 – *j* 15.75 Ω (see Fig 11-81).

Using the calculator called "calculating array drive impedances" we calculate the drive impedances Z_{in1} and Z_{in2}:

- We specify the currents as 1 ∠0° and 1 ∠–105°.
- The program calculates Z_{in1} = 20.66 – *j* 15.04 Ω and Z_{in2} = 51.09 + *j* 23.19 Ω.

The calculation of the input impedances is based on: Z_{in1} = Z11 + Z12 × I2/I1 + Z13 × I3/I1 etc (see Section 3.3.2).

What does this mean? If we have two quarter-wave elements, spaced λ/4, and if these elements show drive impedances that are equal to the values Z_{in1} and Z_{in2} respectively (as calculated above), those two elements will have element drive currents (I1 and I2) as specified above (being currents of equal magnitude and 105° phase shift).

This is where the black box comes into the picture. We can think of the array as being a black box with two connections, one to element 1 and one to element 2 (see **Fig 11-83**). The black box is characterized by Z_{in1}, Z_{in2}, I1 and I2, which we know (we calculated the impedances and specified the currents).

As we feed the black box with equal voltages, we can convert the black box shown at the top of Fig 11-83 to another box, by simply applying Ohm's law.

As the black boxes are fed from the same source with a λ/2 feed line between them (which results in 180° phase shift), we can write that V2 = –V1. The voltage at the left port is

$$V1 = \frac{I1}{Z_{in1} + j\,A}$$

and at the right port:

$$-V1 = \frac{I2}{Z_{in2} + j\,B}$$

From this we can easily calculate the ratio I2/I1:

$$\frac{I2}{I1} = \frac{-(Z_{in1} + j\,A)}{(Z_{in2} + j\,B)} = \frac{-(R1 + j\,X1 + j\,A)}{(R2 + j\,X2 + j\,B)} \qquad \text{(Eq 11-27)}$$

From which we can calculate:

$$B = \frac{R1 + R2 \times k \times \cos\theta}{k \times \sin\theta} - X2 \qquad \text{(Eq 11-28)}$$

and

$$A = -X1 - (X2 \times k \times \cos\theta) - (R1 \times k \times \sin\theta) - (B \times k \times \cos\theta) \qquad \text{(Eq 11-29)}$$

Example

From the example above we know that Z_{in1} = 51.09 + *j* 23.11 Ω and Z_{in2} = 20.66 – *j* 15.04 Ω. We have specified I1 = 1∠0° and I2 = 1∠–105°, or k = 1 and θ = 105. If we plug these values into Eq 11-28 and 11-29 we obtain:
A = –30.9 Ω (capacitive reactance)
B = +62.42 Ω (inductive reactance)

Calculating the array feed impedance becomes very simple — it is just the value obtained by paralleling the impedances of (Z_{in1} + *j* A) and (Z_{in2} + *j* B):

$$Z_{array} = (Z_{in1} + j\,A) \times \frac{Z_{in2} + j\,B}{Z_{in1} + Z_{in2} + j\,A + j\,B} \qquad \text{(Eq 11-30)}$$

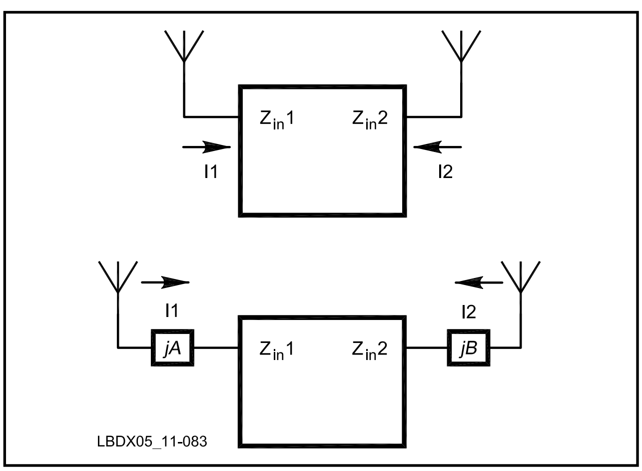


Fig 11-83 — The equivalent circuit of the "black box" method. See text for details.

3.4.9.2.2. Calculating the Loading Devices Using a Simple Spreadsheet Program (Black Box Method)

The spreadsheet *w1mk-on4un-oppvolt.xls* for this method is shown in a blue frame. Let's work out the same example of our 2-element end-fire array. See **Fig 11-84**.

Step 1. Measuring the Elements

Self or naked impedance: Z11 = Z22 = (41 + *j* 0) Ω

Coupled impedance, measured as: Z1,2 = Z2,1 = 19.75 – *j* 15.75 Ω

ACTION	real	imag	what?
ENTER Z11 (EL1)	41	0	self-Z EL1
ENTER Z22 (EL2)	41	0	self-Z EL2
ENTER Z12 = Z21	19.79	-15.75	Mutual Z12
	magn	angle	
ENTER I1 (mag, °)	1	0	curr el # 1
mag, rad	1.00	0	
ENTER I2 (mag, °)	1	-105	curr el # 2
mag rad	1.00	-1.83	
DRIVE IMPEDANCES	Real	Imag	
Zin EL1 =	20.66	-15.04	Ω
Zin EL2 =	51.09	23.19	Ω

Fig 11-84 — The 2-element drive impedance spreadsheet calculates the drive impedances at the elements, given the self impedances, the mutual impedance and the current magnitude and phase relationship between the elements. EL1 is the back element, EL2 the front element.

Step 2. Calculating the Mutual Impedances and the Element Feed Impedances

See Fig 11-84. For k = 1 and θ = 105° we obtain: Z_{in1} = 20.66 – *j* 15.04 Ω and Z_{in2} = 51.09 + *j* 23.19 Ω.

Step 3. Calculating the Loading Elements

We can now plug these values into the spreadsheet (calculator for Black Box method, framed in blue) and obtain the values of the loading elements *j* A and *j* B, as shown in **Fig 11-85**.

Note that the spreadsheet according to the black box method yields the same results for *j* A and *j* B as obtained using the "mutual impedance" method (Fig 11-82).The feed point impedance of the array is 28.5 + *j* 15.75 Ω.

Remark: Instead of starting from measured self and coupled impedances, we could simply have modeled the array using a model that specifies the element feed currents, such as *Ch11-2el-endfire-90-105.ez*. That model calculates the impedances at the base of the element as Z1 = 17.72 – *j* 14.65 Ω and Z2 = 47.53 + *j* 22.42 Ω, close to what we calculated in Figs 11-84 and 11-85 but not identical.

Using these impedance values we can use the black box spreadsheet program to calculate the loading elements: 1556 pF at element 2 and 2.58 µH at element 1.

METHOD B: THE BLACK BOX METHOD				
action		real part	imag part	
ENTER Zin1 (EL1) →	in Ω	20.66	-15.04	
ENTER Zin2 (El2) →	in Ω	51.09	23.19	
ENTER k →		1		
ENTER θ →	in °	-105		
	θ in rads =	-1.8326		
	Load EL2=	-30.9	Ω	
	Load EL1=	62.4	Ω	
ENTER frequency →	Fo =	3.65	MHz	
	Load EL2=	1411.63	pF	
	Load EL1=	2.72	uH	
		real part	imag part	
Zarray	Z =	28.50	15.75	Ω

Fig 11-85 — The black box spreadsheet tool to calculate the loading elements and the array feed impedance.

Modeling the Opposite Voltage Fed Array

The *EZNEC* modeling program makes it possible to include transmission lines in a model. This means that once you have developed an array using the method described above, you can plug in the data for the element loading coil and loading capacitor, as well as the half-wave transmission line and see what comes out.

In Sections 3.4.9.1 and 3.4.9.2 we calculated the values of the loading elements (for k = 1 and θ = –105°). Plug these values into the *EZNEC* model (file *Ch11-2el-endfire-0-105-volt-fed.ez*). *EZNEC* calculates the pattern shown in **Fig 11-86** which is really the kind of vertical pattern we were looking for.

3.4.9.4. Operational Bandwidth of the 2-Element End-Fire Four Square Opposite Voltage Feed System

Figs 11-87 and **11-88** show the operational characteristics and the radiation patterns of a 2-element end-fire array, λ/4 spacing and θ = 90°, designed for f = 3.65 MHz and fed according the opposite voltage feed system. To assess the SWR bandwidth, the model includes an L-network (L-series = 0.29 µH and C-shunt = 503 pF) at the input. This provides a perfect 50-Ω input impedance at 3.65 MHz. Note that the SWR remains very flat over the entire range. The operational bandwidth is about 50% higher than with the Lewallen L-network method (see Section 3.4.5.4.5).

If we want to make a 2-element end-fire array that covers 3.5 to 3.8 MHz with excellent directivity, we will need to switch the loading devices. We will need a set of devices for 3.5 MHz, another one for mid band (3.65 MHz) and a third one for 3.8 MHz. If you are only interested in the high end and the low end of the band, two sets will do.

3.4.9.5. Measuring and Tuning

We all know that the radiation pattern of an array is determined by the *current* in each of the elements. Arrays that use a *current-forcing* feed system (feed lines that are λ/4 or uneven multiples of λ/4 long) make it easy on us. In that case, we can measure the voltage at the end of the feed lines, voltage which equals the current (at the antenna) multiplied by the feed line impedance.

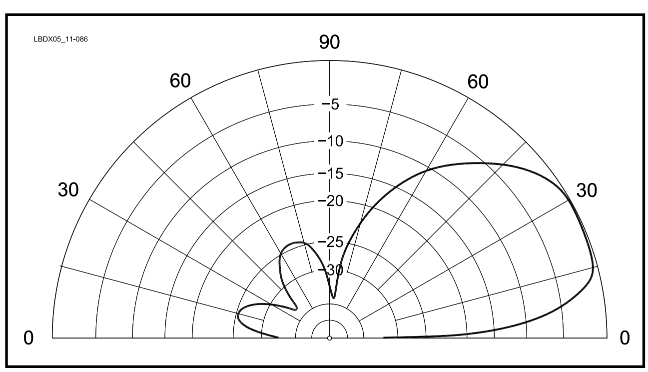


Fig 11-86 — Vertical radiation pattern for the array using loading element values as calculated with the spreadsheet program without any optimizing.

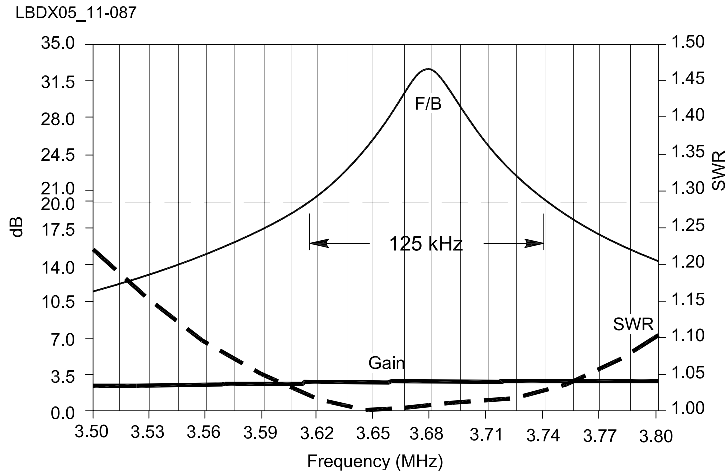


Fig 11-87 — The 20 dB F/B bandwidth of this 2-element end-fire array (λ/4 spacing and θ = 90°) is 125 kHz. Modeling file: *CH11-2el-endfire-0-90-volt-fed.xls*. (Plot generated with W8WWV's *LBDXView* software.)

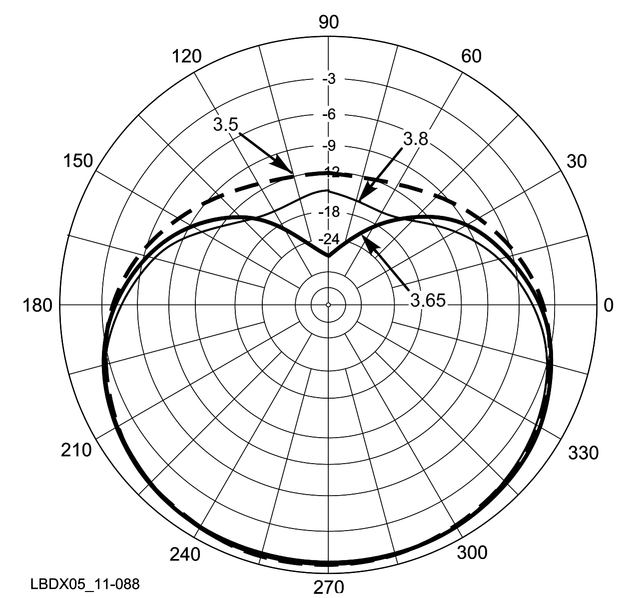


Fig 11-88 — Horizontal radiation patterns at mid-band and at the two band edges for the voltage-fed 2-element end-fire array (k = 1, θ = 90°). (Plot generated with W8WWV's *LBDXView* software.)

Table 11-15
Pi Type Attenuator Design
Values for Z = 50 Ω for k-ratios between 1.05 and 2

ratio (k)	*dB*	*Rser (Ω)*	*Rpar (Ω)*
1.05	0.43	2.5	2020
1.10	0.83	4.8	1050
1.15	1.21	6.9	720
1.20	1.58	9.1	551
1.30	2.28	12.8	396
1.40	2.92	17.1	300
1.50	3.52	21.0	250
1.66	4.40	26.4	202
1.75	4.86	29.5	183
2.00	6.02	37.5	150

In systems where we use no current-forcing feed lines, we need to measure the element current directly (at each of the elements).

The top-of-the-line test setup for adjusting an array is undoubtedly a VNA together with a vector scope (Section 3.6.2). W1MK's measuring setup (Section 3.6.1), which uses a 90° hybrid, is also very suitable for adjusting two signals for 90° phase shift and equal magnitude. **Fig 11-89** shows such a hybrid coupler setup for use with any array that does not employ current forcing (λ/4 feed lines). In this case we need to measure the antenna element currents at the elements using a *current probe* (see Section 3.5.5.2).

3.4.9.5.1. The Attenuators

Table 11-15 lists the resistor values for making pi-type attenuator (Z = 50 Ω) to be used in the measuring setup in cases where k ≠ 1.

Once the extra length of cable and the attenuator are installed (if necessary), one needs only to adjust the loading components (typically a coil and a capacitor) while watching the output of the power meter on remote readouts that can be installed at the base of the antenna elements.

3.4.9.6. Direction Switching

While you might be tempted to switch the loading devices (*j* A and *j* B) as shown in **Fig 11-90A**, the feed line is λ/2 long so you can put the loading element at either end of the line (see equivalent circuits in Figs 11-90B and 11-90C). The final arrangement with the direction-switching relay shows that it requires half the number of loading devices and a single DPDT relay instead of two DPDT relays to do the job (Fig 11-90D). In addition the switching is all done at one physical location.

3.4.9.7. The Opposite Voltage Feed System for Other 2-Element Arrays

The opposite voltage feed system can obviously also be used on horizontally polarized arrays, such as those described in Chapter 12.

I worked out an example of a 2-element 40-meter antenna on an 8-meter long boom, using shortened elements 16

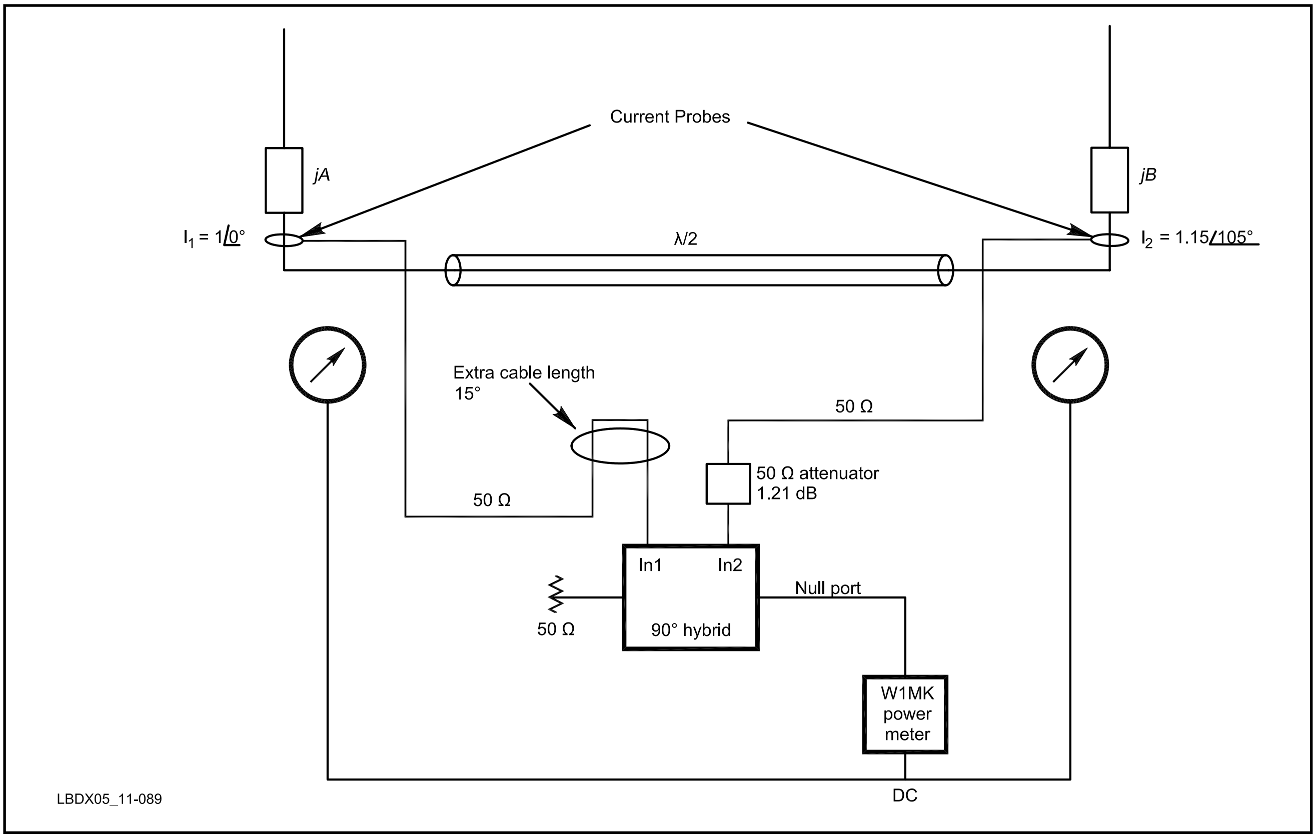


Fig 11-89 — Example test setup for an array with k = 1.15 and θ = 105°. An attenuator of 1.21 dB (20 log 1.15) is inserted in the measuring line going to the element with the highest current. An extra 15° long measuring line ensures that the signals picked up by the current probes are 90° out of phase at the hybrid for θ = 105°.

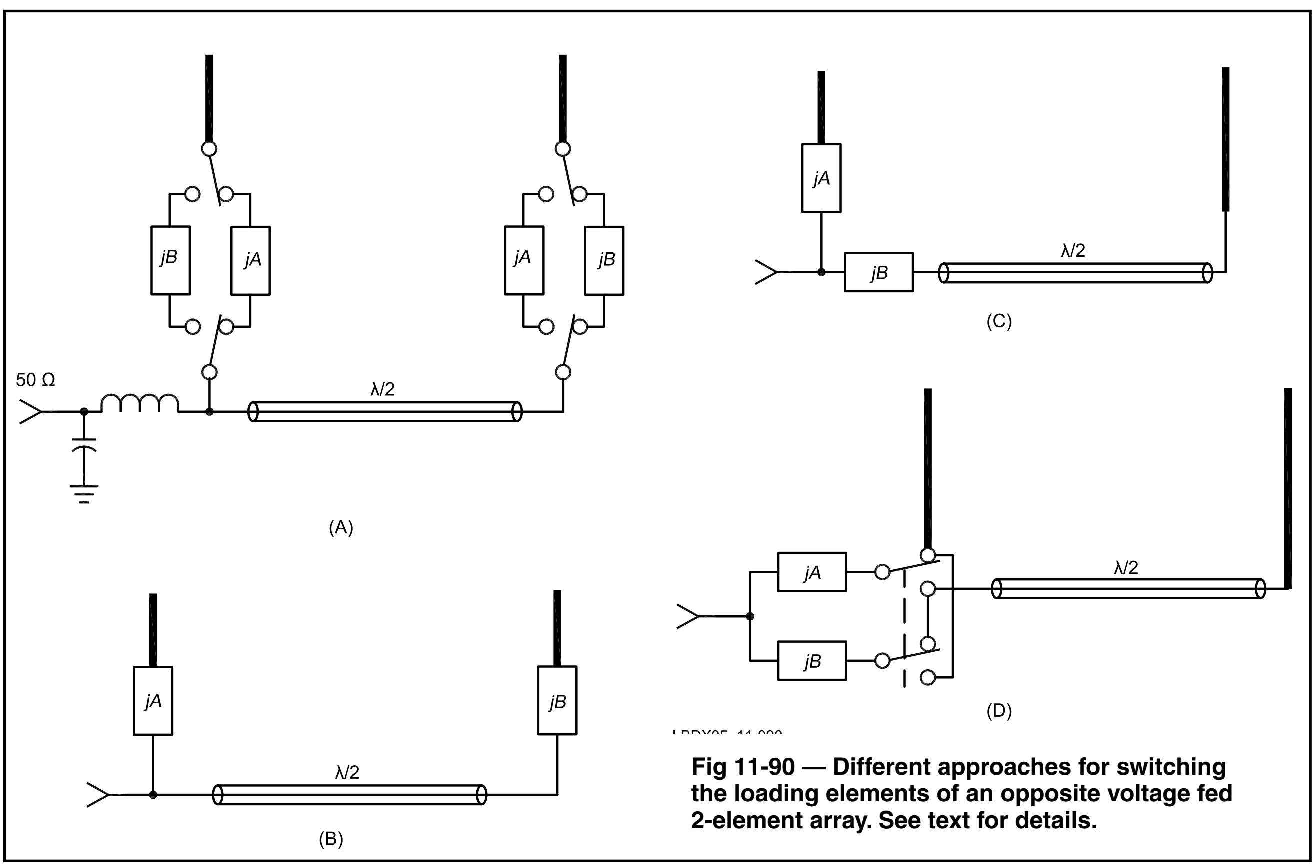


Fig 11-90 — Different approaches for switching the loading elements of an opposite voltage fed 2-element array. See text for details.

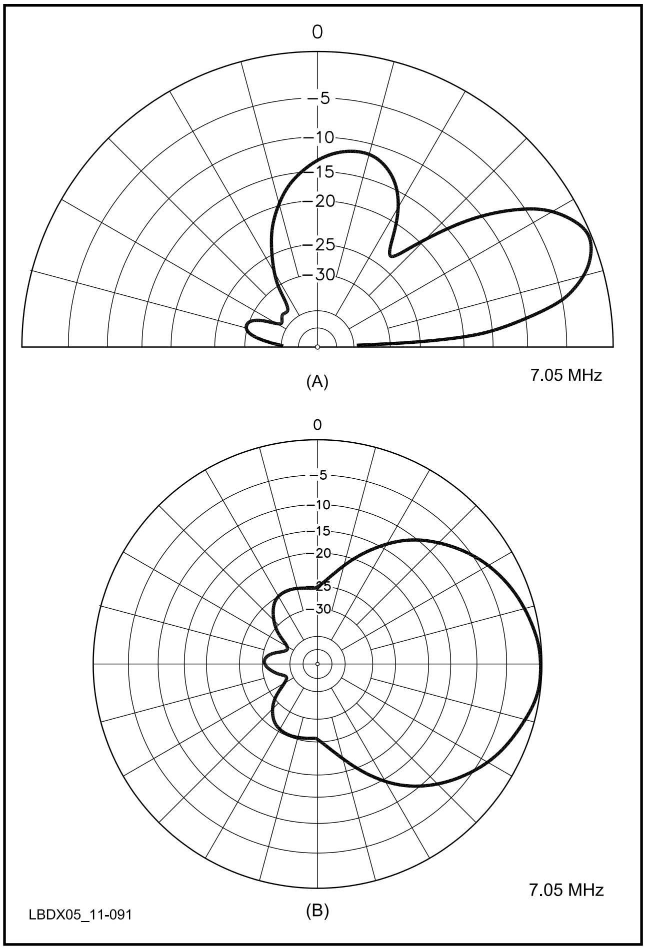


Fig 11-91 — Radiation patterns for the 2-element array at 20 meters height. The F/B over a wide angle in the back is between 20 and 30 dB, which is a value than can never be obtained with a 2-element Yagi. In addition the SWR is less than 1.3:1 over more than 100 kHz.

meters long (70% of full size). Pekka, OH1TV, has modeled a number of such 2-element arrays for the different HF bands, and has worked out examples where a single physical design could be successfully used on two bands. It would certainly be possible to model a 7 and 10 MHz array. But let us stick to a single-band 40-meter design.

If you model this antenna in free space, you can get some startling radiation patterns. However, real life is not free space. After having made an initial free-space design using *EZNEC*, I further optimized the antenna at a real height of 20 meters ($\lambda/2$), and optimized the loading device values for best directivity and good SWR bandwidth between 7.0 and 7.1 MHz. See *EZNEC* file *Ch11-2el-40m-short-array-voltage fed-at20m.ez*. The directivity on 7.05 MHz is shown in **Fig 11-91**.

Gain is 11.5 dBi (calculated over average ground) and F/B is better than 20 dB over 90° in the horizontal plane and between 0 and 50° in the vertical plane.

The feed point impedance is $14 + j\ 11\ \Omega$ at 7.05 MHz. (SWR 1:1 after matching). The SWR curve is short of spectacular: 7.0 MHz: 1.2:1, 7.1 MHz: 1.3:1. As the elements are substantially shorter than $\lambda/2$, the loading components in both elements are coils (two 2.45 µH coils and two 3.1 µH coils respectively).

By playing around with the values of the loading elements (eg 2.3 µH and 3.1 µH) you can achieve a model that can be used over a wider bandwidth (a lower-Q antenna). The change results in a slightly higher feed impedance (approximately $18 + j\ 11\ \Omega$) and a wider SWR bandwidth that is typically 1.5:1 at 7.0 MHz, 1:1 at 7.1 MHz and 1.5:1 at 7.2 MHz. The trade-off for such design is that the directivity is not as good.

The half-wave transmission line connecting the two "dipoles" should be equipped at both ends with a common-mode choke (balun). As the separation between the elements is only 8 meters, and a half wave of coax such as RG-213 at 7.05 MHz measures approximately 14 meters, it is a good idea to use a 14-meter length of RG-213 inside the boom of the Yagi. Coil up approximately 3 meters at each end of the boom, where the two

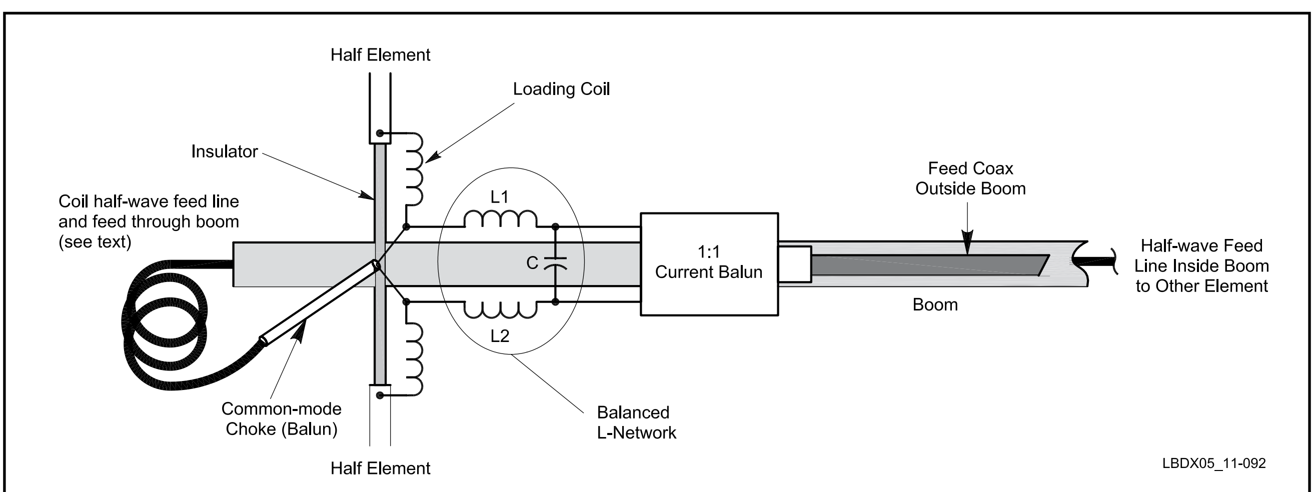


Fig 11-92 — Top view of the feed arrangement for the 2-element array fed by the opposite voltage feed system. All coils and capacitors should be mounted inside a watertight box. See text for further details.

feed lines are attached to the elements. The inductive reactance of these coils is too small however (<500 Ω) to make a "good" balun, unless you pass the coil turns through four large-diameter ferrite cores evenly spaced on the coil.

Another solution is to use a stack of ferrites on a short piece of Teflon cable (see Chapter 6, Fig 6.18B).The Wireman (**www.thewireman.com**) sells the cores and the Teflon coax as a kit. The choke can be made using beads, or a length of RG-303 Teflon coax can be wound on a properly dimensioned ferrite core (eg a stack of two FT-240-61 cores as used for the 180° transformer from Section 3.4.6.3.5). Make sure you include the length of transmission line equipped with ferrite beads or wound on a ferrite core when calculating the physical length of the half-wave line. The feed system is shown in **Fig 11-92**.

Matching to the 50-Ω feed line can be done via an L-network. Watch out: The feed point is symmetrical, which means that you need to use a balanced L-network. Design an unbalanced L-network (using the "L Network" module of the *New Low Band Software*) for an impedance that is half the impedance you need to match into a 25-Ω load (½ of the 50-Ω line impedance). L1, L2 and C are the L-network components, where L1 = L2. The balanced L-network must be followed by a transmitting-type current balun to achieve an unbalanced feed point for the coax going to the station. The common-mode choke, as shown in that figure is part of the half-wave transmission line connecting the two elements.

3.4.9.8. More Elements

Let's have a look at the popular Four Square array. Here too we have a few options. We could develop a feed system based on a configuration where all four elements are connected to a loop made of four λ/2 long feed lines. A second approach is to use a star feed configuration where we run four λ/2 feed lines from a common point to the base of the four elements where we would place the required loading elements. The third and most attractive solution is the one where we apply the same star feed line principle, but have the loading elements at the opposite end of the feed lines, which means at a common location. That means we require substantially fewer loading elements. In addition it is nice to have all these components in the same place, which makes it much easier to work on. See **Fig 11-93**.

Robye, W1MK, developed the mathematics and the spreadsheet that make designing such a feed system child's play. The *4sq-voltagefeed-calculator.xls* spreadsheet makes it possible to calculate the loading devices A, B, A' and B' for any Four Square array (measuring λ/4 on the side) going from classic quadrature (k1 = k2 = 1 and θ1 = θ2 = 90°) to an array with odd k and θ values but in a practical range (eg k1 = 0.9, k2 = 0.85, θ1 = –120° and θ2 = –240°).

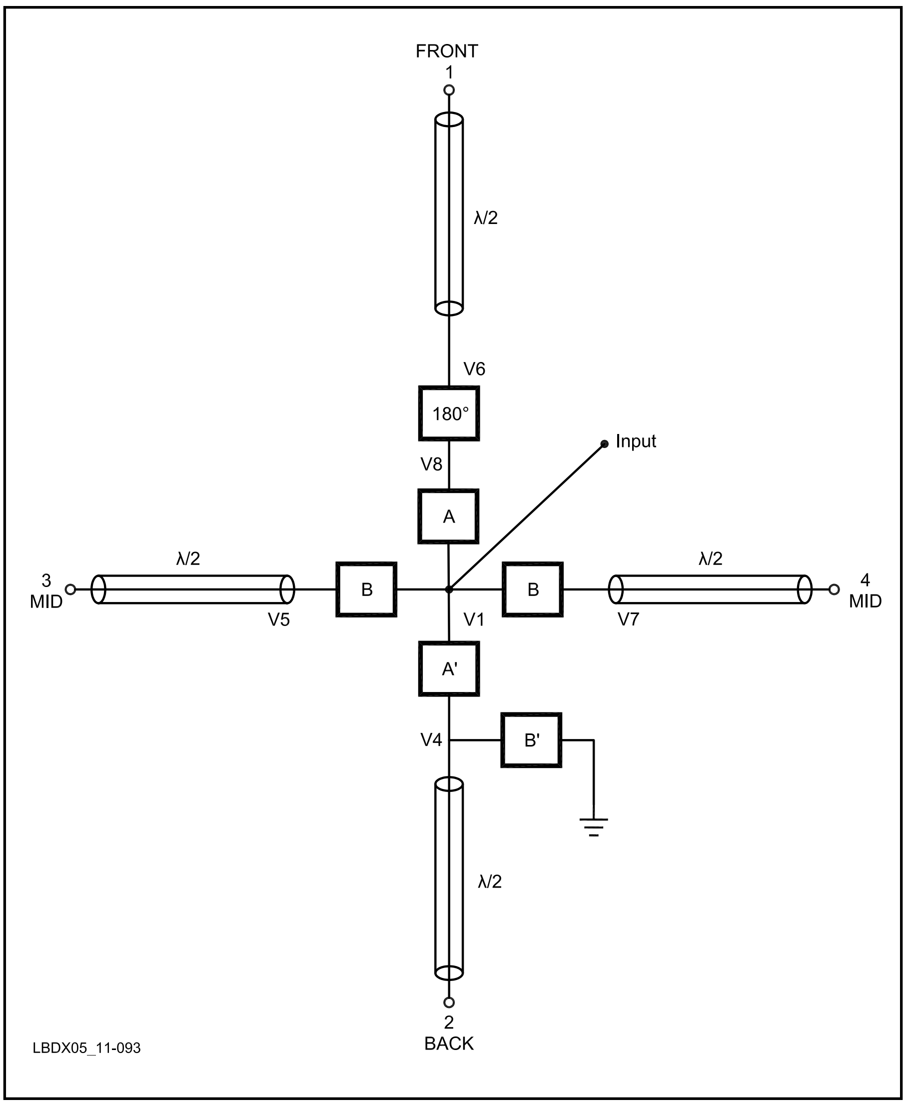


Fig 11-93 — Four Square star-shaped feed configuration for the opposite voltage feed system. The denominations used in this drawing (element number, connections V1 through V7) correspond with those used in the modeling files mentioned in the text.

The detailed procedure for how to use the spreadsheet program is given in the spreadsheet program itself. A generic *EZNEC* modeling file for this purpose is available on this book's CD (*Ch11-4sq-voltage-feed.ez*) as well as a few examples (*Ch11-4sq-volt-90-180.ez, Ch11-4sq-volt-0.85_-90-1_-180.ez,* and *Ch11-4sq-volt-WA3FET.ez*). See also Section 4.7.6.

3.4.9.9. Operational Bandwidth of the Four Square Opposite Voltage Feed System

The *EZNEC* models of an array fed with the opposite voltage feed system include all elements (cables, coils, capacitors), so operational bandwidth assessment is made easy. **Figs 11-94** and **11-95** show the operational characteristics and the radiation patterns of a Four Square array (quadrature fed),

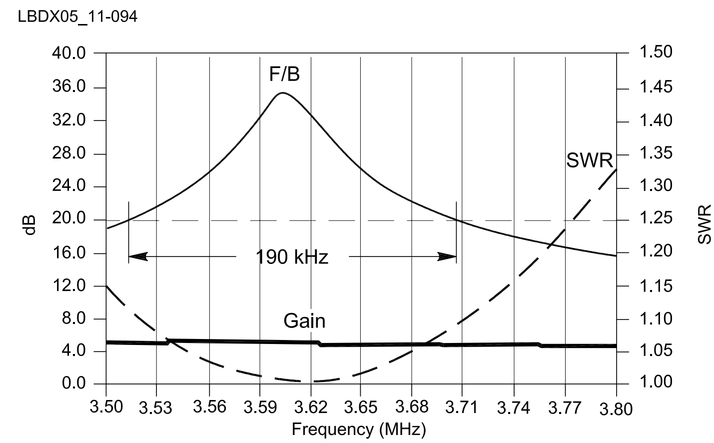


Fig 11-94 — The 20 dB F/B bandwidth of this Four Square array (k = 1, θ = 90°) is approximately 190 kHz. Modeling file: *CH11-4sq-volt-90-180.ez.* (Plot generated with W8WWV's *LBDXView* software.)

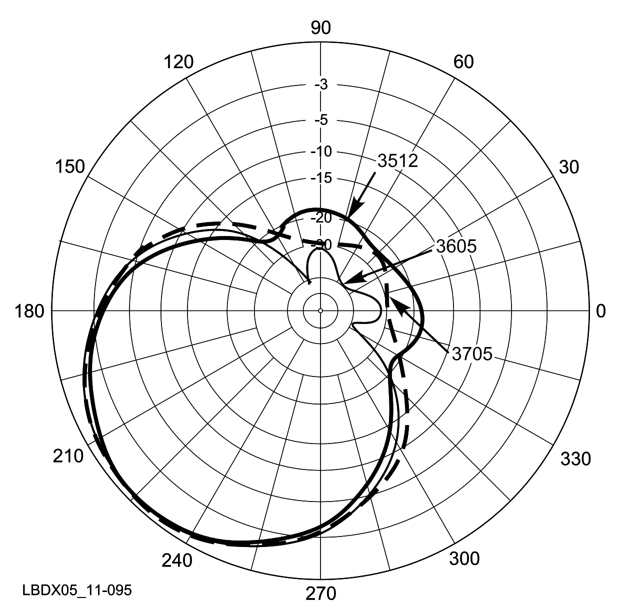


Fig 11-95 — Horizontal radiation patterns for the voltage fed Four Square (k = 1, = 90°). The plots were made at the limits where the F/B was 20 dB and mid-way in between those frequencies. (Plot generated with W8WWV's *LBDXView* software.)

designed for f = 3.65 MHz fed by the same feed system. The model (*Ch11-4sq-volt-90-180.ez*) also includes an L-network to match the input impedance to 50 Ω at 3.65 MHz: L-series = 0.81 µH and C-shunt = 1600 pF.

To compare, the same array using the Lewallen L-network feed systems has a 20 dB F/B bandwidth of ~80 kHz. With one of the compensated hybrid coupler feed systems though, we reach 300 kHz.

3.4.10. Choosing a Feed System for Your Array

Until Gehrke published his excellent series on vertical arrays, it was general practice to simply use feed lines as phasing lines, and to equate electrical line length to phase delay under all circumstances. We now know that there are better ways of accomplishing the same goal (see Section 3.3).

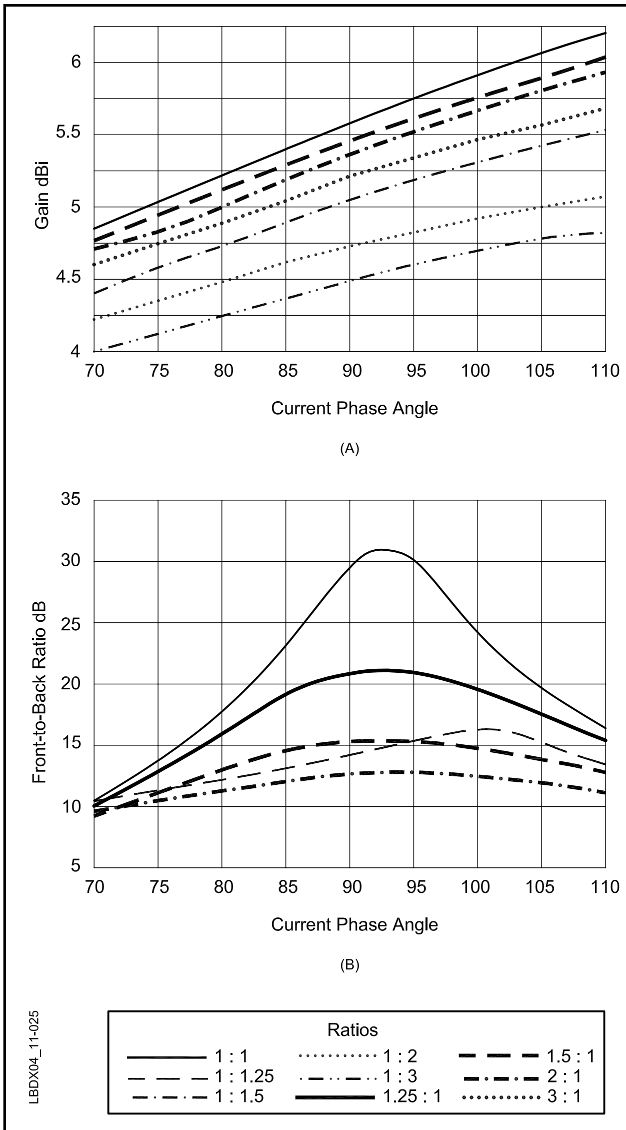


Fig 11-96 — Calculated gain and the front-to-back ratio of a 2-element end-fire array versus current magnitudes and phase shifts. Calculations are for very good ground at the main elevation angle. The array tolerates large variations so far as gain is concerned, but is very sensitive so far as front-to-back ratio is concerned.

Fortunately, as Gehrke states, these vertical arrays are relatively easy to get working. **Fig 11-96** shows the results of an analysis of the 2-element cardioid array with deviating feed currents. The feed-current magnitude ratio as well as the phase angle is quite forgiving as far as gain is concerned. As a matter of fact, a greater phase delay (eg 100° versus 90°) will increase the gain by about 0.3 dB.

The picture is totally different as far as F/B ratio is concerned. To achieve an F/B of better than 20 dB, the current magnitude as well as the phase angle need to be tightly controlled. But even with a "way off" feed system it looks like you always get between 8 and 12 dB of F/B ratio, which is indeed what we used to see from arrays that were incorrectly fed with coaxial phasing lines having the electrical length of the required phase shift.

When choosing a feed system one should keep in mind:

- Will the array be used as a receive antenna as well? If so, directivity is a main concern. If not, gain will be the main concern.
- Is bandwidth a main criterion? If you want to use the array to cover from 3.5 to 3.8 MHz you will need to go for a 90° hybrid based system or split up the band in sections (depending on the element Q).

3.4.10.1. Christman System

The Christman method makes maximum use of the transformation characteristics of coaxial feed lines, thus minimizing the number of discrete components required in the feed network. This is an attractive solution and should not scare off potential array builders. For a 2-element cardioid array this is certainly a good way to go. Of course you need to go through the trouble of measuring the impedances.

With arrays of more elements, it is likely that identical voltages will be found only on two lines. For the third line, lumped-constant networks will have to be added. In such cases the Lewallen or Lahlum/Lewallen method is preferred.

3.4.10.2. Lewallen and Lahlum/ Lewallen Systems

The following sections will provide more information on the Lewallen and Lahlum/Lewallen methods.

3.4.10.2.1. The Quadrature Lewallen System

The Lewallen feed system has been used very successfully by many array builders, especially those who want no compromises and care only for peak performance. The system can produce the right phase angle and feed current magnitude for any load impedance, and one can adjust ("tune") the values of the L-network to obtain the desired values.

In the last several editions of *The ARRL Antenna Book*, Roy Lewallen, W7EL, published a number of L-network values for the 2-element cardioid and the 4-element square arrays, which a builder can use for building the L-network without doing any measuring.

3.4.10.2.2. Any Phase Angle With Lahlum's Approach

Lahlum's approach introduces an extra feature that allows you to program any phase angle at any feed current magnitude. This so-called Lahlum/Lewallen method, which employs two L-networks, is "fully adjustable," which is a great advantage. Using quarter-wave (or ¾ wave) feed lines to your array elements, you can measure the voltage (magnitude and phase) at the start of these lines and tune the L-networks elements until you obtain exactly what you want. A simple procedure to do that is outlined in Section 3.6.1. An even more attractive measuring and tuning method using a vector scope is described in Section 3.5.2.

3.4.10.2.3. My Experience

After having used the non-optimized hybrid system very successfully for almost 15 years, I built and installed a feed system according the Lahlum/Lewallen system (**Fig 11-97**). In Section 3.6, I cover some test equipment I used for tuning the array (see also Chapter 7).

The design parameters for my particular Four Square (using one elevated radial, as described in Section 6) were:

Front element: I = 1.5 ∠–220° A
Center elements: I = 1 ∠–111° A
Back element: I = 0.85 ∠0° A

The network values were adjusted using a five-channel oscilloscope (three channels used) as shown in **Fig 11-98**. See also **www.seed-solutions.com/gregordy/Amateur%20 Radio/Experimentation/HexArray/UsingScope.htm.**

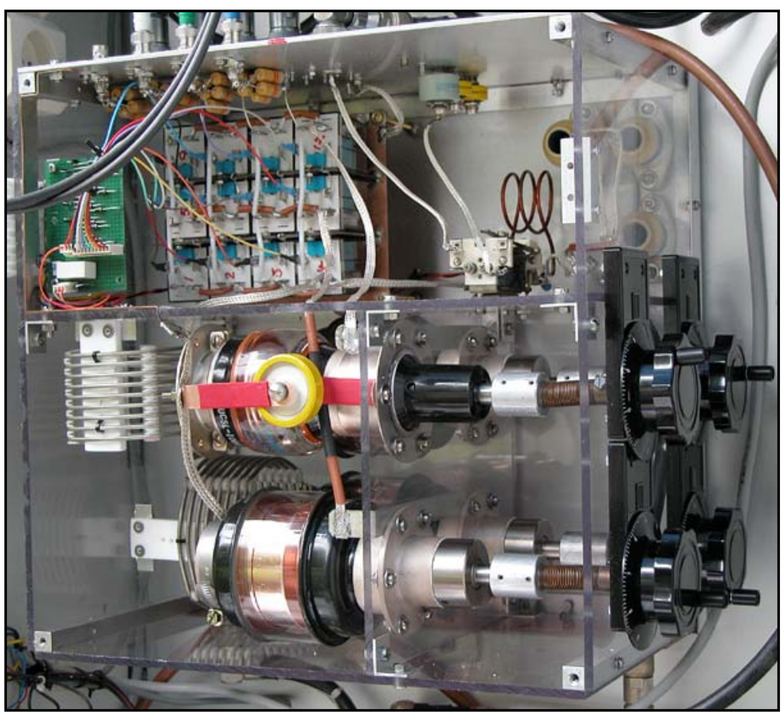


Fig 11-97 — Four-square feed system according the Lahlum/ Lewallen method, as built by the author with the help of Roger, ON6WU. In the two L-networks both components are continuously adjustable. The coil was replaced by a coil and vacuum capacitor in series to provide that facility. The unit includes an L-network for a perfect match to the feed line as well as an omnidirectional position.

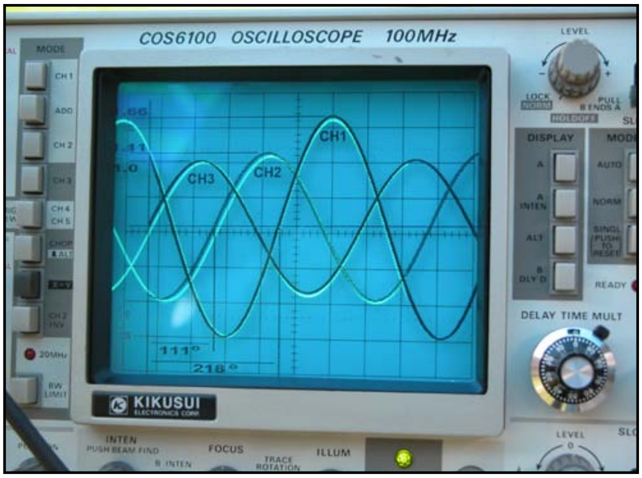


Fig 11-98 — A template showing the phase and magnitude relationship between the three drive currents of the optimized Four Square (see Section 4.7.2) was overlaid on the multichannel scope screen. All that was required to adjust the array was to tune the four vacuum capacitors until the scope pattern coincided with the template pattern.

These feed currents give about 0.4 dB more gain (see Section 4.7.2) than a perfectly working quadrature feeding solution and the directivity is significantly enhanced. I used a vector voltmeter to measure the voltages at the start of the λ/4 feed lines, and transformed to the feed currents at the elements. The measurement was confirmed using the method described in Section 3.6. A multichannel scope brought further confirmation. The design phase and amplitude are obtained through carefully adjusting the network.

More recently I reinstalled the Comtek hybrid. By the time this book is published it should be updated with one of the optimizations described in Section 3.4.6.5 and 3.4.6.6.

3.4.10.2.4. Bottom Line on the Lahlum/Lewallen Feed System

If you are not after operational bandwidth (say you only work phone or CW on 80 meters), and if you need to use your Four Square as your best receiving antenna, then the Lewallen/Lahlum feed system is a good choice. Calculating the component values is simple, thanks to the mathematical analysis provided by Robye, W1MK. The spreadsheet (*Lahlum-Lnetwork.xls*) is available on this book's CD, and the detailed explanation in Section 3.4.5.8 of this chapter lets you design your own system in a matter of minutes.

The most significant drawback is its limited operational bandwidth (approximately 80 kHz for ≥20 dB F/B) on 80 meters. An improved hybrid coupler feed system reaches approximately 300 kHz operational bandwidth.

To successfully build and properly implement and adjust the Lewallen/Lahlum system is, however, not a task for a "plug and play" ham. It requires good technical understanding and the proper test equipment to do so. One equipment manufacturer stopped advertising this system only for that reason.

3.4.10.3. Hybrid Coupler Feed System

Over the years Jim Miller, K4SQR, and his company Comtek must have supplied a very large majority of the hybrid drive systems for arrays used by the Amateur Radio community. In 2008, K4SQR sold Comtek to DX Engineering. Roger, ON6WU, measured a couple of Comtek units as well as the more recent system from DX-Engineering. **Fig 11-99** shows the Comtek and DX Engineering units.

The functional hybrid tests were done in the middle of the band (3.65 MHz) as these commercial units are sold for covering both the CW and phone ends of the band. These measurements were done with complex load impedances that are typical for a quadrature-fed Four Square installation at the end of the λ/4 75-Ω feed lines:

Back element = 11 + *j* 373 Ω
Front element = 44.9 – *j* 40.5 Ω
Parallel center elements = 58 + *j* 30 Ω

Table 11-16 shows the measurement data for the Comtek coupler and the DX Engineering coupler. The results are very similar. These measured results confirm the data obtained by modeling.

The DX Engineering unit provides an omnidirectional position, where all four elements are fed in phase. Another very useful feature is a built-in circuit that prevents you from going on the air while switching directions. (How many of us have not replaced any relays on a Comtek unit?)

The DX Engineering unit certainly excels when it comes to design and workmanship. The omnidirectional position and

Fig 11-99 — A look inside the Comtek (left) and DX Engineering (top) couplers.

Table 11-16
Operational Data for Commercial 80 Meter Hybrid Couplers

Model	*Port 4*	*Back El*	*Center El*	*Front El*
Comtek	–15 dB	1 ∠0°	0.92 ∠–82°	1.01 ∠–189°
DX Engineering	–19 dB	1 ∠0°	0.81 ∠–84°	1.06 ∠–191°

the safety feature certainly make this a good choice, although at a slightly higher price than the proven Comtek model.

Now that we know all about the ins and outs of the hybrid coupler feed system (Section 3.4.6), I expect that some of the more technically oriented low banders will do some experiments with new optimized hybrid systems.

3.4.10.4. Opposite Voltage Feed System

The concept of the opposite voltage feed system, developed by Pekka, OH1TV, had, so far, never been published. Only a few arrays have been constructed according to this principle and are reported to be working as anticipated.

No doubt we will see many of the more technical oriented low banders experiment with this feed system, and I have no doubt that further improvements and more detailed designs for various types of arrays will become available.

Conclusion

This system has advantages and disadvantages:

- It requires twice as much feed line as using current-forcing feed system.
- The operational bandwidth is substantially better compared to the Lewallen L-networks feed system, but not as good as the optimized hybrid coupler feed system.
- It is nice to be able to switch the "loading elements" at a common central location.

3.5. Measuring

None of the arrays described in this chapter can be built or set up without any measuring. The simplest array is a quadrature configuration using a hybrid coupler. This configuration uses no intentional optimization, but usually obtains the required k and θ within acceptable tolerances. Even in that simple case, the elements of the array will have to be tuned to proper resonance.

You can use the SWR meter in your transceiver to bring the elements to resonance where you want them. Just assume the point of lowest SWR is the resonant frequency (which is not quite true), and you will be close enough for an array fed in quadrature with a hybrid coupler.

The only other thing you should measure in such an array is the power dumped in your hybrid termination resistor. This should never be more than about 10% of the power going into the hybrid.

Okay, so far we have not needed any special test equipment!

3.5.1. Homemade Test Equipment and Surplus Professional Equipment

Until only a few years ago most of us needed to rely on rather simple homemade test equipment or to buy some relatively expensive secondhand professional test equipment for serious impedance and/or phase measurements. Professional network analyzers are, in principle, the ideal tools for measuring impedances. There are various types on the market, and secondhand you may be able to get a system comprising the analyzer and the generator for between $1500 and $3000. Nowadays you can buy a brand new VNA to work with a PC for just over $1000!

Things have changed a lot over the years and, as an example, excellent impedance measuring equipment has become available that can compete with the better professional equipment. In the past editions of this book I have described a number of test methods using simple test setups that relied on homemade adapters and interfaces with basic test equipment. These methods are still valid and may be very valuable, but in this edition I will mainly describe the use of commercial test equipment designed especially for Amateur Radio use.

3.5.2. The Antenna/Network Analyzer

There are a number of *antenna analyzers* on the market. These are single port test instruments that can measure the impedance of the load connected to that port. A serious antenna builder should have one of those. The better ones can perform a lot of tasks. The most important ones are:

- Perform precise impedance measurements.
- Display the complex impedance in several formats, as well as the SWR and return loss).
- Measure exact cable length.
- Measure cable attenuation.

The second group is the family of *vector network analyzers* (VNA). A VNA is a two port measuring device that allows you to measure filters, transformers and other components that have "in" and "out" ports.

Let us quickly go through what is available on the market and what is commonly used by hams.

3.5.2.1. The MFJ-259B

The MFJ-259B antenna analyzer is different from the older MFJ-259. It uses a microprocessor and four voltage detectors in a bridge to directly measure reactance, resistance and SWR. With so much information available, uses are limited mostly by your imagination and technical knowledge.

One main application for antenna builders is its capability of measuring SWR (also in terms of return loss). It will also measure the resistive part and the absolute value of the reactive part of complex impedances. The MFJ-259B isn't smart enough give the sign of the reactive part without some minor help. You must vary the frequency slightly and watch the reactance change to determine the sign of the reactance and the type of component required to resonate the system. If adjusting the frequency slightly higher increases reactance (X), the load is inductive and requires a series capacitance for resonance. If increasing frequency slightly reduces reactance, the load is capacitive and requires a series inductance for resonance. This general rule works with most antennas, but not necessarily all of them.

The only thing I did not like is the large number of batteries that go dead in the middle of a measuring session in the field. I strapped a 12 V sealed lead acid battery (2 Ah) to the analyzer, and now have plenty of portable power (see **Fig 11-100**). I often use the MFJ for initial measurements in

Fig 11-100 — I always use my MFJ-259B antenna analyzer outdoors with a small 12 V lead-acid battery strapped on its side, as I hate to run out of power in the field.

the field (I call it "ballpark' measurements). When I need more accurate data I switch to my AIM 4170 or VNA 2180.

What I like very much about the MFJ-259B is that you have an analog display and a tuning knob, which really gives you the feeling of having full control over the equipment. If you need to recalibrate your MFJ-259B visit **www.w8ji.com/mfj-259b_calibration.htm**.

3.5.2.2. The AIM 4170

Ever since the AIM 4170 antenna analyzer, developed by Bob Clunn, W5BIG, (**www.w5big.com**) and Array Solutions (**www.arraysolutions.com**) was introduced on the market, it has been widely accepted as being the best value for its money. Tests show its accuracy compares favorably with professional test equipment such as the HP3577A. At the time of writing this piece of equipment is widely considered as the Rolls Royce of affordable antenna test equipment. See **Figs 11-101** and **11-102**.

Some of the AIM 4170's outstanding features are:

- Very accurate (accuracy limited by quality of calibration standards).
- The SWR reference Z_0 can be set at any value (50 Ω, 75 Ω or even 550 Ω (see Chapter 7, Section 2.5.4 and Fig 7-91). High impedances (eg Beverage impedance) can be measured as accurately as low impedances.
- Can be calibrated at the end of a cable or even through a filter. This is very important as you can now measure at ground level, where you set up the AIM 4170 and a laptop in a car or a small tent.
- Excellent and easy way of measuring line loss (see Chapter 6, Fig 6-2).
- Excellent and easy software, which is regularly updated (**www.w5big.com/prog_update.htm**), and which includes a Smith Chart display.
- Excellent strong signal handling capability. The AIM 4170 can handle broadcast band (BC) signals that are up to 15 dB stronger than an MFJ-259B before the results are influenced. In addition, as the AIM 4170 can be calibrated with BC filter in the measuring line. This makes it possible to measure 160 meter antennas very close to BC antennas.

It would be nice if the unit used a USB interface instead of requiring a RS-232 serial port interface to the computer.

3.5.2.3. The N2PK VNA (Vector Network Analyzer)

This vector network analyzer, developed by N2PK, has for several years been one of the best VNAs (**Fig 11-103**). The unit was never commercialized but several kits were made available. I obtained one from Greg, W8WWV, and I can confirm that the results obtained are comparable to those obtained by very expensive network analyzers.

This unit is capable of both transmission and reflection measurements from 0.05 to 60 MHz, with about 0.035 Hz

Fig 11-101 — The W5BIG AIM 4170 antenna analyzer only measures 13 cm wide, 4 cm high and 10 cm deep. The unit pictured uses a SO-239 connector, but a version using an N connector is also available.

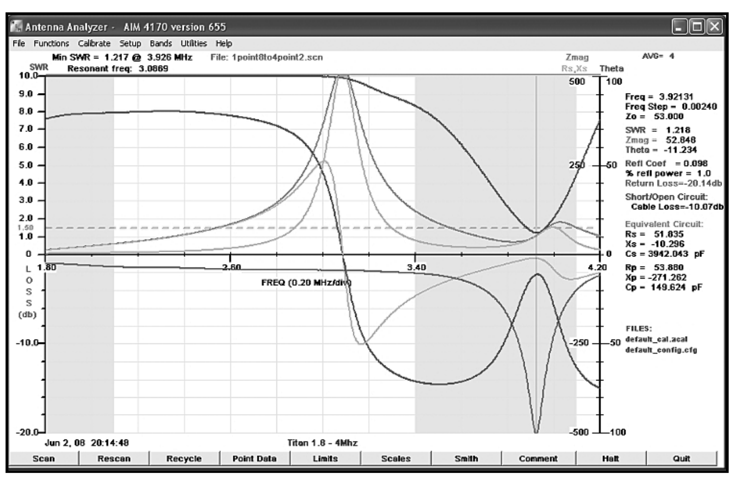


Fig 11-102 — The AIM 4170 software produces a very clear and colorful display, also displaying the numeric values of all parameters on the screen. The measured data can be saved in several formats, and the graph as a .bmp file.

Fig 11-103 — The N2PK-designed VNA. The cover was removed from the impedance bridge. The two interconnected N connector are used for doing two port measurements (network analyzer mode).

frequency resolution and >110 dB of dynamic range. Its transmission measurement capabilities include gain/loss magnitude, phase and group delay. Its reflection measurement capabilities include complex impedance and admittance, complex reflection coefficient, VSWR and return loss. Unlike other impedance measuring instruments that infer the sign of the reactance (sometimes incorrectly) from impedance trends with frequency, a VNA is able to make this determination from data at a single frequency. This is a direct result of measuring the phase as well as the magnitude of an RF signal at each test frequency

Paul, N2PK, impressed hams looking for an affordable network analyzer by the level of documentation that he has made available for anyone wanting to build a unit (**n2pk.com/VNA/VNAarch.html**). At the time the VNA design was published, its performance was *much* better than what could be obtained from anything else you might be able to build or buy for the same amount of money required to build this VNA. However, building a VNA is not for a first-time kit builder, although there are interest groups supporting potential builders (**www.seed-solutions.com/gregordy/Amateur%20Radio/Experimentation/N2PKVNA/N2PKVNA.htm**).

3.5.2.4. The VNA 2180

The VNA 2180 was developed by Bob Clunn, W5BIG (**www.w5big.com**) and Array Solutions (**www.arraysolutions.com**), the same team that brought us the AIM 4170 antenna analyzer. See **Fig 11-104**. This VNA has specifications that are very similar to those of the AIM 4170 analyzer, but this VNA has the great advantage that it has two ports. That makes it possible to measure the S21 transfer characteristics of filters and transformers, and even the diagonal isolation in a symmetrical array (see Section 3.4.6.4.3).

The signal output at port A is adjustable to +8 dBm maximum (compared to –18 dBm for the AIM 4170). This makes it possible to use the equipment on antennas where very strong signals from local stations are a problem when using a lower power level test signal. The VNA 2180 has an impressive dynamic range of more than 100 dB up to 50 MHz (still 80 dB at 160 MHz).

Impedances can be measured up to 10 kΩ. Unlike the AIM 4170, this unit is already equipped with a USB interface to the computer. The VNA 2180 also has a port for accessories, which can be controlled from the VNA 2180 software. Such an accessory is a multiplexer box making it possible to quickly scan through up to four inputs and display the antenna feed currents (magnitude and phase) on a vector scope image on your PC (see Section 3.6.2.).

What makes this piece of equipment unique is that Bob, W5BIG, used a number of novel design features for which patents have been obtained. The patents are based on a new way of making impedance measurements. This method uses an RF sine wave applied through both series and shunt connected resistors to get a voltage and phase measurement that is applied to mixers. The mixers down convert the results of the measurements to a low audio frequency of 2 kHz (to be compared to an IF in our receivers). Then the signals are sent through a band pass filter to an A/D converter and from there to a microprocessor for processing the data and generating the signals to be sent to the program on the PC connected to the VNA 2180.

This approach using the mixer is unique and has the advantage of being able to improve noise rejection, selectivity and dynamic range without having to use frequency tracking

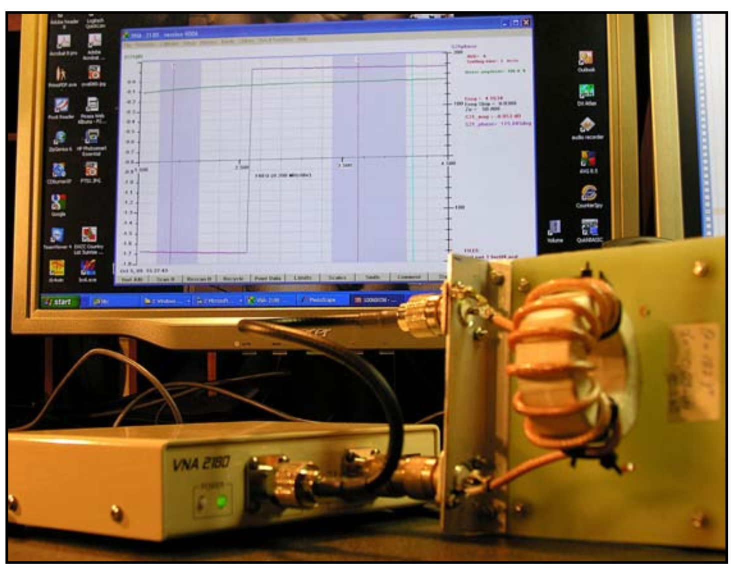


Fig 11-104 — The VNA 2180 vector network analyzer has specifications as good as many of the commercial brand-name VNAs costing many times as much. In this picture we see the unit measuring the insertion loss and phase angle for the 180° phase inversion transformer described in Section 3.4.6.3.

schemes. The use of a couple of resistors instead of diode bridges (as usually employed in VNAs) makes the front-end very tolerant to large RF voltages, nearby lightning strikes and a level of static discharges that would normally kill a diode bridge. The patent refers to the heterodyning of the measurement and processing of it in the microprocessor and PC.

3.5.2.5. Ten-Tec TAPR VNA 655

Another VNA that has been available since 2005 is the TAPR VNA 655, built and commercialized by Ten-Tec. This very nice unit has one disadvantage, its limited dynamic range of only 70 dB, which makes it a poor candidate to measure filters. The unit uses a wide-band detector that makes it susceptible to interference from strong signals at nearby frequencies (such as is the case with the MFJ-259B).

3.5.2.6. Which Antenna/Network Analyzer?

If you will never have to measure two-port components, such as transformers, amplifiers or filters, the AIM 4170 will be, at this time (mid-2009), without any doubt, your best buy.

If you want to be able to work with two-port components, the VNA 2180 is your logical choice. The multiplexer and the vector scope software makes this a most attractive tool for the serious array builders.

Rudy Severns, N6LF did an in depth comparison testing on a number of VNAs and antenna analyzers (N2PK-VNA, AIM 4170, Ten-Tec/TAPR VNA 655, MiniVNA and MFJ-259B). His report is available at **www.antennasbyn6lf.com/files/vna_comparisons.pdf.**

The ARRL also published a comparative test report between different makes of antenna analyzers and here too the AIM 4170 came out at the top. An overview of these test reports can be downloaded from **www.w5big.com/TestResultCombined.htm**.

3.5.3. Measuring Impedances in an Array

Obviously, measuring impedances (and calculating SWR/return loss) is the main and most direct task of an antenna analyzer or VNA.

Considering the radiating elements of an array, we need to measure their self impedance and their coupled impedance(s). As explained in Section 3.3.1, we can measure the *self- impedance* of a naked (uncoupled) element, and we can measure the *coupled impedance*, when the element is reacting to the presence of another element in its close vicinity.

Caution: If you use a coax between your antenna analyzer or VNA and the antenna feed point, make sure you have good quality common mode choke (high impedance) on that "transparent" feed line. If not, and if the ground system of the vertical is not perfect (0 Ω), there is a chance that the outside of your coax shield will act as a common mode source and upset the impedance measurement.

3.5.3.1. Self Impedance

Try not to measure the element impedance on a single frequency, but rather measure the impedance over a relatively wide bandwidth. For example, on 160 meters measure between 1.6 and 2 MHz. If on the graph representing the R and X values of the impedance you see sudden bumps, it is likely that the measurement is being influenced by strong BC signals. In that case you need to use a so-called "transparent" BC filter, which attenuates the BC signals but does not influence the measured impedance. The influence of the filter can be taken into account if you perform the calibration process with the BC filter in line.

Similar problems can exist at the high end of the 40-meter band (especially in Europe) at the time of day that BC signals above 7300 kHz are S9 + 50 dB. Try to make your impedance measurements during daytime, when D-layer absorption on the low bands is at maximum. This won't, of course, help you with strong ground wave signals.

You should not only measure the self impedances, but you should try to make them equal. This is important if you make an array that you can switch in two or more directions, and if you want to get maximum gain. Equalizing the *resonant frequency* of the naked elements can be done by changing the radiator length of the vertical elements, while equalizing the *self impedance* can be done by changing the number of radials used. If you start putting down perfectly identical and symmetrical radial systems, you will likely get very similar values for the resistive part of the various elements. If you cannot easily get equal impedances, you will have to suspect coupling from one or more of the array elements into another antenna or conducting structure.

Do not change the length of one of the radiators to get the equal values for the resistive parts of the elements. The elements should all have the same physical height (within a few percent).

3.5.3.2. Unwanted Mutual Coupling

When you have done your impedance measurement over a certain frequency range, look at the results on a *Smith Chart* plot. The results should be a gradually bent, curled curve with no sudden irregularities in it. If the curve shows an abrupt twist, it likely indicates that you have heavy mutual coupling to another antenna or tower. This should not be the case if you are doing a naked element (self impedance) measurement. If such a sudden twist is apparent on the chart, locate the other antenna or metal structure that is coupling to the element being measured.

If you happen to have towers or other metal structures or antennas within λ/4 of one of the elements of the array, it is possible that you will induce a lot of current into that tower by mutual coupling. The tower will act as a parasitic element, which will upset the radiation pattern of the array and also change the feed impedances of the elements and the array.

To eliminate the unwanted effect from the parasitic coupling, proceed as follows:

- Decouple all the elements of the array with the exception of the element closest to the suspect parasitic tower (leave the vertical elements floating).
- Measure the self impedance of the vertical under investigation and watch the shape of the Smith Chart curve.
- Detune the offending tower using one of the methods described in Chapter 7, Sections 2.11.1. and 3.10.
- After having detuned the offending tower, measure the self impedance of the vertical again. If you have properly detuned the parasitic tower, you will likely see a rise in impedance and a shift in resonant frequency, and the twist in the impedance curve will be gone.

• Another way to detune the offending tower is to fire the array toward the suspected tower and sample the RF current in the tower (see **Fig 11-105**). Detune the tower by minimizing the current in the tower.

3.5.3.3. Coupled Impedance

When you measure the *coupled impedance*, you want to see the influence of the nearby other element of your array on the impedance of the first element. In this case you must see an irregularity (twist) in the impedance curve on the Smith Chart near the resonant frequency of the elements.

3.5.5.4. Measuring Mutual Impedance?

In addition to the naked (uncoupled) impedance of each element, we also need to know the mutual impedance (between pairs of elements) in order to be able to do calculations on an array. As explained in Section 3.3.2., one cannot *measure* mutual impedance. Mutual impedance can only be *calculated* (from self impedance and coupled impedance). This can be done with the "Mutual Impedance and Driving Impedance" software module of the *New Low Band Software*, or with the *w1mk-on4un-oh1tv-arrays.xls* spreadsheet.

The same software module will also allow you to calculate the actual feed impedance of each element (being driven with a given current magnitude and phase). Check if the values you calculated are in the same ballpark as the results you obtained through modeling.

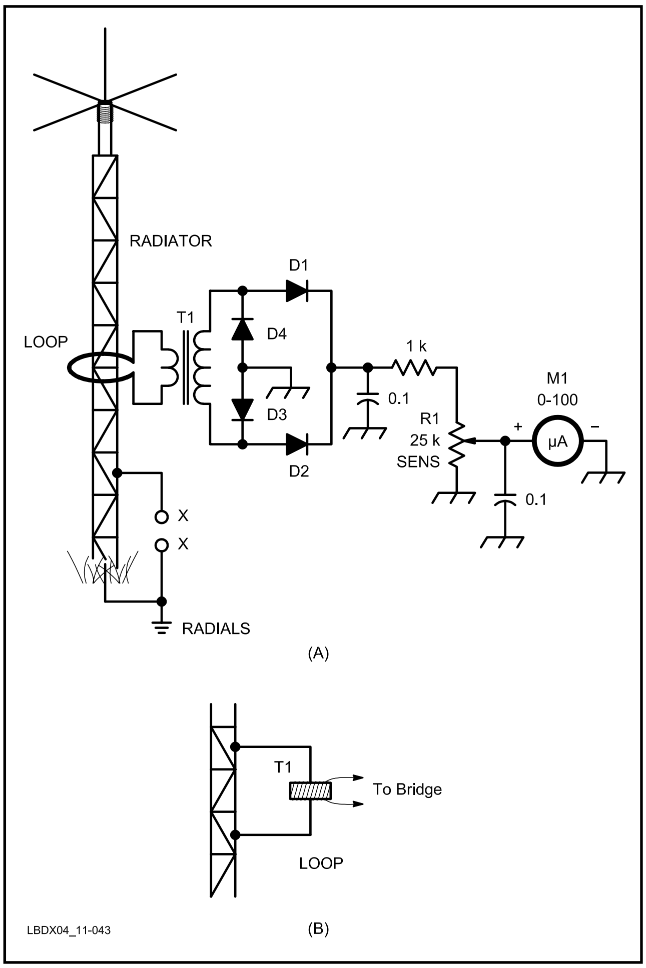


Fig 11-105 — Current–sampling methods for use with vertical antennas, as described by DeMaw, W1FB. Method A requires a single-turn loop of insulated wire around the tower. The loop is connected to a broadband transformer, T1. A high-mu ferrite toroid, as used with Beverage receiving antennas (see Chapter 7 on Receiving Antennas), can be used with a 2-turn primary and 2- to 10-turn secondary, depending on the power level used for testing.

3.5.3.5. Too Little Mutual Coupling Where You Want It

If you measure little or no difference between the self impedance and the coupled impedance, then have a look at the value of the self impedance. It is likely that the resistive part of the impedance is much higher than it should be. What is "should be"? It should be the impedance you have calculated by modeling the antenna.

Example: If you use inverted-L elements that are ⅛ λ vertical, you should expect a self impedance of approximately 17 Ω over a perfect ground. If you measure 50 Ω, it means that you have an equivalent loss resistance of 33 Ω! With so much loss resistance you will, even with very close coupling, as in the case in an array with ⅛ λ spacing, see only a little difference between self impedance and coupled impedance. Such an array will still show the same directivity, but its gain will be way down. In the above example the gain will be down 4 to 5 dB from what it would be over an excellent ground system. So, if you see no effect of mutual coupling where you should see it, suspect you have large losses involved somewhere.

3.5.4. Cutting Feed Lines of a Specific Length

Building a phased array *always* involves feed lines of a specific length. Cutting feed lines to an exact electrical length can be done in different ways, but it's best to use an antenna analyzer or VNA. Most of the analyzers have a special function to help you cut quarter-wave stubs.

Assume you need a length of coax of 70° on 3.5 MHz. This is equivalent to 90° on 3.5 × 90/70 = 4.5 MHz. Assume you are using RG-213 (VF = 0.66). The calculated length is 300 /(4.5 × 4) = 16.66 meters. Now cut a length of cable that is 17 meters long and measure the resonant frequency using your antenna analyzer. Look at the parallel impedance where the inductive part jumps from a very high positive value to a very high negative value at resonance. Now shorten the coax in small increments, until you reach resonance where you want it. Some analyzers such as the AIM 4170 have a special function to help you do that ("¼ wave stub").

3.5.5. Measuring Feed Current Magnitude and Phase

So far we have measured the elements and adjusted them to be as electrically identical as possible (all self impedances and coupled impedance nearly the same).

We also have chosen the feed system we want to use (see Section 3.4), and cut the feed lines and possible phasing lines to the required lengths, using the method described in Section 3.5.4 to do so.

All we need to do now is measure the current magnitude (k) and phase (θ) in each of the elements to see if the currents meet the design values.

3.5.5.1. Measuring Voltage Instead of Current

If the array elements are fed by current forcing, which means through feed lines that are ¼ λ or ¾ λ long, we can rely on a specific property of quarter-wave lines to measure voltage in order to tell us current. This property says that the current at the end of a quarter wave line, multiplied by the characteristic impedance of that line, equals the voltage at the other end of the line — of course with a 90° phase shift.

In many arrays the feed lines are ¼ λ or ¾ λ long and connected to a centrally located "phasing box." If that is the case we can simply measure the voltage at the phasing box with a vector voltmeter.

The HP8405A vector voltmeter is an ideal tool, provided you can find one with probes in good condition. Surplus HP8405As very often have defective probes! Do I really need such a lab-grade test-equipment? No. A very attractive, simple and inexpensive, but very accurate test method is described in Section 3.6.1. If you are serious about antenna measurements, a VNA and a multiplier box plus vector scope is the ultimate (see Section 3.6.2).

We should be aware of the fact that feed lines are never without losses, which means that the current-forcing principle does not really apply 100%. On a lossless cable, 75 V of RF at one end of a λ/4 long feed line will result in a 75 V / 75 Ω = 1 A feed current at the other end.

Take the case of a real feed line with high SWR, seeing, for example, a load impedance of $10 + j\,365\ \Omega$ (a typical feed impedance for a reflector element of a quadrature-fed Four Square). Using a λ/4 RG-11 type foam feed line with a nominal loss of 0.3 dB/100 ft, the current magnitude will be off only about 0.5%. The phase angle will be 94.3° instead of 90° in case of a lossless cable. In practice we should not be overly bothered by this difference, but, if we want to know the exact feed current at the elements, only current measurement using a current probe will give the correct answer.

3.5.5.2. Measuring Current

If we want to measure and compare the feed current at the base of the different array elements, this requires simultaneous access to the feed points of the different elements. The advantage of this method is that it is the most direct method of measuring the feed currents (without any intervening "lossy" elements, as is the case when measuring the voltage at the end of current-forcing feed lines).

The combination of a VNA and a multiplexer, as described in Section 3.6.2, is the ideal test setup for measuring and tuning the elements of a Four Square array or other antenna. For measuring and comparing feed currents in an array we will need as many "identical" current probes as we have elements. These current probes are in reality current transformers of which the main requirements are:

- They should be *transparent*, meaning that they do not influence the current that is being measured.
- The feed lines for the different probes (transformers) bringing the measured signal to a central point should be calibrated to be of exactly the same *electrical length*.

See **Figs 11-106** and **11-107**.

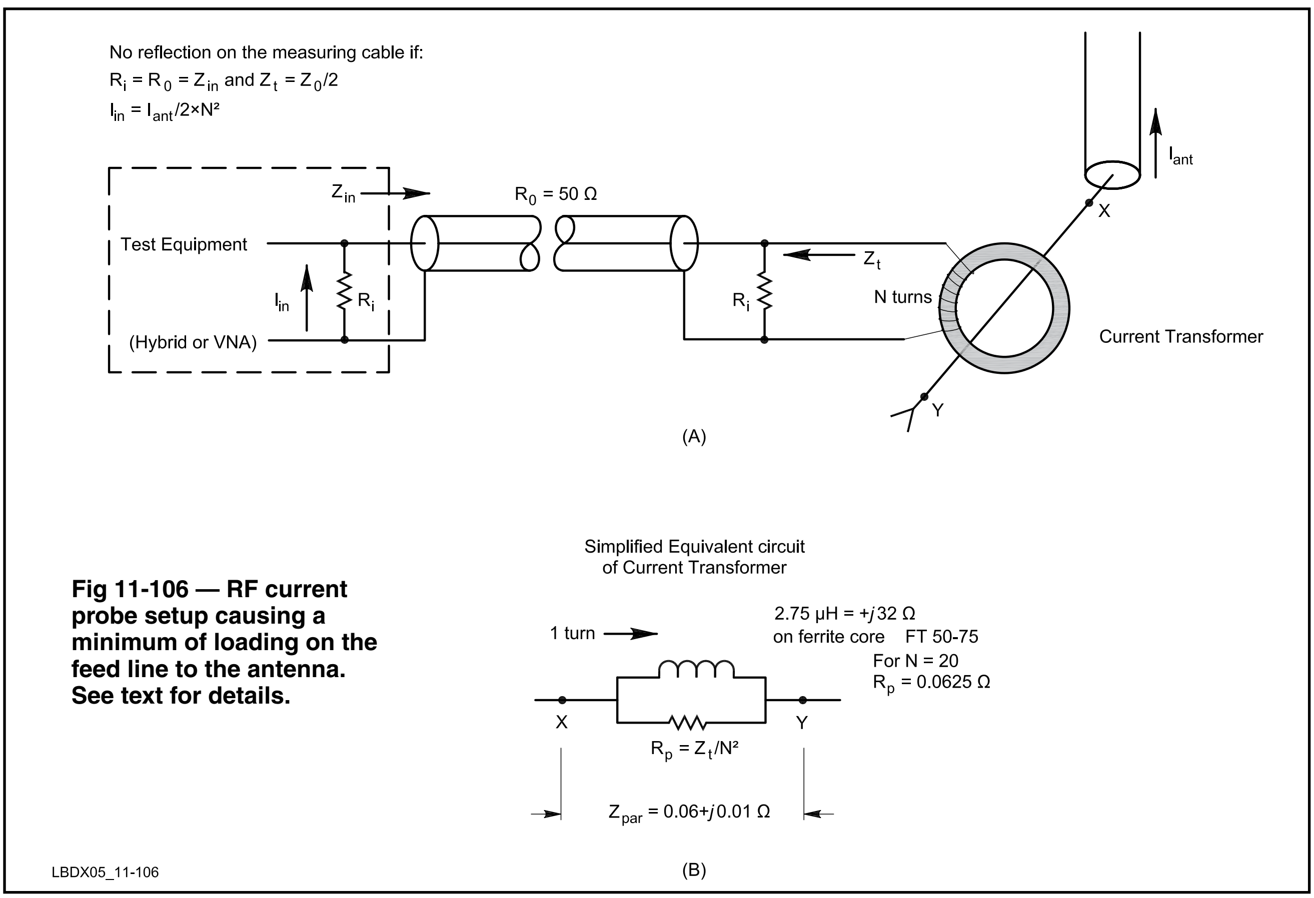


Fig 11-106 — RF current probe setup causing a minimum of loading on the feed line to the antenna. See text for details.

Fig 11-107 — Set of four identical RF current probes as described in Fig 11-106. Each transformer, along with its own cable (RG-58, about 25 meters long), was calibrated to be as identical as possible. The transformation loss varies between 32.5 dB on 160 meters and 32.8 dB on 80 meters.

Transparency

The insertion of a current transformer should interfere as little as possible with the impedance of the element under test. In Fig 11-106B we see a simplified equivalent circuit of the primary of the current transformer. The wire XY going through the center of the transformer core causes a series inductive reactance at the antenna feed point. Using a FT-50-75 (ferrite) core, a single turn through the core causes an inductance of approximately 2.75 µH (equal to + *j* 32 Ω on 1.8 MHz). In parallel with this we have the transformed impedance Z_t (total), which in this case is 25 Ω (using R_{out} and R_{in} = 50 Ω). This impedance is transformed in the current transformer (with 20 turns on the secondary) into R_p (parallel) = $25/20^2 = 0.0625$ Ω.

This resistance, in parallel with the + *j* 32 Ω impedance we calculated above (and which has little or no influence) is the total impedance Z_t inserted at the base of the antenna when we use such a current transformer. Z_t is so low that it will cause very little measurement error. To keep the error at a minimum, Z_t must be as small as possible. This is important especially if we are measuring impedances of short elements, where R_{rad} may be less than 10 Ω. If we had used only 10 turns on the primary, R_p would have been as high as 0.25 Ω, which would be meaningful when measuring antennas with a low R_{rad}.

If we want to make correct current measurements with this current probing system, we must take care that the cable between the transformer and the test equipment sees Z_0 in both directions. That is the only way to guarantee no reflections on the line (see Fig 11-106A).

Assume that $R1 = Z_0 = 50$ Ω. The transformation coupling of the current transformer is given by

$$I_{in} = \frac{I_{ant}}{2 \times N} = \frac{I_{ant}}{40}$$

where N is the number of secondary turns, in this case 20. The "loss" is 20 log (40) = 32 dB.

Identical Line Length

Make your probe measuring feed lines long enough so that you can reach the elements of future project arrays (eg 25 meters). Start with making the four feed lines to the current transformers so that they have an identical electrical length. Minor differences can be calibrated out using the VNA calibration procedure. The calibration is done by connecting each of the four probe sets (transformer plus feed line) one by one to a 50-Ω load resistor, and then letting the software take care of minor differences and calibrate probes 2, 3 and 4 with reference to probe 1 (the standard). If you will use the lines with a vector scope (see Section 3.6.3) you will be able to calibrate the probes using the VNA software (if you use the VNA 2180 and its associated multiplexer and software).

It's a good idea to equip these probes with plenty of common-mode choking on both ends.

If you use the probes with the 90° coupler for a zero adjustment using a 90° hybrid as shown in Fig 11-89, you may need an extra length of cable in one of the measuring lines (if the phase difference is different from 90°). If the magnitudes of the feed currents are different (k ≠ 1) you will need to use a suitable attenuator in the measuring line going to the element with the higher current. That makes it possible to align arrays using any θ (phase angle) and any k (current magnitude ratio), as shown in Fig 11-89.

3.6. Tuning an Array

Do we really need to *measure* feed current or feed voltage at the end of quarter-wave feed lines? No. We are not interested in the *values* of the feed currents, all we want is that they relate to one another as specified for that array (correct k and correct θ).

An attractive way to tune an array is to display the feed currents of the different array elements simultaneously on a *multi-channel scope* (see Fig 11-98). The word "simultaneous" is what characterizes this method. Tuning the variable components in an array fed with the Lahlum/Lewallen feed method interacts on the different channels, and unless you can see that interaction this can be very time consuming.

3.6.1. The Null Detector Method

The null method has been described in Section 1.27 of Chapter 7 on receiving antennas.

The 90° hybrid can be used as the heart of a simple but very effective phase-measuring device for quadrature-fed arrays. If two voltages of identical magnitude but 90° out of phase are applied, the bridge circuit will be fully balanced and the output is null. The design also comes from Rob, W1MK, (Ref 968). **Fig 11-108** shows the hybrid in a simple test circuit for a quadrature-fed Four Square. After having built the hybrid for the test circuit, use the layout described in **Fig 11-109** to test the hybrid (**Fig 11-110**).

Note that the principle can be used with phase angles other than 90° as well. Let's work with an example. **Fig 11-111** shows the feed system for the WA3FET Four Square, described in Section 4.7.2. The elements are fed via λ/4 feed lines, which means we can measure the voltages at the end of these lines to determine the currents at the antenna feed point (current equals voltage divided by feed line impedance).

Using a voltage divider (with a high enough dividing

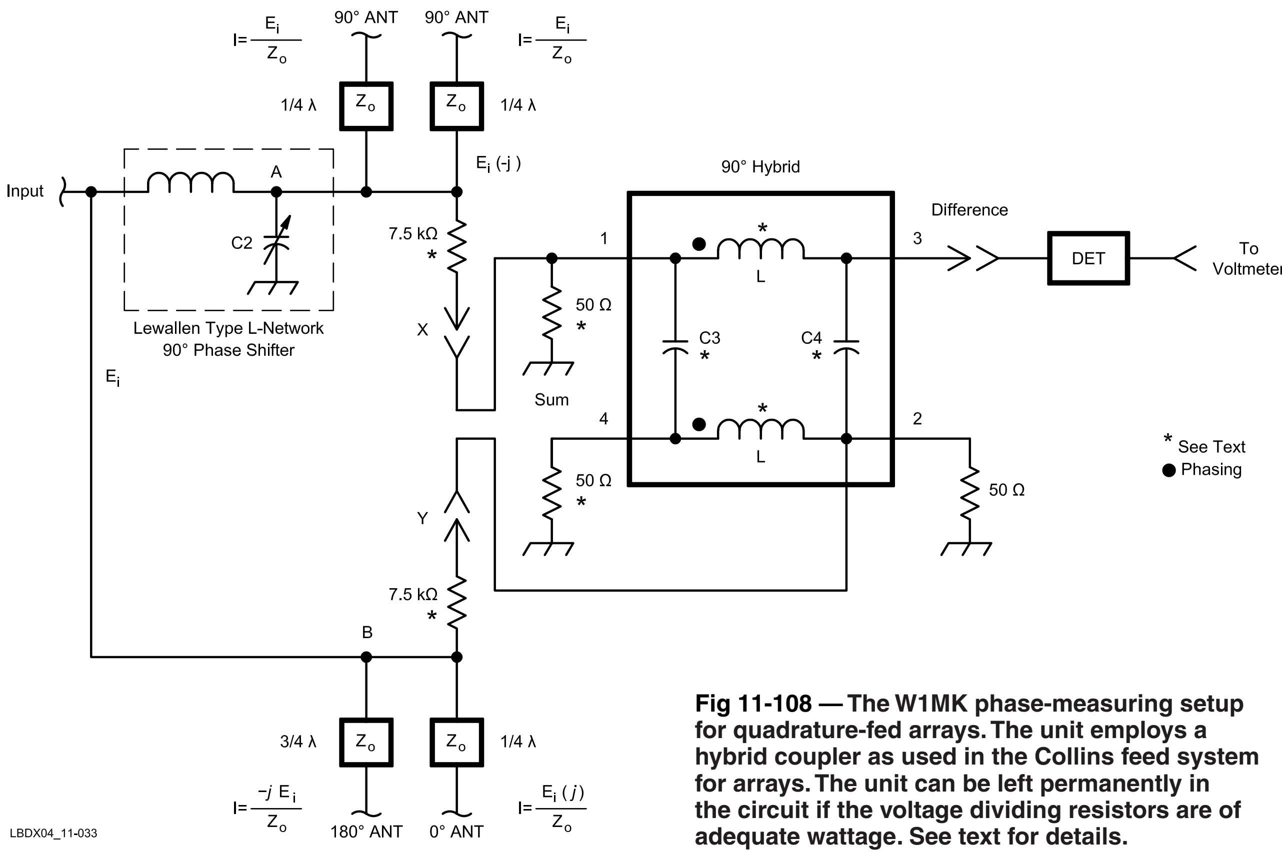


Fig 11-108 — The W1MK phase-measuring setup for quadrature-fed arrays. The unit employs a hybrid coupler as used in the Collins feed system for arrays. The unit can be left permanently in the circuit if the voltage dividing resistors are of adequate wattage. See text for details.

ratio so as not to disturb the impedance involved), we sample some voltages at those points and bring them with equal length coaxial cables to our hybrid-coupler test setup. Three possibilities exist:

- Assume first that the array is fed in 90° increments (quadrature feeding). The sampled voltage at the end of our probe lines will be 90° out of phase and the output of the hybrid coupler will be zero.
- Assume that we are feeding with 90° phase shift but with slightly unequal current magnitudes. In this case we need to compensate for that with a calibrated attenuator in the probe line at the hybrid coupler input. It is essential that the probe coaxial cables are terminated in their characteristic impedances so that line length equals phase shift.
- Assume the array is not fed in 90° current increments, but with a phase difference of 111° (such as between the center elements and the back element in the WA3FET Four Square). All we need to do in that case is insert an additional line length of (111 – 90) = 21° in the line going to the element with the leading phase, so that the net result again is 90°. See Fig 11-111.

In the same example the phase difference between the center elements and the director is –107°, hence we need an additional line length in the measuring setup of 17°. When measuring between points A and C, we need to insert a 1 dB (a 0.89:1 voltage ratio) attenuator in the line to point B to compensate for the unequal drive currents. The value of the sampling resistors depends on the power you want to do the testing with, and the detector's sensitivity.

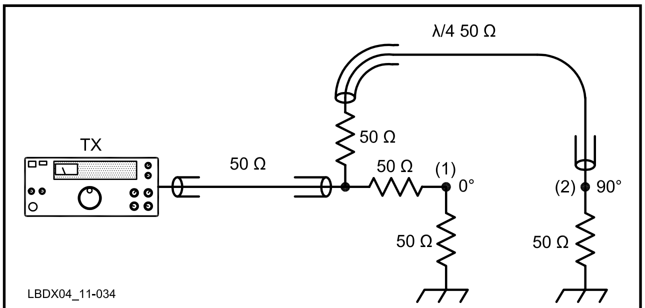


Fig 11-109 — In this phase-calibration system for the quadrature tester, RF voltage from the transmitter is divided down with two 50-Ω series resistors (to ensure a 1:1 SWR), routed directly to a 50 Ω load, and through a 90° long 50-Ω line (RG-58) to the second 50-Ω load. For a frequency of 3.65 MHz, the cable has a nominal length of 13.56 meters. The cable length should be tuned using the method described in Chapter 6 on feed lines and matching.

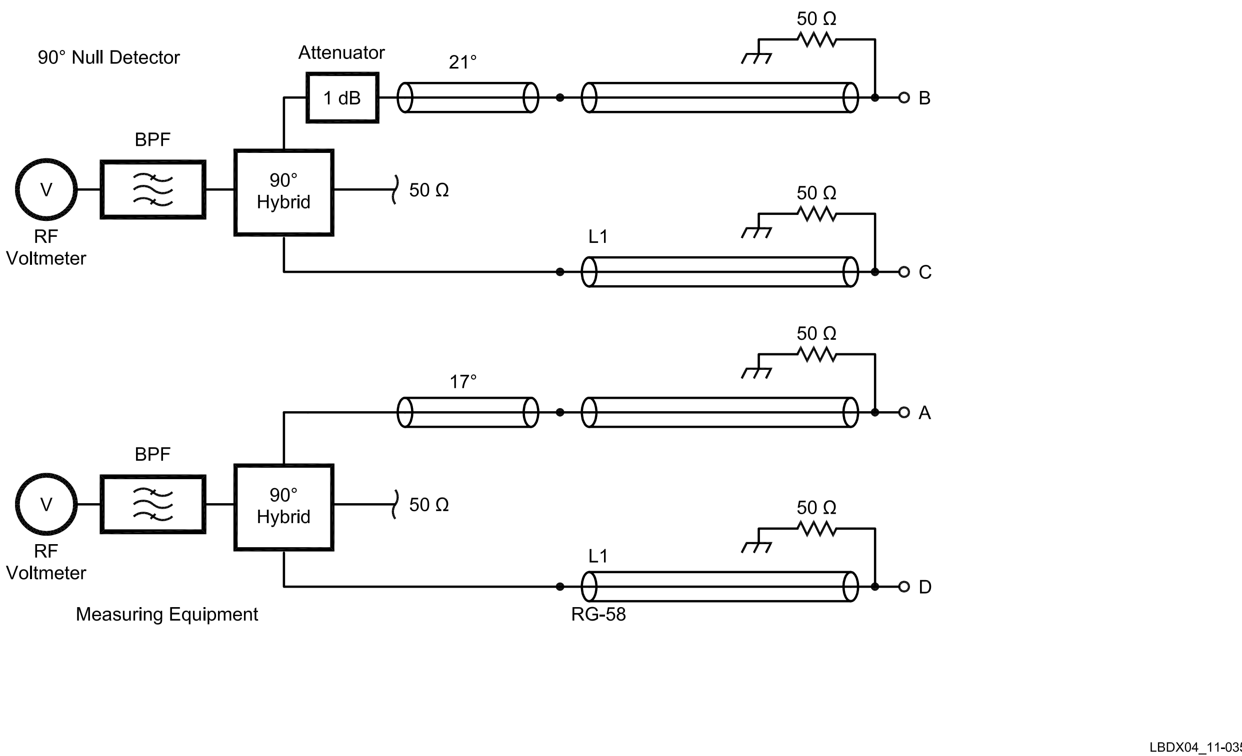


Fig 11-110 — Some RF is sampled at the end of the λ/4 lines going to the antenna elements. This is fed via RG-58 voltage sampling lines of equal length to the measuring equipment. Short line lengths and small attenuators can be inserted to compensate for non-quadrature setups and unequal drive currents. The schematic of the 90° hybrid is given in Fig 11-22. Section 3.4.6 explains how to calculate Xs1, Xp1, Xs2 and Xp2. V is a detector, which can be the detector/wattmeter described in Section 3.6.1 or a receiver. BPF is a bandpass filter.

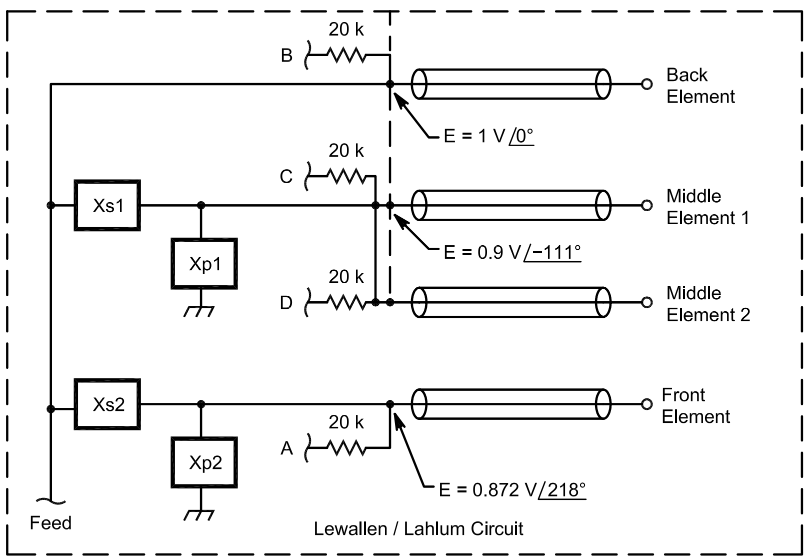
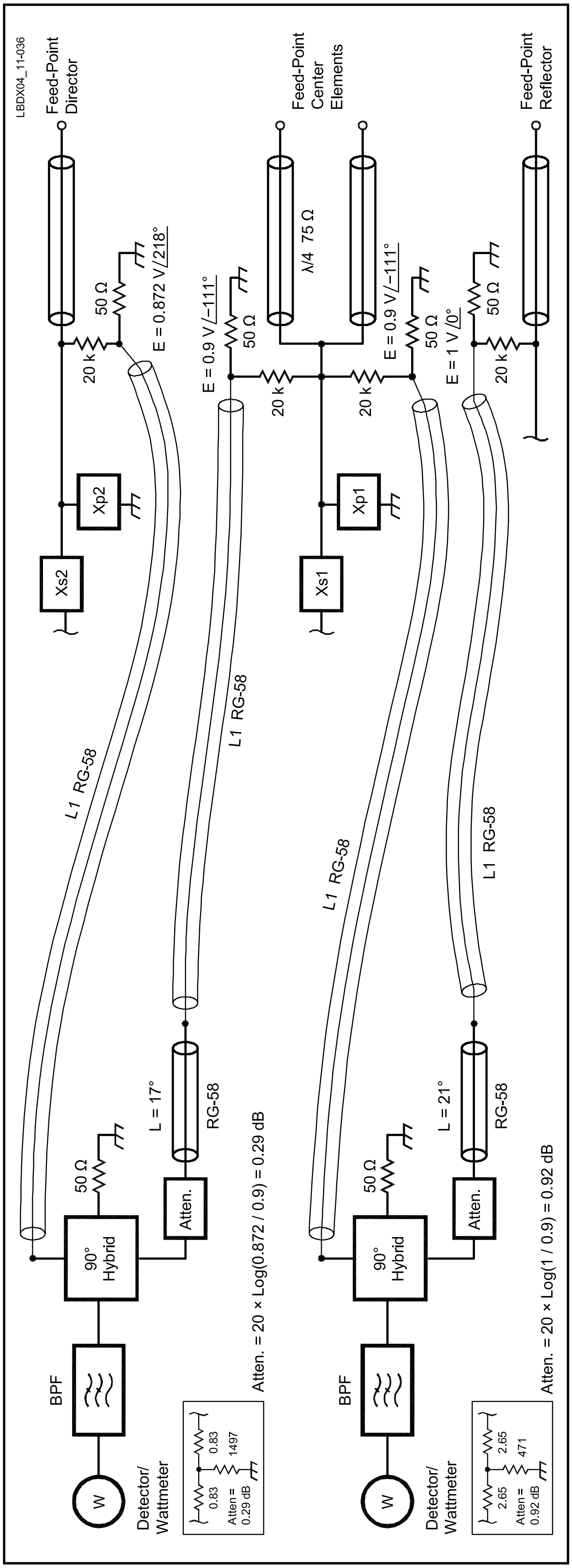


Fig 11-111 — Detailed schematic of the test setup for the WA3FET optimized Four-Square.



3.6.1.1. The W1MK Detector/Power Meter

Instead of using a receiver as null indicator, you can of course also use a dedicated detector/power meter as built by Rob, W1MK, and shown in **Fig 11-112**. This dual channel detector/wattmeter (a modified W7ZOI design), uses two AD8307 log amps that give a sensitivity of better than –70 dBm. In this circuit we see two identical detector/amplifiers, with three outputs: one for channel A, one for channel B and one for the sum of channels A and B. This comes in very handy when adjusting a Four Square array using the Lewallen/Lahlum feed methods using two independent L-networks.

The output of all three ports goes between 0 and 2 V, where 2 V equals 0 dBm and 0 V equal –80 dBm (see calibration chart in Fig 11-112). The maximum sensitivity is approximately –75 dBm. The unit has a bandwidth of approximately 500 MHz.

The circuit shown in **Fig 11-113** makes it possible to read the power in dBm on the scale of the digital voltmeter used as indicator.

The scaling is as follows: Power in dBm = mV/10. Some examples:

Power in = – 50 dBm → –500 mV
Power in = – 35 dBm → –350 mV
Power in = 0 dBm → 0 mV

M³ Electronics sells a similar power meter and frequency counter kit, model FPM1, at an attractive price (**www.m3electronix.com**).

3.6.1.2. Required Signal Levels, BC Interference, and Detector Sensitivity

Ideally we would want to be able to do some testing with an antenna analyzer such as the MFJ-259B as a signal source, and using a small detector/wattmeter as described in Section 3.6.1. This way we can work on the antenna with really portable equipment. This should do for initial tuning even if you are not able to get a null better than 30 dB. As a final touch up, you can always use the station transmitter as a signal source for doing final alignment.

What are the limiting factors?

- BC signals or even broadband noise.
- Detector sensitivity (noise figure).
- Available testing power.

W1MK says that when he starts a measurement session, he first measures the level of background signals or noise on the antenna. For that you simply connect the detector/wattmeter to the antenna you will be testing. A broadband noise level of –35 dBm for 80 meters and even more on 160 is not uncommon, and in some case can be much higher (10 or 20 dB higher!). These values will of course be different in different locations.

Adding a band-pass filter (BPF) in front of the broadband detector should drop the meter readings significantly. The values, of course, will be different for different locations. For example, W1MK experiences very high levels (–45 dBm) even with a BPF in front of the detector due to strong BC interference levels. In most situations the majority of the power hitting the detector is from out-of-band signals that, if not filtered out by a selective circuit, will reduce the amount of null that can be obtained. If the interference is inside the BPF passband, you can apply more power or use a receiver to provide more selectivity.

For minimum measurement error, a sampling resistor

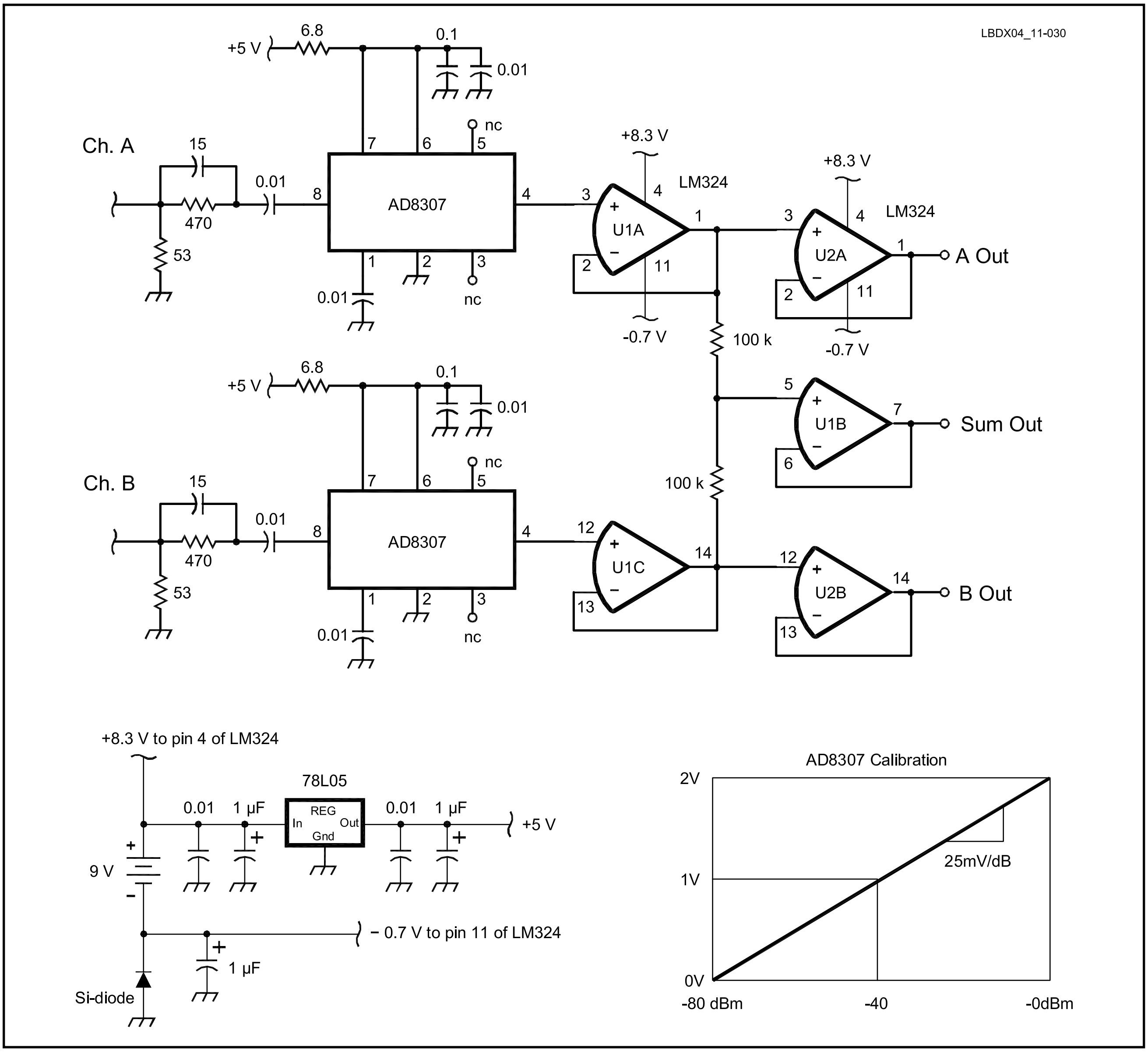


Fig 11-112 — Schematic circuit of the W1MK detector/power meter circuit. First connect one input and adjust the RF drive for 2 V output. Then the components of the LC circuit(s) are adjusted until the sum output (A + B) reads minimum.

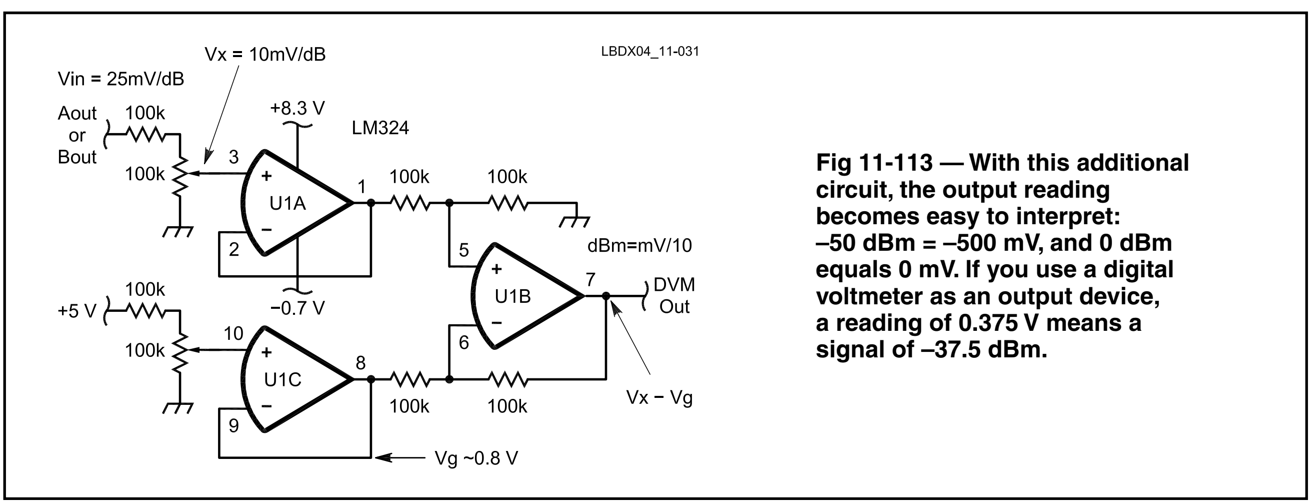


Fig 11-113 — With this additional circuit, the output reading becomes easy to interpret: –50 dBm = –500 mV, and 0 dBm equals 0 mV. If you use a digital voltmeter as an output device, a reading of 0.375 V means a signal of –37.5 dBm.

value of 20 kΩ is recommended. This means that the sampled signal will be approximately 52 dB down from the applied power. If we apply power with the MFJ-259, the level will be +13 – 52 = approximately –40 dBm.

If we use the detector/wattmeter described in Section 3.6.1.1 (which has a maximum sensitivity of –75 dBm) and if we are not limited by BC signals, we can see a null down as far as –35 dB. This is not bad for starters! An S9 + 40 dB signal represents –32 dBm, which means that the sensitivity of the detector/wattmeter matches pretty well with the level of a S9 + 40 dB signal, and even with such strong broadcast signals you will be able to see nulls of approximately –30 to –35 dB.

In case of very stubborn noise/interference problems you can, of course, use your receiver as a null detector. It has surplus sensitivity and should have enough selectivity to reject offending signals.

Your ability to obtain a deep null with a simple detector/wattmeter will always be either noise limited (the internal noise or the noise figure of the detector/wattmeter) or interference limited. If it is out-of-band interference, a BPF will help. If the interference is *on* your desired testing frequency you can move the test frequency slightly, or even better apply more power.

You might use 10-kΩ sampling resistors if sensitivity is a problem, but that is the limit. It is better to use higher testing power. A simple testing procedure is the following:

- Always use a bandpass filter at the input of the detector/wattmeter.
- Start your session with a portable source, such as the MFJ-259 antenna analyzer.
- Adjust the L-network values for maximum null. You should be able to obtain a null of at least –30 dB.
- If you are satisfied with a 30 dB null, now use your exciter as a signal source and apply 10 W (+40 dBm). This is about 27 dB better than the MFJ-259, which means that under the same circumstances you now will be able to see a null down to 50 dB.

For fine trimming of the phase and amplitude you must be able to make fine adjustments to both the series and the parallel reactances of the L-network. A variable capacitor is an obvious choice for fine trimming. You can make the equivalent of a variable inductor with a little trick. For example, if the network requires a coil with a reactance of +50 Ω, make a coil with double the reactance (100 Ω or 4.2 µH at 3.8 MHz) and connect in series a variable capacitor with (at maximum capacitance) a reactance of –50 Ω or less. If you use –25 Ω (1675 pF at 3.8 MHz), the series connection of the two elements will now yield a continuously variable reactance (at 3.8 MHz) of +25 (or less) to +75 Ω. See **Fig 11-114**.

The nice feature of such a test setup is that you can leave it permanently connected. Make sure that your sampling resistors are of high wattage if you run high power. Using 20-kΩ sampling resistors and running 1500 W the resistors dissipate 3.75 W, so try two 40-kΩ, 2-W resistors in parallel.

The sampled power level going into the hybrid is 50 to 60 dB down from the transmit power, which puts it in the 1 to 10 mW (0 to +10 dBm) level for 1000 W (+60 dBm) transmit power. A 40-dB null would show up as –30 to –40 dBm on your detector/wattmeter in the shack.

A –30 dBm level is 7 mV in 50 Ω. If you just want a kind of alarm system that tells you when things are really wrong, a simple germanium diode detector and a sensitive analog microamp meter (eg, 50 mA full scale) could be used.

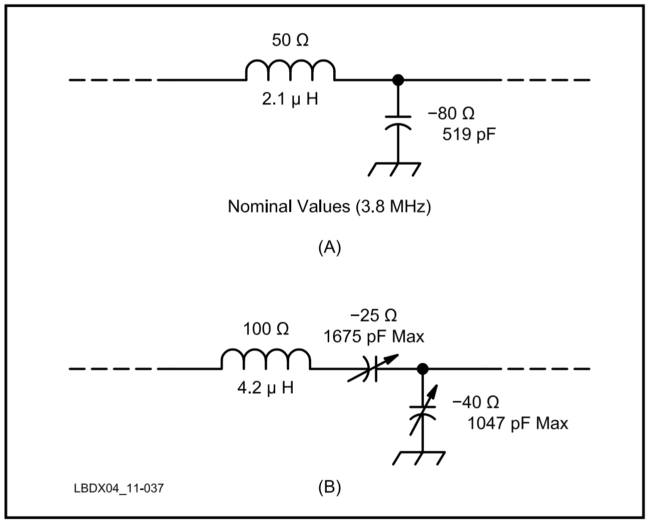


Fig 11-114 — To make the Lewallen L-network continuously adjustable, replace the coil with a coil of twice the required value and connect a capacitor in series. The net result will be a continuously variable reactance. With the values shown, the nominal +50 Ω reactance is adjustable from +75 to +25 Ω (and less). The two capacitors can be motor driven to make the phase-shift network remotely controllable.

Don't expect to have enough nulling sensitivity with this setup to properly adjust the L-network components. For that you need the sensitive wattmeter shown in Section 3.6.1.1. To avoid overdriving the detector-wattmeter you should provide a 10/20/30-dB step attenuator when running high power.

3.6.2. Using a Multiplexer

Tuning an array with more than two elements can be a tedious job. The null-detector method (Section 3.6.1) makes life a lot easier, but one thing is missing: When the tuning is off, you will know it is off (no null), but you have no information on *what* is off (magnitude or phase), in which leg and to what degree.

The multi-channel scope method (see Fig 11-98) is "real time," giving you all the details, but it but lacks accuracy.

This is where a multiplexer and the RVM (relative vector meter) method, first described by Greg Ordy, W8WWV, comes into the picture. This idea and setup was subject of a presentation by Greg at the Antenna Forum in Dayton in 2008 (**kkn.net/dayton2008/W8WWV08.pdf**).

The multiplexer is a fast switch that is controlled from the VNA software. It selects between a number of input channels. Each channel, when not selected is terminated in 50 Ω. The multiplexer allows the VNA to make an automated series of measurements. The multiplexer scans the lines from the current (or voltage) probes (see Section 3.5.5.2.), measuring the element feed currents at the antenna base or the feed voltage at the end of the current-forcing feed lines.

Based on this principle Array Solutions and Bob Clunn, W5BIG, developed a matching six port extender (or multiplexer) to work with the VNA 2180 (Section 3.5.2.4). The unit was in

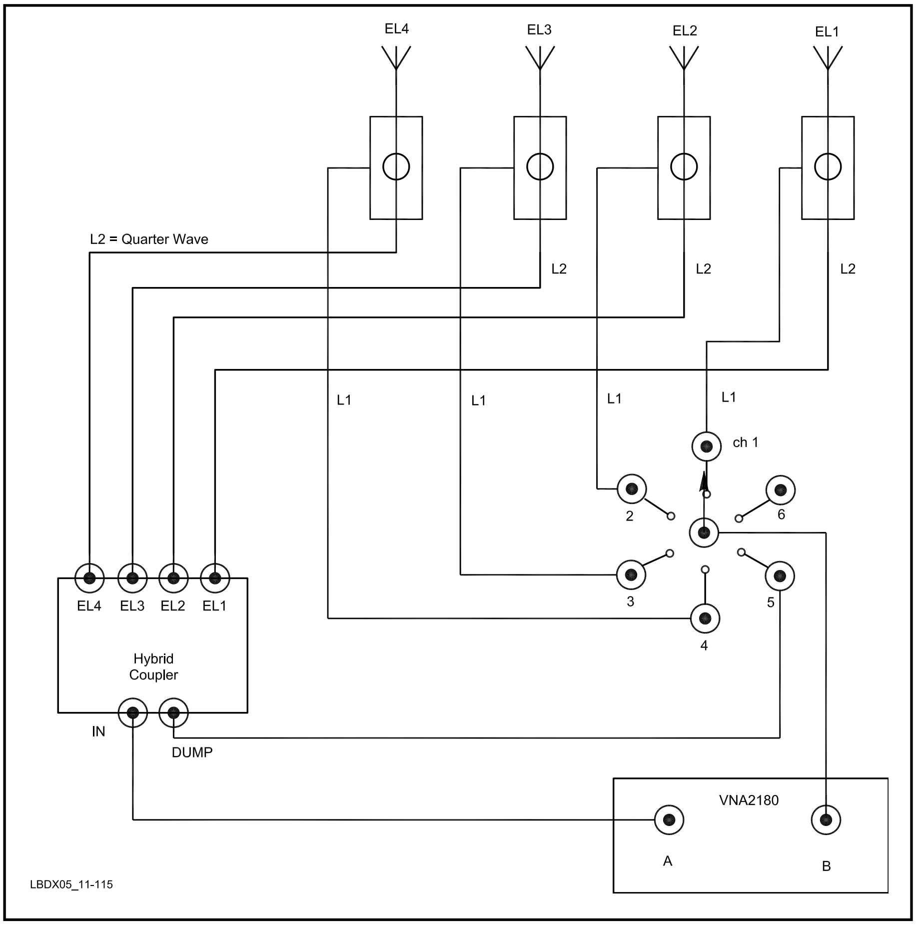


Fig 11-115 — RVM (relative vector meter) setup using the VNA 2180 and the matching port extender. See text for details.

development and not yet available when this section was written.

As shown in **Fig 11-115**, five of the six ports are used to analyze a Four Square array driven by a hybrid coupler. Four channels will measure the currents at the base of the four elements, channel 5 measures the power dumped by the hybrid coupler at its port 4.

This port extender can be used in two configurations: as a frequency scanning device, measuring data sequentially in port 1 through 5, or as a port scanning device measuring data at all the ports in fast sequence on a single frequency.

In the first configuration the VNA scans the set spectrum (in user defined frequency steps) and shows the measured data on one of two user-selected plots (see **Fig 11-116**) showing either the relative element feed current magnitude or the relative phase of these feed currents (relative to the values measured by channel # 1 of the multiplexer). The coupler's input SWR as well as dump power (in case of a hybrid coupler) are also shown on these graphs. The charts shown in Figs 11-116 and 11-117 were available during the development phase of the software. The final product charts may look slightly different.

The big advantage of this method is that it uses a swept frequency technique, which means you do the measurement not on one frequency but over a certain frequency range.

In the second (vector scope) configuration, the setup is used in single frequency mode together with the RVM screen. The feed current data are displayed as vectors on screen (see **Fig 11-117**). This screen is an almost real-time screen, where you can see the relative magnitude and relative phase of the feed currents change as you make adjustments to the feed system. This "almost real time" method is ideal for tuning

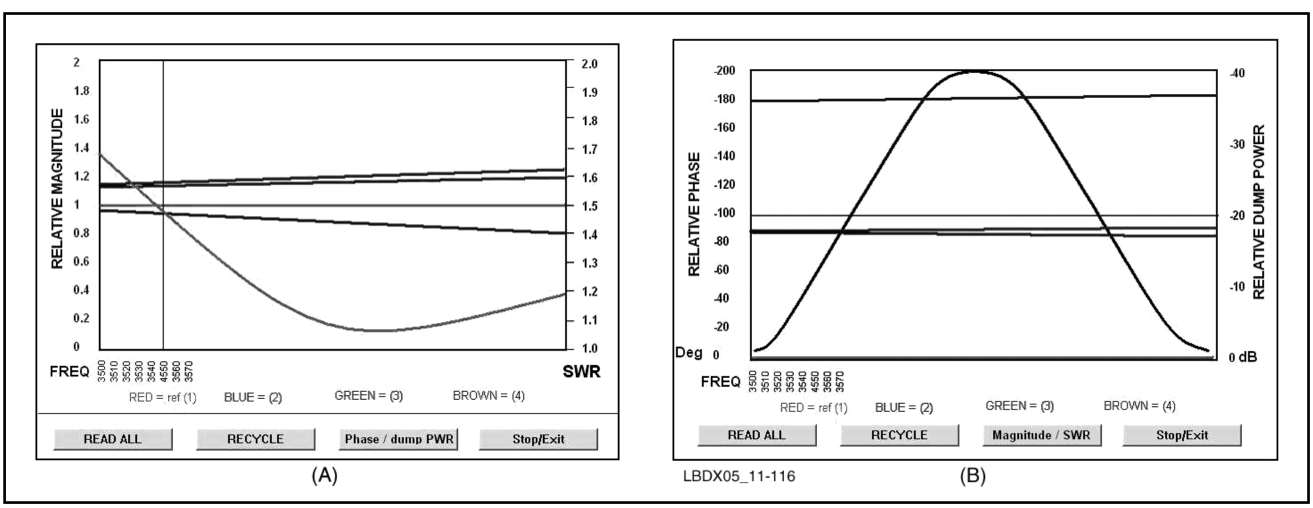


Fig 11-116 — The software that comes with the multiplexer allows you to generate charts showing you all the important parameters — phase angle and current magnitude for each element, input SWR and dump power (if using a hybrid coupler).

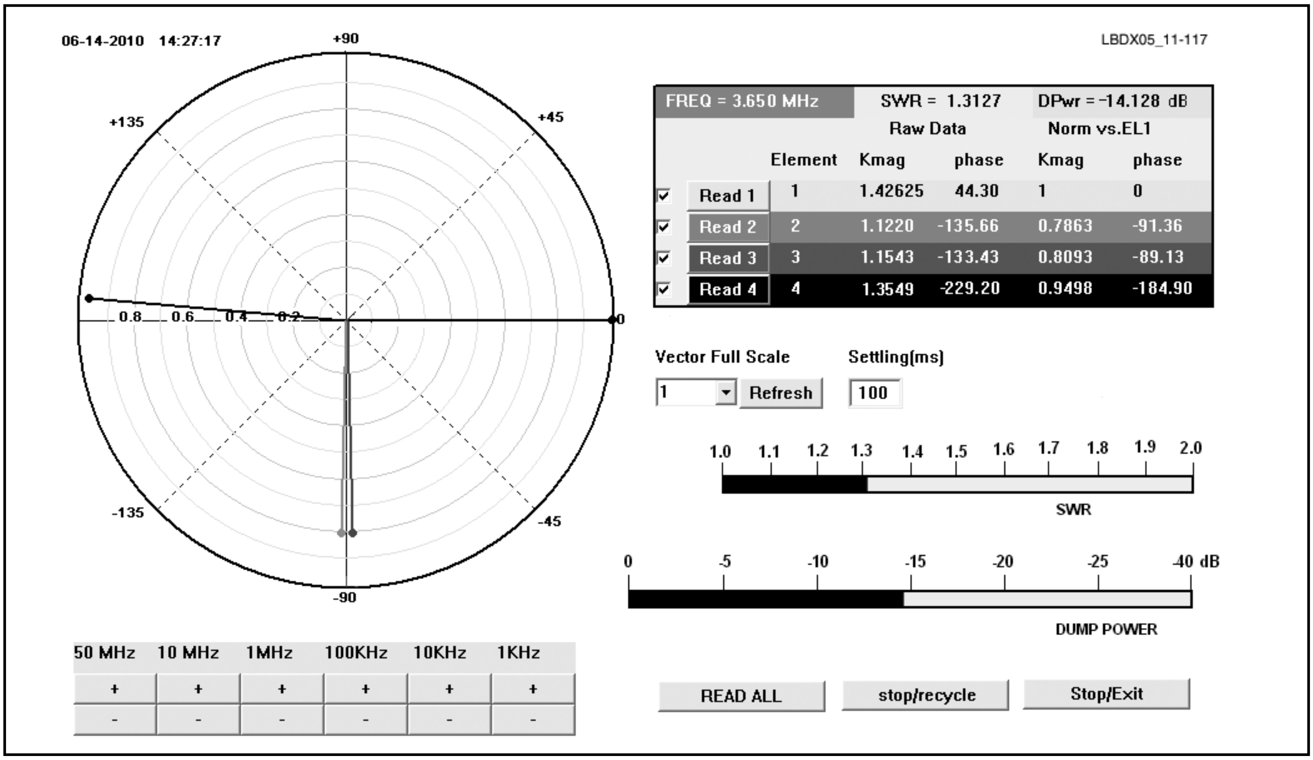


Fig 11-117 — The vector scope screen of the software developed by Bob, W5BIG, and that works with the VNA 2180 vector network analyzer. The circular vector plot shows the relative feed current magnitude (the length of the vectors) and phase angle for up to five measurement channels. The absolute current value, and the values normalized vs the current in element 1 are also shown in the table next to the circular display. The dumped power level and hybrid input SWR are also shown as bar graphs.

the array. One can see the current vectors change length and rotate while adjusting the feed system.

I am confident that using this port extender/multiplexer system and the RVM screen is *the* way to go for correctly and swiftly measuring and adjusting the driving/phasing systems in arrays.

3.7. Network Component Ratings

When designing array feed networks using the computer modules from the *New Low Band Software*, you can use absolute currents instead of relative currents.

Example: The feed currents for the 2-element cardioid array used as a design example have so far been specified as:

$I1 = 1\text{ A } \angle{-90°}$
$I2 = 1\text{ A } \angle 0°$

The feed-point impedances of the array are:

$Z1 = 51 + j\,20\ \Omega$
$Z2 = 21 - j\,20\ \Omega$

With 1 A antenna current in each element, the total power taken by the array is 51 + 21 = 72 W. If the power is 1500 W, the true current in each of the elements will be

$$I = \sqrt{\frac{1500}{72}} = 4.56\text{ A}$$

Using this current magnitude in the relevant computer program module "Coaxial Transformer" will now show the user the real current and voltage information all through the network design phase. The components can be chosen according to the current and voltage information shown.

If there is any question as to the voltage rating of any of the feed lines that are used as transformers in our designs (all have an SWR greater than 1), the program "Feed Line Voltage" with the real current as an input can be used to calculate the highest voltage at any point on the line. We find that for the 2-element array with a cardioid pattern (fed according to the Christman method), the highest voltage on a feed line of any length to element 1 (which has a 1.48:1 SWR) is only 397 V with 3 kW applied. For feed line 2, the maximum voltage is 352 V. For the Four Square array with ¼ λ spacing, the feed-line-voltage values are 234 V, 253 V, 253 V and 391 V. This should not represent any problem with good-quality RG-213 cable. In a similar fashion the voltages across capacitors or currents through capacitors in the lumped-constant networks can be determined.

When evaluating coils, use the following guidelines: For up to 5 µH, it is advisable to use air-wound coils. The best Q factors are achieved with coils having a length-to-diameter ratio of approximately 1:1. For higher values, use powdered-iron toroidal cores if necessary (never use ferrite material for these applications). Information on this subject as well as on the subject of dimensioning capacitors in a network is given in Chapter 6, on feed lines and matching. The module "Coil Calculation" of the *New Low Band Software* may be helpful in designing the coils. If you plan to build your own Lahlum/Lewallen network, it's a good idea to stick to air-wound coils (have a look at Fig 11-97) for in-

Table 11-17
Maximum Inductance with Various Toroid Cores
Maximum inductance for a single layer winding, as a function of wire diameter

Type	A_L	*#10*	*#12*	*#14*	*#16*	*#18*	*#20*
T-106-2	135	3.9	6	9	15	22	35
T-157-2	140	12	18	24	47	68	110
T-200-2	120	16	25	40	65	95	153
T-200A-2	218	29	46	73	119	172	278

ductances up to approximately 5 µH. Above this value you will have to revert to toroidal cores. Ferrite cores should not be used in this application as they tend to be unstable under certain circumstances. Only use powdered-iron cores. The red cores (mix 2) are a good choice for 160, 80 and 40 meters. How large a core do you need to use? The rule is never to wind more than a single layer. **Table 11-17** gives you the maximum inductance that you can get with a given wire size for a given core.

Example: Assume you need a reactance of 800 Ω. On 1.83 MHz that represents 69 µH. You may marginally make it on a T-157-2 core with #18 AWG wire. In most cases where such high values of inductance are involved, current through the coil will be very small, and #18 AWG enameled wire will be just fine. In cases where inductances of between 10 and 15 µH are required, I would use a T-200 or T-200A core with #10 AWG or even #8 AWG wire.

4. POPULAR ARRAYS

Whereas in previous editions of this book I described in detail how various feed systems can be applied to various arrays, with one exception I will only use two feed systems when describing a number of arrays in detail. Where applicable, for quadrature feeding, I will use:

- The hybrid coupler method (the optimized hybrid coupler version).
- The Lewallen/Lahlum L-network feed method, which allows utmost flexibility, but has limited operational bandwidth.

All arrays were modeled using *NEC-2* over "Average Ground" (ρ = 5 mS/m, ε = 13), with an extensive radial system that accounts for an equivalent series loss resistance of 2 Ω (for each element). The element feed-point impedances shown include this 2 Ω of loss resistance. If you want to calculate your feed system for different equivalent ground loss resistances, apply the following procedure:

- Take the values from the array data (given later). The resistive part includes 2 Ω of loss resistance. If you want the feed-point impedance with 10 Ω of loss resistance, just add 8 Ω to the resistive part of the feed-point impedance shown in the array data. The imaginary part of the impedance remains unchanged.
- Follow the feed-system design criteria as shown, but apply the new feed-point impedance values.

Most of the arrays in Section 4 were modeled on 80 meters (3.65 MHz), where the length of the radiating elements was adjusted for naked self resonance on 3.65 MHz. The modeling was done using an element diameter of 40 mm. If modeling was done on 160 meters, a vertical diameter of 250 mm (tower section) was used. The arrays that were modeled on 160 meters have elements that are self resonant on 1.83 MHz.

The gain is expressed in dBi (over good ground as specified above). In most models the *NEC-2* engine was used (available to everyone), but some modeling was done using *NEC-4*, especially in order to obtain more realistic gain figures. In such cases this will be mentioned in the text.

For each array we also calculated the directivity, expressed in RDF (Receiving Directivity Factor) and in DMF (Directivity Merit Figure). See Chapter 7 for an explanation of these terms.

In some arrays you will see a negative impedance for the "back" element of the array. The negative impedance merely means that the feed network is not supplying power to that element but rather "taking" power from that element. The different modules of the *New Low Band Software* as well as the *Lahlum-Lnetwork.xls* spreadsheet program handle these negative values without problems.

All Lahlum/Lewallen feed networks are calculated without taking into consideration the effects of cable losses. These effects are quite small if good cables are used. Only with very long cable lengths are losses significant (for example, ¾ λ current-forcing feed lines plus a 180° phasing line). I made several calculations between "ideal case" (no losses) and the "real world" case, and the differences of the L-networks values were well within the typical tuning range of the components. When you take into account the losses, the feed impedance of the network will be slightly higher (typically a few percent).

4.1 The Two-Element End-Fire Array

The principles of operation of the 2-element end-fire array were explained in detail in Chapter 7, Section 1.6. Most of us

Table 11-18
Data for 2-Element End-Fire Arrays
For Average Ground (conductivity = 5 mS, ε = 13). Includes 2 Ω equivalent ground loss resistance.
Reference single vertical element gain = 0.34 dBi.

Spacing (°)	*Phase (°)*	*Gain (dBi)*	*3-dB angle (°)*	*RDF (dB)*	*DMF (dB)*	*Z (Ω) front el*	*Z (Ω) back el*	*EZNEC modeling file*
105	–90	3.38	178	8.14	13.1	55+*j*14	20–*j*14	CH11-2el-endfire-105-90.EZ
90	–90	3.36	178	8.12	12.3	53+*j*19	22–*j*19	CH11-2el-endfire-90-90.EZ
90	–105	3.84	160	8.70	14.4	48+*j*23	18–*j*15	CH11-2el-endfire-90-105.EZ
90	–110	3.99	155	8.89	15.1	46+*j*23	17–*j*13	CH11-2el-endfire-90-110.EZ
75	–120	4.14	146	9.14	15.6	36+*j*26	15–*j*15	CH11-2el-endfire-75-120.EZ
60	–135	4.29	135	9.50	17.2	24+*j*25	12–*j*15	CH11-2el-endfire-60-135.EZ
45	–145	4.22	132	9.57	16.6	17+*j*24	14–*j*22	CH11-2el-endfire-45-145.EZ

probably think of a λ/4 spaced array where the elements are fed 90° out of phase, but this is not necessarily the best solution. If you want to use 90° phase shift, for instance because you want to use a hybrid coupler to feed the array, then a spacing of about 105° achieves a little better DMF than a spacing of 90°. Staying with quarter-wave spacing, a phase difference of approximately 110° is recommended. The wider the elements are spaced, the better the bandwidth, and that shows from the element impedances. Small arrays, such as with λ/8 spacing, give excellent directivity but the element feed impedances become low, causing drop in gain (for a given ground system) and bandwidth over which directivity will hold.

4.1.1. Array Data

Table 11-18 gives the main data for a range of 2-element end-fire arrays. The first impression is that 60° spacing with 135° phase shift is best, but note the relatively low feed impedance, which means narrower bandwidth than a wider spaced array.

The gain figures are over Average ground (ε = 13 and ρ = 5 mS). You can calculate the gain with reference to a single element (under the same circumstances of ground quality and ground loss) by deducting the single element gain (in dBi) from the listed dBi gain. Example: For a 90° spacing / 90° phase shift array, the gain over a single element is 3.36 – 0.34 = 3.02 dB

The modeling files listed in the table are available on this book's CD.

4.1.2. Feed Systems

Several feed methods for 2-element end-fire arrays were illustrated with a 2-element end-fire array in Sections 3.4.1 through 3.4.5.6, in Section 2.4.6.8.1.and Section 3.4.8.

4.1.2.1. Christman Feed System for the 2-Element End-Fire Array

This approach uses a minimum of components, but it does not using current-forcing feed lines so you cannot measure voltage to know feed current. This means you either need to be able to measure feed current (not so easy to do accurately), or you need to do some precise element impedance measurements (coupled and uncoupled), calculate the mutual coupling and from there calculate the feed impedances. Use the module "Mutual Impedance and Driving Impedance" from the *Low Band DXing Software*.

Fig 11-118 shows how you can switch the array in the two end-fire directions. When both elements are fed in-phase the array will have a bidirectional broadside pattern (see Section 4.2) with a gain of 1 dB over a single vertical. The front-to-side ratio is only 3 dB. The feed impedance of two quarter-wave-spaced elements fed in-phase is approximately 57 – *j* 15 Ω, assuming an almost-perfect ground system with 2 Ω equivalent-ground-loss resistance. Notice that both elements have the same impedance, which is logical since they are fed in-phase.

We can easily add the broadside direction (both elements fed in phase) by adding a switch (relay) that shorts the 71° long phasing line.

L-networks can be designed to match the array output impedance to the feed line. Don't forget that you need to measure impedances in order to calculate the line lengths that will give you the required phase shifts. Merely going by published figures will not get you optimum performance!

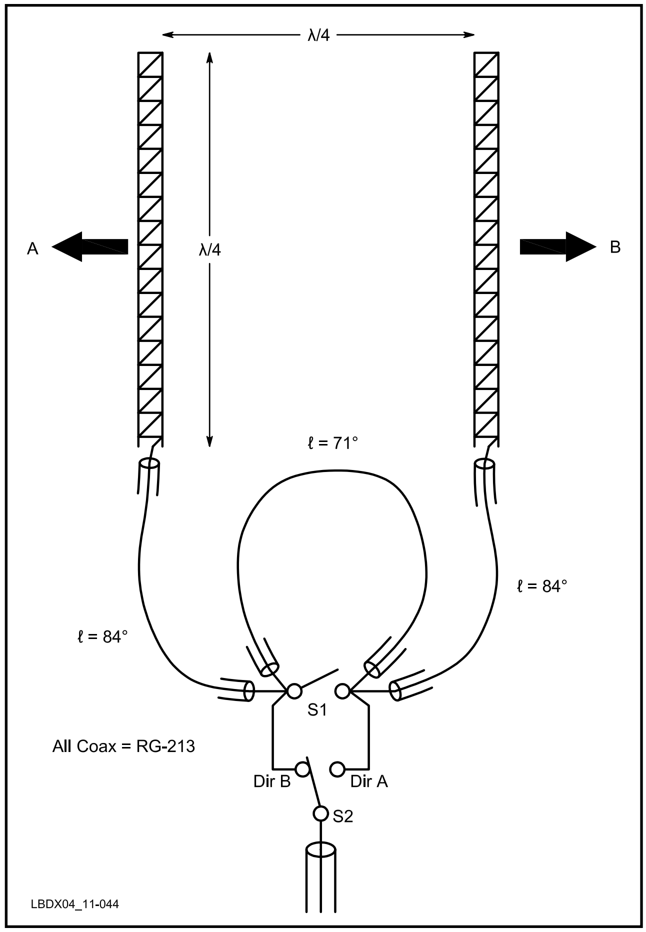


Fig 11-118 — The 2-element vertical array (λ/4 spacing) can be fed in-phase to cover the broadside directions. I added switch S1 to the Christman feed system as described in Fig 11-8. When S1 is closed, both antennas are fed in-phase, resulting in bi-directional broadside radiation.

4.1.2.2. Lewallen Feed, 2-Element End-Fire Array

The application of the Lewallen feed method for the 2-element end-fire array was described in detail in Section 3.4.5.4.

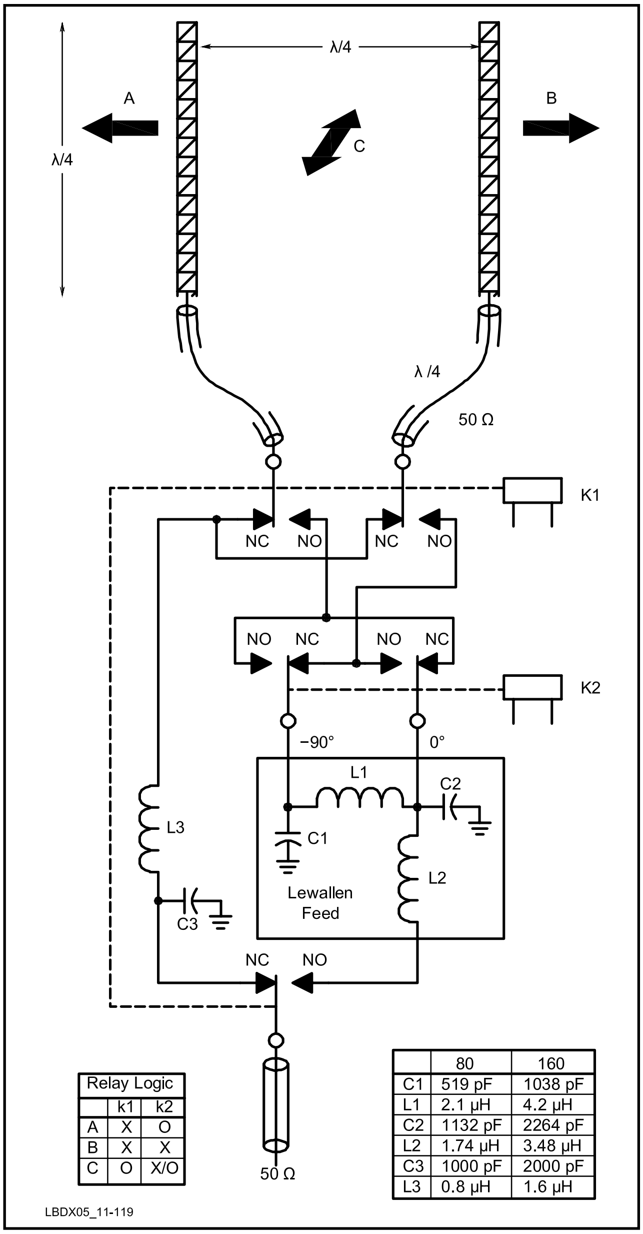
Fig 11-119 shows a direction switching system that includes a bidirectional broadside direction (see Section 4.2).

Two element end-fire arrays are often used in a broadside/end-fire combination to increase directivity and gain (see Section 4.8).

Using the Lewallen feed system one can adjust the L-network component values to obtain the proper feed current magnitude and phase shift, using the simple test method and equipment developed by Robye, W1MK, and described in Section 3.6.

4.1.2.3. Hybrid Coupler Feed, 2-Element End-Fire Array

The non-compensated hybrid coupler is not a good solution for this array, as both k (voltage magnitude ratio) and θ (voltage phase angle) at the ports of the hybrid will be way off (see Fig 11-66).



Relay Logic		
	k1	k2
A	X	O
B	X	X
C	O	X/O

	80	160
C1	519 pF	1038 pF
L1	2.1 µH	4.2 µH
C2	1132 pF	2264 pF
L2	1.74 µH	3.48 µH
C3	1000 pF	2000 pF
L3	0.8 µH	1.6 µH

Fig 11-119 — The 2-element vertical array (λ/4 spacing) can be fed in-phase to cover the broadside directions. The circuit shows the L-network feed system. Relay K1 chooses between the end-fire and the broadside configurations. Relay K2 switches directions in the end-fire position.

You can use one of compensated hybrid feed systems which will give you 300 kHz operational bandwidth on 80 meters. Such a compensated feed system has been calculated in Section 3.4.6.10.1.

The compensated hybrid coupler feed system gives you excellent bandwidth, but you are limited to θ = 90°. The L-network feed system gives you total freedom for θ and k, but you will be limited to an operational bandwidth of approximately 80 kHz.

4.2. The Two-Element Broadside Array

If you feed two elements in phase, they will produce a broadside bidirectional figure eight pattern, provided spacing

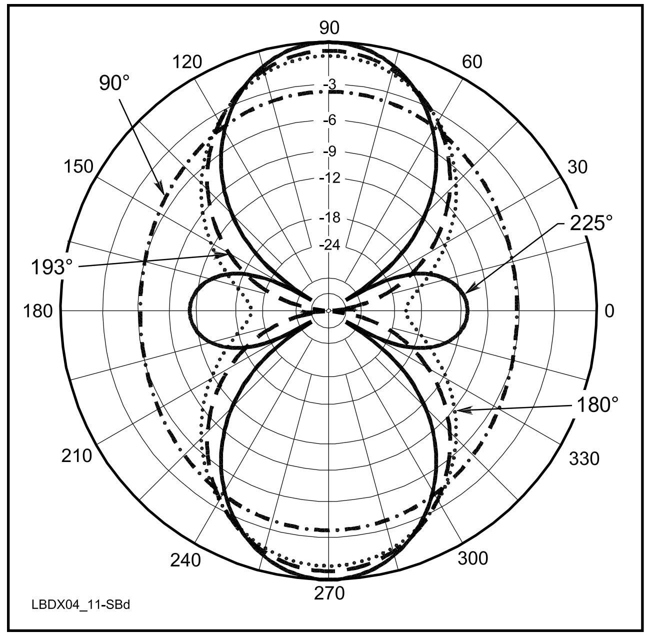


Fig 11-120 — Pattern (at 20° elevation angle) for broadside operation with variable spacing (90°, 180°, 193° and 225 °) between the two elements. Note the sizeable sidelobe that appears for the 225° (5λ/8) case.

is wide enough. (Broadside means radiation in a direction perpendicular to the line connecting the two elements.)

The array with 90° spacing is often used as a "third" direction with an end-fire array, and gives approximately 1 dB gain over a single vertical (see **Fig 11-120**).

4.2.1. Data, 2-Element Broadside Arrays

Narrow spacing yields a wide forward pattern. When we reach λ/2 spacing, and up to about ⅝ λ spacing, the forward lobe is at its narrowest without excessive side lobes. At λ/2 spacing the rejection off the side is maximum at zero wave angle. Increasing the spacing lifts the maximum rejection off the ground, and hence a better directivity and higher gain (by way of narrower forward lobe) are achieved.

Table 11-19 lists the main characteristics of 2-element broadside arrays of various spacings.

4.2.2. Feed Systems, 2-Element Broadside Arrays

As the elements are fed in-phase we can feed them with equal length feed lines to a common point where you parallel the ends of the feed lines. In principle the array can be fed with two feed lines of any equal length. Feeding via λ/4 or 3λ/4 feed lines, however, has the advantage of "forcing" equal currents in both elements, whatever the difference in element impedances might be. Even when using foam dielectric coax (VF = 0.83), two λ/4 feed lines will only allow a element spacing of 0.83 λ (150°), so ¾ λ feed lines will be required for wider spacing if you want to use current-forcing feed lines (which is not strictly necessary).

Using the "Coax Transformer/Smith Chart" and the "Parallel Impedances" modules of the *New Low Band Software* we can easily calculate the feed impedance of this antenna. Let's work out the example of a broadside array with 193° spacing:

Table 11-19
Data for 2-Element Broadside Arrays
For Average Ground (conductivity = 5 mS, ε = 13). Includes 2 Ω equivalent ground loss resistance.
Reference single vertical element gain = 0.34 dBi.

Spacing (°)	*Gain (dBi)*	*3-dB angle (°)*	*Front to side (dB)*	*Z (Ω)*	*EZNEC modeling file*
90	1.87		2	55–*j*16	CH11-2el-broadside-space90.EZ
135	3.14	93	7	40–*j*19	CH11-2el-broadside-space135.EZ
180	4.60	66	48	29–*j*14	CH11-2el-broadside-space180.EZ
193	5.08	59	36	26–*j*11	CH11-2el-broadside-space193.EZ
208	5.40	54.5	17	34.5–*j*8	CH11-2el-broadside-space208.EZ
225	5.60	50.2	11	23.4–*j*4	CH11-2el-broadside-space225.EZ

Table 11-20
Data for 3-Element Broadside Array
For Average Ground (conductivity = 5 mS, ε = 13). Includes 2 Ω equivalent ground loss resistance.
Reference single vertical element gain = 0.34 dBi.
Current magnitude in center element = 2 × current magnitude in outside elements.

Type	*Elem Spacing*	*Gain (dBi)*	*3-dB angle*	*Front to side (dB)*	*Z (Ω) center el*	*Z (Ω) outside el*	*EZNEC modeling file*
3 el	0.5 λ	6.05	47	35	29–*j*14	24.5–*j*20	CH11-3el-broadside-space0.5wve.EZ

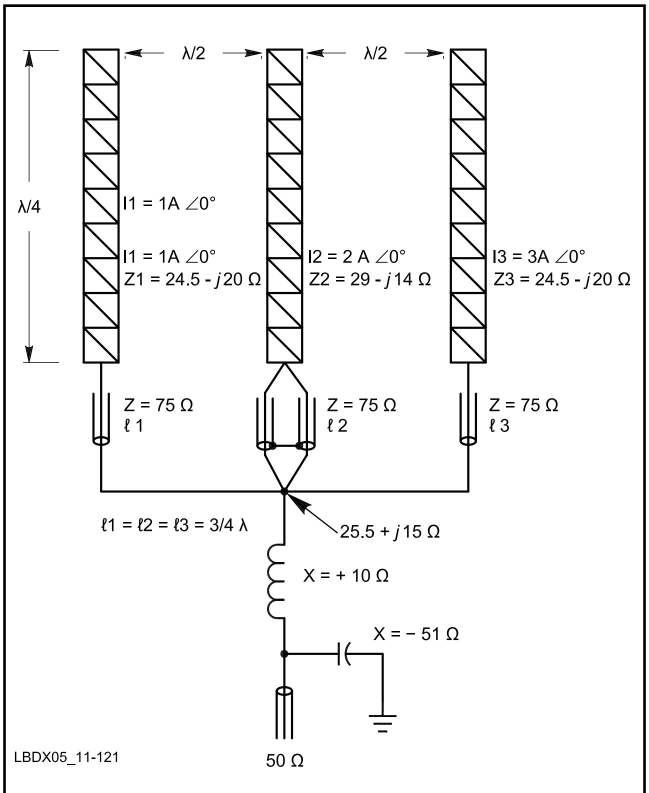


Fig 11-121 — Feed system for the 3-in-line broadside array with using λ/2 spacing and binomial current distribution. The center element is fed via two parallel 75 Ω feed lines in order to obtain double feed current magnitude. The current-forcing method ensures that variations in element self-impedances have minimum impact on the performance of the array.

- Feed impedance: Z = 28 – *j* 12 Ω.
- Assume loss-free cables. At the end of ¾ λ long 50 Ω current-forcing feed lines, impedance is: Z' = 75.4 + *j* 32.3 Ω.
- Paralleling the two feed lines: Z = 38.7 + *j* 6.1 Ω.

Now run the "Shunt/Series Impedance Network Module" and find out that by putting a reactance of –109 Ω (a capacitor) in parallel with this impedance transforms it into 45 Ω, an almost perfect match for the 50-Ω feed line.

4.3 The Three-Element Broadside Array

If more than two elements are used in a broadside combination (all in-line and fed in phase), the current magnitude should taper off toward the outside elements in order to obtain best directivity and gain. This current distribution is what they call the *binomial current distribution*. Multi-element broadside arrays are also covered in Chapter 7 on receiving arrays in Section 1.35.

4.3.1. Array Data

Data is given in **Table 11-20**. The radiation pattern is similar to what is shown in Fig 11-120, only the patterns get narrower and the gain increases as we use more elements.

4.3.2 Feed Systems, 3-Element Broadside Array

If we design the array with 0.5 λ spacing between the elements, our feed lines will need to be ¾ λ long if we want to follow the current-forcing principle. In this case, in order to obtain double feed current magnitude in the center element, we will need to feed the central element with two parallel feed lines. See **Fig 11-121**.

Using 75-Ω coax for the feed lines we have at the end of those feed lines:

Outer elements: Z1' = Z3' = 137.8+ j 112.5 Ω
Center element: Z2' = 39.3 + j 19 Ω

Connected in parallel we obtain an array feed impedance of: 25.5 + j 15 Ω, which we can easily match with an L-network to 50 Ω.

4.4. The Three-Element End-Fire Array

We have covered the 2-element end-fire arrays in Section 4.1. As we have 2 and 3 element Yagis, we also have 2 and 3 element end-fire arrays. As we have seen with 2-element end-fire arrays (Section 4.1) there is nothing sacred about spacing or phase angles. It is true of course that an array with quadrature feeding (phasing angles that are in 90° steps, and identical current magnitudes) have a certain attraction as they make it possible to use the hybrid coupler (Collins) feed system. **Figs 11-122** and **11-123** show radiation patterns for several configurations discussed in this section.

4.4.1 Array Data

The main characteristics of 3-element end-fire arrays with different spacings are listed in **Table 11-21**. Note the negative impedance, which happens frequently in multi-element arrays with the element in the back, especially at close spacings. This simply means that this element is not taking any power from the feed network (through the feed line) but rather delivering excess power it has received through mutual coupling into the feed line.

4.4.2 Hybrid Coupler Feed System, 3-Element End-Fire Array

The 3-element end-fire in-line with 70° equal spacing makes it possible to feed the outer elements with quarter-wave current-forcing feed lines with VF ~ 0.82. That saves a lot of coax! The quadrature feed may not yield the highest gain, but will undoubtedly yield the best operational bandwidth if fed with an optimized hybrid coupler feed system.

EZNEC files *Ch11-3el-endfire-spacing70deg-quadrat.ez* and *Ch11-3el-endfire-spacing70deg-blackbox.ez* contain the description of the model for this array.

The black box principle for this array was explained in

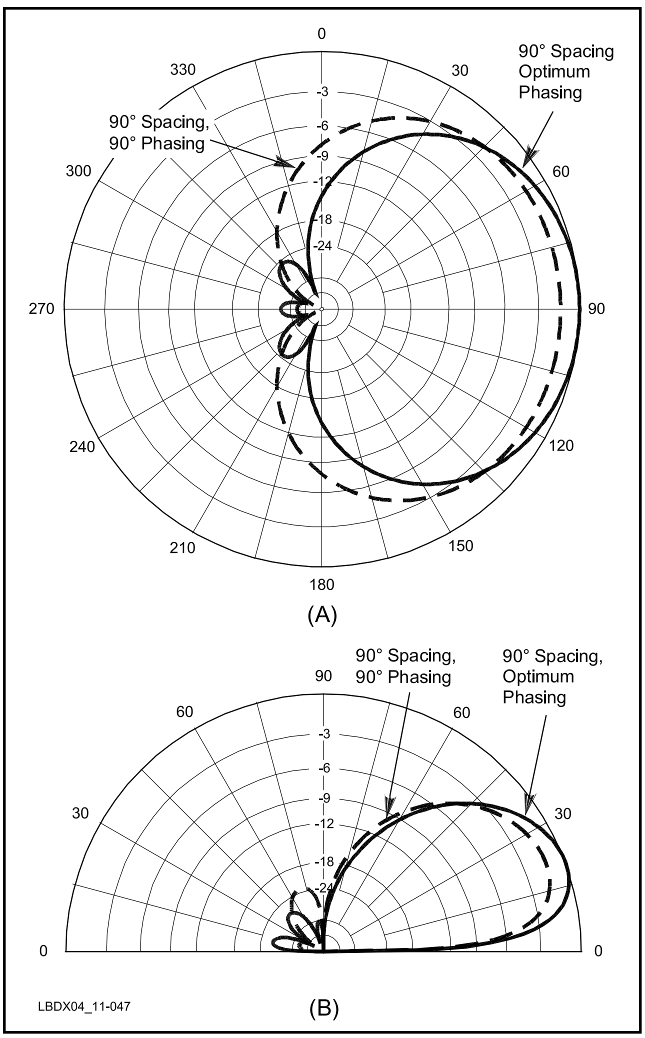


Fig 11-122 — At A, the solid line shows the azimuthal pattern (at 20° elevation) for the quadrature-fed, 3-element in-line end-fire array, with spacings of λ/4. The dashed line is for an array fed with optimized k and θ. At B, elevation pattern comparisons.

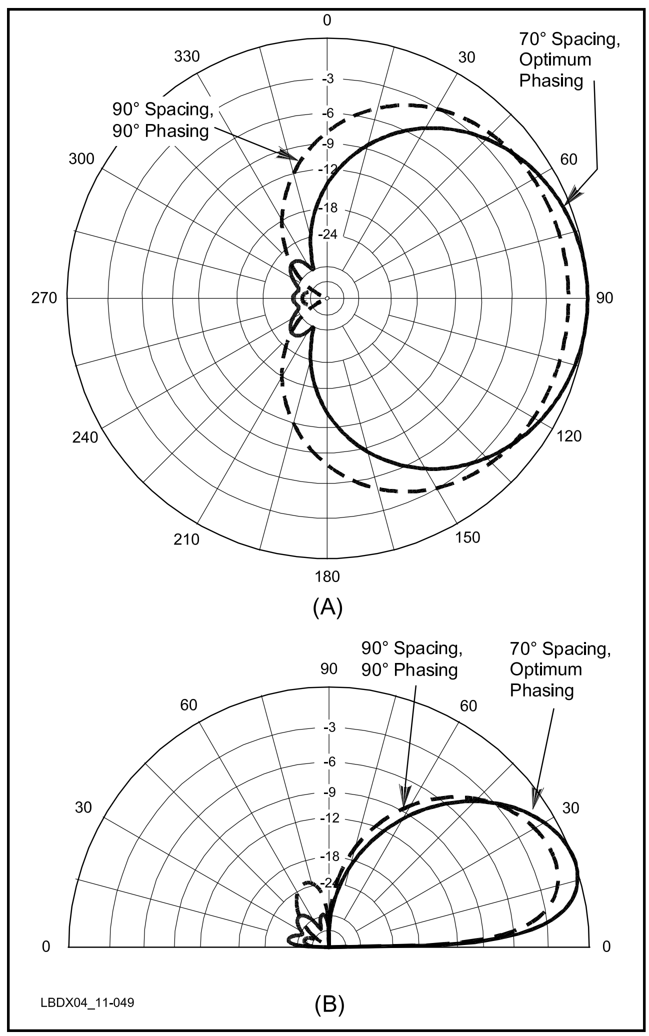


Fig 11-123 — At A, the solid line shows the azimuth pattern (at 20° elevation) for Lahlum/Lewallen feed-optimized array using 70° spacings. The dashed line is a reference with 90° spacings and 90° and 180° phasing. At B, elevation pattern comparisons.

Table 11-21
Data for 3-Element End-Fire Arrays
For Average Ground (conductivity = 5 mS, ε = 13). Includes 2 Ω equivalent ground loss resistance.
Reference single vertical element gain = 0.34 dBi.

Spacing (°)	*Gain (dBi)*	*RDF (dB)*	*DMF (dB)*	*Feed Currents back, mid, front*	*Z (Ω) back, mid, front*	*EZNEC modeling file*
90	5.31	9.27	17.8	1∠0°	14.9–*j*23	CH11-3el-endfire-spacing90deg,V1.EZ
				2∠–90°	36–*j*1	
				1∠–180°	77+*j*50	
90	6.56	10.88	27.9	1∠0°	11.4–*j*14	CH11-3el-endfire-spacing90deg,V2.EZ
				1.75∠–125°	25.6+*j*9.5	
				0.9∠–250°	30+*j*60	
70	6.50	11.17	28.7	1∠0°	8.6–*j*15.8	CH11-3el-endfire-spacing70deg.EZ
				1.85∠–135°	19.2+*j*7	
				0.92∠–270°	–2.3+*j*48	
45	5.12	11.05	27.5	1∠0°	5–*j*15	CH11-3el-endfire-spacing45deg.EZ
				1.9∠–150°	10.5+*j*1	
				0.95∠–300°	–18+*j*11	

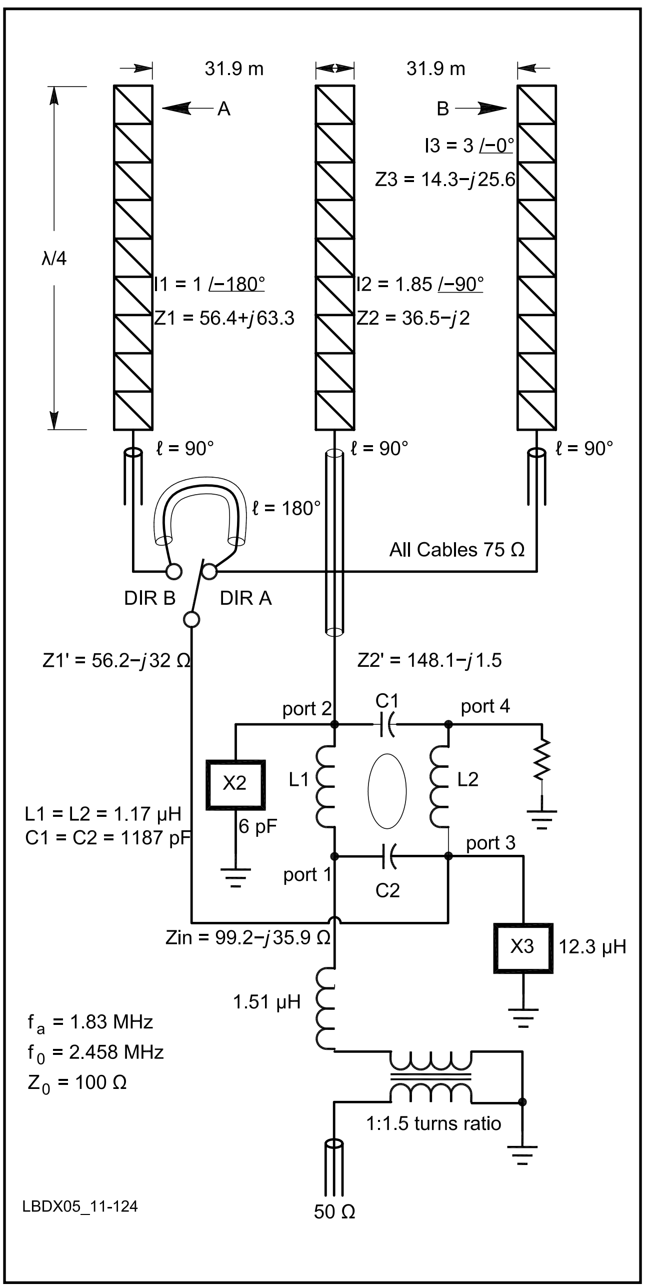


Fig 11-124 — Hybrid coupler feed using W1MK's phase compensation optimization method for the 3-in-line end-fire array using quadrature feeding. The center element is fed via two parallel 75 Ω feed lines to obtain double the feed current magnitude. Direction switching is accomplished with one simple DPDT-relay.

Section 3.4.6.4.1 and both optimized hybrid network feed systems were developed in that section. The operational bandwidth is also discussed in that section.

Fig 11-124 shows the total feed system according to W1MK's two-shunt-element optimized hybrid feed system.

4.4.3. Lahlum-Lewallen System, 3-Element End-Fire Array

The array with 70° spacing between the elements has the advantage of not requiring ¾ λ current-forcing feed lines, if we use coaxial lines with a VF of approximately 0.8. In addition it has substantially (1.3 dB!) more gain, and much better RDF and DMF, but it suffers from much less operational bandwidth (approximately 30 kHz on 160 meters).

Fig 11-125 shows the feed network, including the direction switching done with a DPDT relay (K1). Fig 11-123 shows the horizontal and vertical radiation patterns. Here, 50-Ω feed lines were used as they prevent the components in the L-networks to the front element from having impedances that are too high. A similar network can be calculated for other spacings and phase angles, using the *Lahlum-Lnetwork.xls* spreadsheet tool and the appropriate *New Low Band Software* modules. The procedure to adjust the L-network values is covered in Section 3.4.5. Don't forget that the operational bandwidth of the 3-in-line array fed with L-networks is not more than approximately 80 kHz (vs 300 kHz when fed using the optimized hybrid coupler system).

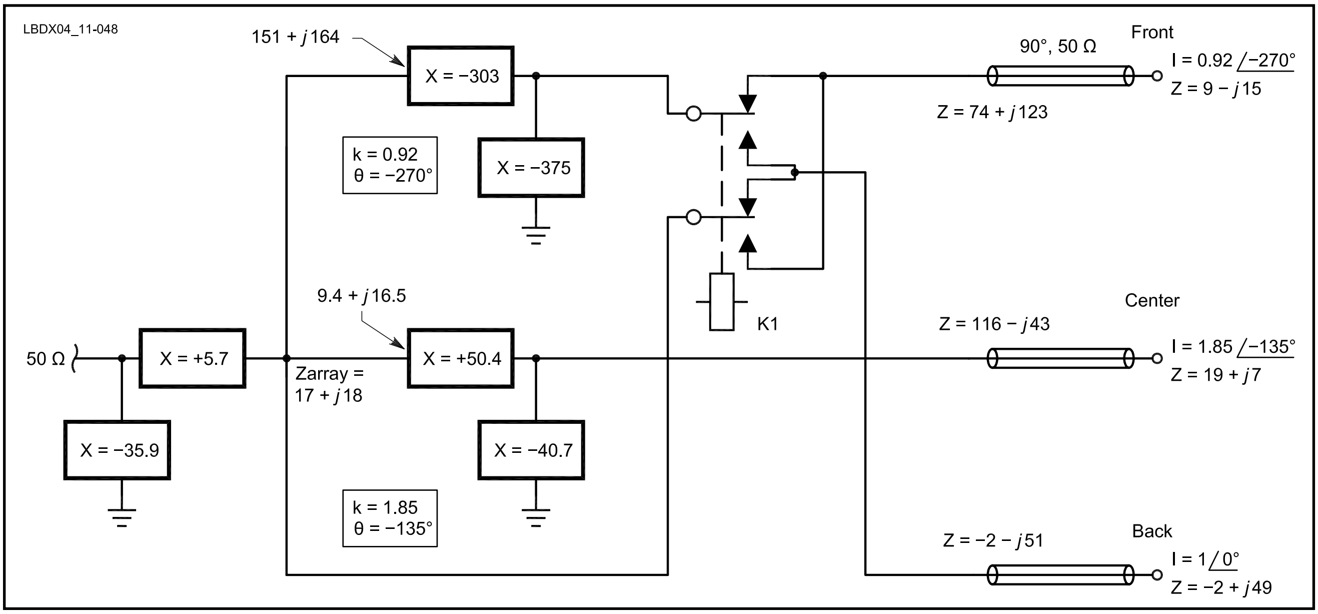


Fig 11-125 — Lahlum/Lewallen feed network for a 3-element in-line, end-fire array with 70° spacing between the elements. This element phasing was chosen to be able to use λ/4 current forcing feed lines (VF = 0.8). Direction switching is included.

4.5. A Bidirectional End-Fire Array

Assume we have a 2-element broadside array with λ/2 spacing. How can we cover the directions that are 90° off? This can be done by feeding the two elements 180° out of phase, which also results in a bidirectional pattern, but with a much broader lobe (beamwidth 115° vs 66° in broadside) and less gain (3.0 dBi vs 4.6 dBi). See **Fig 11-126**.

4.5.1. Array Data

Spacing: λ/2
Feed currents: I1 = 1 A ∠1°; I2 = 1 A ∠–180°
Feed point impedance: Z1 = Z2 = 43.5 + *j* 14.5 Ω
The modeling file (on this book's CD) is:
Ch11-bidirectional-endfire.ez.

4.5.2. Current-Forcing Feed System

We will run a ¾ λ long feed line to the element with the leading current, and a ⁵⁄₄ λ long feed line to the element with the lagging feed current (that's a lot of coax!). With the lines being odd multiples of λ/4 long, we enhance the current-forcing principle (currents will be equal in magnitude even though element impedances may be slightly different). A ¼ λ and a ¾ λ-long feed line are too short for the array, as the elements are spaced λ/2. To preserve symmetry, the T junction, where the lines to the elements join, must be located at the center of the array.

The impedances at the end of the feed lines can be

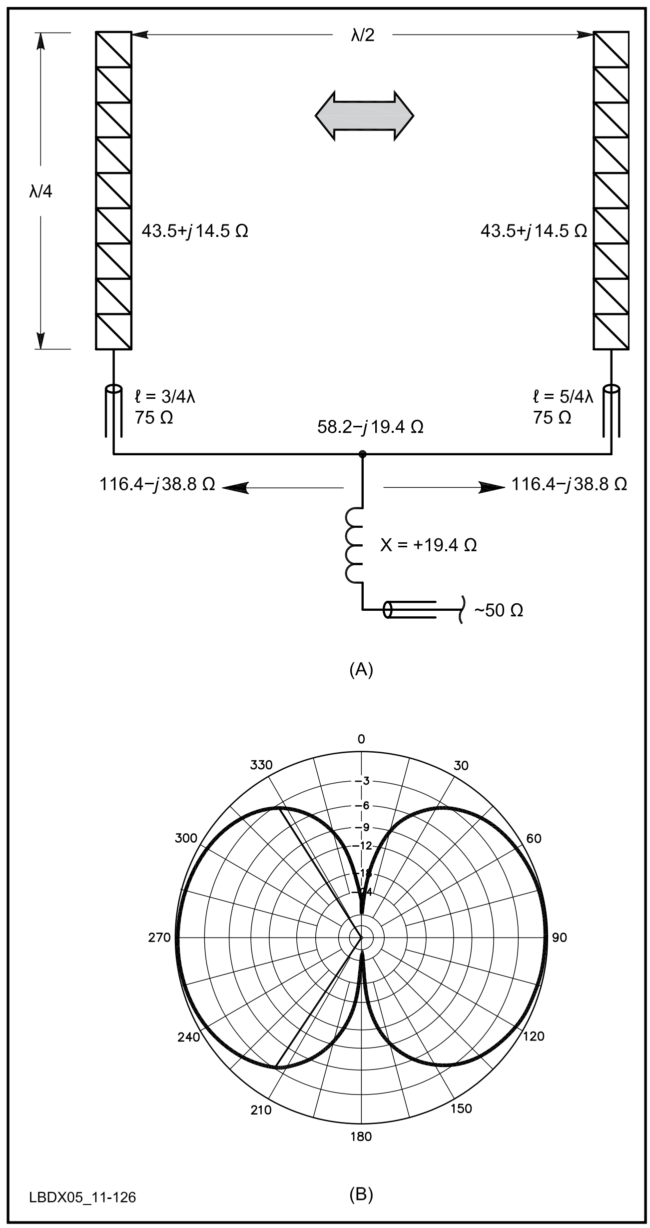


Fig 11-126 — Horizontal radiation pattern (at a 20° elevation) for the 2-element out-of-phase, end-fire array with λ/2 spacing. Elements are in the 90°-270° plane.

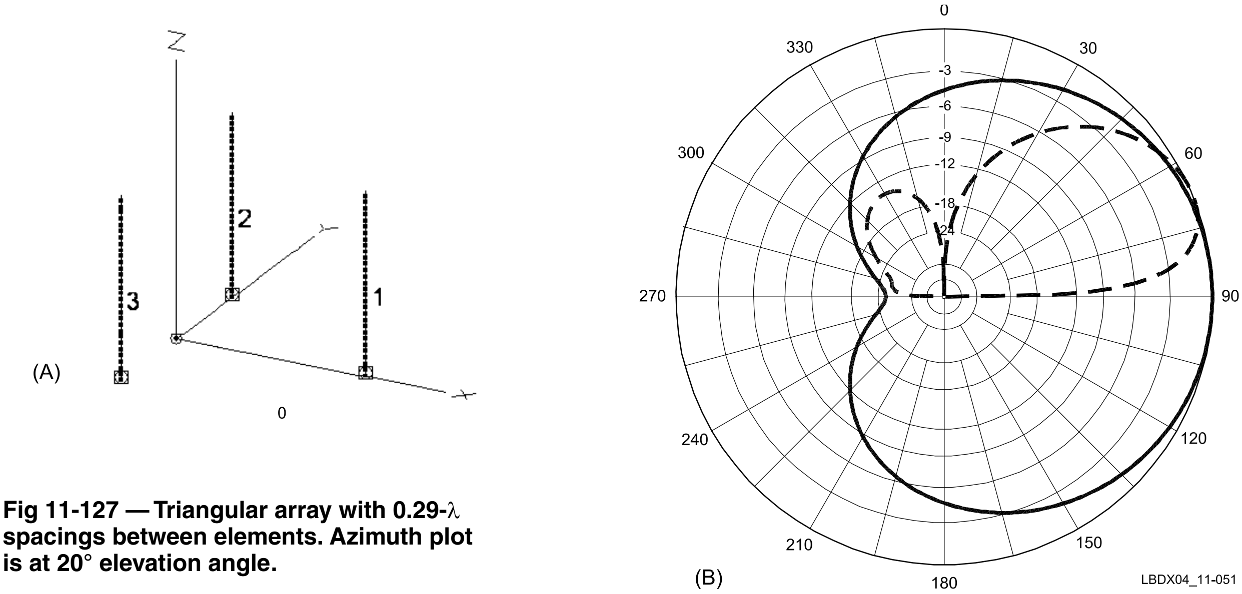


Fig 11-127 — Triangular array with 0.29-λ spacings between elements. Azimuth plot is at 20° elevation angle.

calculated with the "Coax Transformer" software module. While the impedances shown in Fig 11-126 were calculated assuming lossless feed lines, I did a quick check to see how much difference there is with real cable as there is quite a bit of cable involved. Assuming 75 Ω coax with 0.2 dB/100 ft (at 1.83 MHz), we calculate a feed impedance of 54 – *j* 14 Ω, which is very close for all practical purposes. In both cases we can tune out the negative reactance with a small series coil, and end up with a feed impedance very close to 50 Ω.

4.6. Triangular Arrays

The original description by D. Atchley, W1CF (SK), concerned a 3-element array where the verticals are positioned in an equilateral triangle with sides measuring 0.29 λ (Ref 939 and 941). This version of the array used equal current magnitude in all elements. Later, Gehrke, K2BT, improved the array by feeding the two back elements with half the current of the front element. This very significantly improved the directivity of the array.

As expected, the performance (gain, beamwidth, directivity) is somewhere in between the 2-element end-fire array and the Four Square array (see Table 11-26 later in this chapter and **Fig 11-127**).

We can operate a triangle array in two different configurations:

- Beaming off the top of the triangle. The top corner (the "front" element) is fed with a phase delay vs the two bottom line verticals which are fed with the reference phase angle (0°).
- Beaming off the bottom of the triangle. In this case the bottom-corner elements are fed by the current with a phase delay vs the top vertical (the "back" element) which is fed with the reference phase angle of 0°.

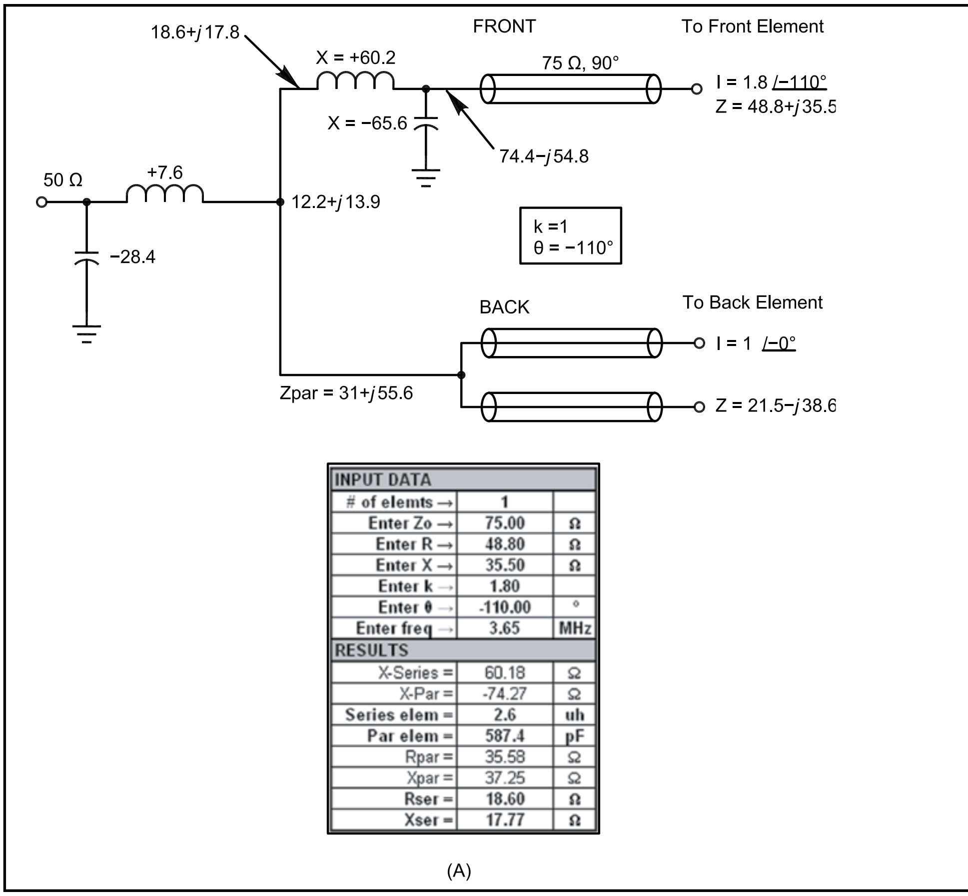
This means that a triangular array can be made switchable in six directions. All directions have the same gain (within approximately 0.1 dB) and a very similar radiation pattern.

We have now fine-tuned this array for best directivity for a quadrature fed configuration, as well as for an optimized configuration where the phase shift is 110° and the current magnitude ratio 1.8. The optimum dimensions were also computed. For the quadrature-fed array the optimum triangle

Table 11-22
Data for Triangular Arrays
For Average Ground (conductivity = 5 mS, ε = 13). Includes 2 Ω equivalent ground loss resistance.
Reference single vertical element gain = 0.34 dBi.
Configuration A: firing through the top of the triangle; B: firing through the baseline

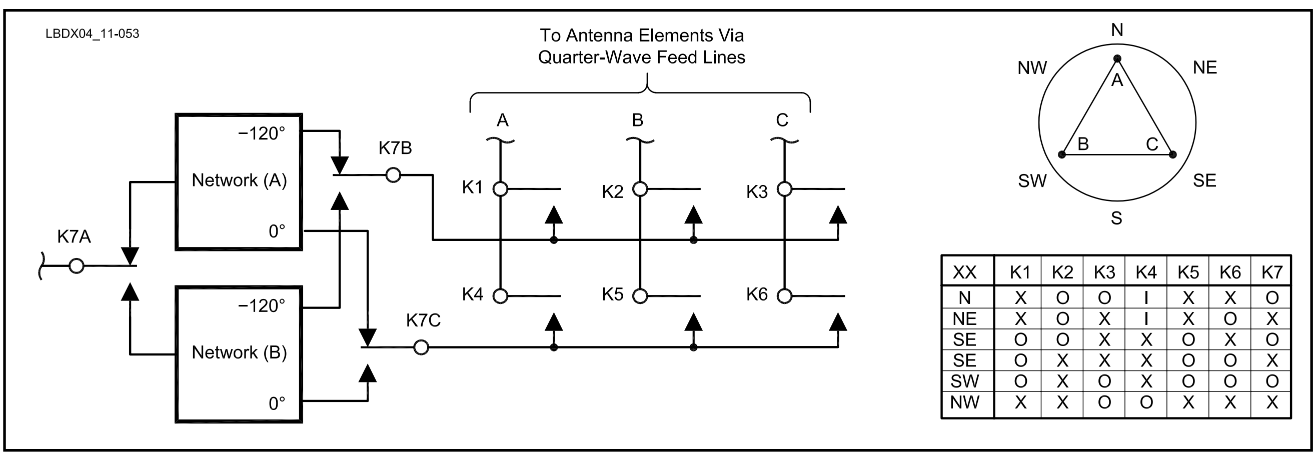
Side (λ)	*Config*	*Gain (dBi)*	*3-dB Beamwidth*	*F/B (dB)*	*RDF (dB)*	*DMF (dB)*	*Feed Currents*	*Impedances (Ω)*
0.307	A	4.32	147°	29.8	8.76	13.8	2∠–90°	28.4+*j*12.4
							1∠0°	16–*j*42.1
							1∠0°	16–*j*42.1
0.307	B	4.42	143°	38	8.87	14.4	2∠0°	22.2–*j*11.7
							1∠–90°	88.5+*j*4.6
							1∠–90°	88.5+*j*4.6
0.246	A	4.59	137°	32.5	9.16	14.4	1.8∠–110°	48.8+*j*35.5
							1∠0°	21.5–*j*38.6
							1∠0°	21.5–*j*38.6
0.246	B	4.68	133°	21.8	9.25	16.1	1.8∠0°	16.8–*j*14.5
							1∠–110°	73.3+*j*26.2
							1∠–110°	73.3+*j*26.2

Fig 11-128 — At A, the feed system for the triangle array when firing off the top of the triangle. At B, the feed system when firing off the baseline of the triangle. If you want six directions, you will need a switching system that selects the proper network, as shown in Fig 11-129. L-networks to match to the 50 Ω feed line are included.



INPUT DATA		
# of elemts →	1	
Enter Zo →	75.00	Ω
Enter R →	48.80	Ω
Enter X →	35.50	Ω
Enter k →	1.80	
Enter θ →	-110.00	°
Enter freq →	3.65	MHz
RESULTS		
X-Series =	60.18	Ω
X-Par =	-74.27	Ω
Series elem =	2.6	uh
Par elem =	587.4	pF
Rpar =	35.58	Ω
Xpar =	37.25	Ω
Rser =	18.60	Ω
Xser =	17.77	Ω

Fig 11-129 — Seven relays, of which six are SPST relays in a matrix, are used to make a six direction switching network/feed system. The networks are shown in Fig 11-128.



XX	K1	K2	K3	K4	K5	K6	K7
N	X	O	O	I	X	X	O
NE	X	O	X	I	X	O	X
SE	O	O	X	X	O	X	O
SE	O	X	X	X	O	O	X
SW	O	X	O	X	O	O	O
NW	X	X	O	O	X	X	X

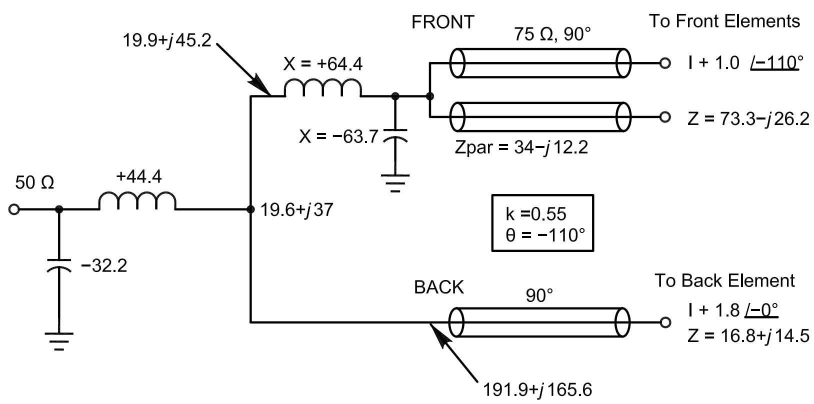
side dimension is 0.307 λ. For the optimized triangle with θ = 110° and k = 1.8, the ideal triangle side dimension is 0.246 λ.

It is obvious that the element feed impedances are different depending on which configuration you use (top firing or base firing). The feed impedances are shown in **Table 11-22**.

4.6.1. Feeding the Triangular Arrays

There are several good reasons why you should not attempt using a hybrid coupler for this array. If you want to get good performance, build two L-networks, one to be used when firing off the top of the triangle, and the other one for firing through the base of the triangle. As the feed impedances are different for the "A" and for the "B" directions, we need two separate phasing networks. In this case you should opt for the current optimized version, which achieves better directivity and gain. Using a Lewallen L-network feed system, the higher current required for the solitary element can be achieved by simply dimensioning the L-network components correctly (specify k accordingly).

Fig 11-128 shows the Lahlum/Lewallen feed networks for both triangle configurations. In **Fig 11-129** we see the direction switching for the array. To do the direction switching, a small matrix of six SPST relays plus a seventh relay with three inverting contacts is required. This may seem complicated, but using the two L-networks makes it possible to adjust the values in order to obtain the exact feed currents as modeled. One of the measuring setups described in Section 3.6 can be used to make the adjustments.



INPUT DATA		
# of elemts →	2	
Enter Zo →	75.00	Ω
Enter R →	73.30	Ω
Enter X →	26.20	Ω
Enter k →	0.56	
Enter θ →	-110.00	°
Enter freq →	3.65	MHz
RESULTS		
X-Series =	64.39	Ω
X-Par =	-63.69	Ω
Series elem =	2.8	uh
Par elem =	685.0	pF
Rpar =	122.35	Ω
Xpar =	54.04	Ω
Rser =	19.97	Ω
Xser =	45.22	Ω

LBDX05_11-128

(B)

4.7. The Four Square Array

In 1965, D. Atchley (then W1HKK, later W1CF, now SK), described two arrays that were computer modeled and later built and tested with good success (Ref 930, 941). Although the theoretical benefits of the Four Square were well understood, it took a while before the correct feed methods were developed that could guarantee performance on a par with the paperwork.

Feed systems based on a 90° hybrid coupler have until now been most popular — not because they give the results, but because the system is supposed to be "plug and play." Unfortunately, until now both the theory of operation and performance of hybrid coupler feed systems have been covered by a veil of mystery and half truths.

To my knowledge, for the first time in Amateur Radio literature, the full story of the hybrid coupler has been told in great detail in this book (Section 3.4.6). The truth is that you cannot just use a plug-and-play hybrid coupler, designed to work correctly with 50 Ω loads, and expect it to produce ideal results when you have the feed lines to a Four Square array connected to it. It requires more work to achieve that. And that is what the hybrid feed system optimization methods, covered in detail in Sections 3.4.6.5, 3.4.6.6 and 3.4.6.7) are all about.

The Four Square is in fact similar to a 3-in-line end-fire array — the center two elements are fed in phase and act as one common element. If all four elements have equal current, total center element current (for both in-phase elements together) is twice the current in each end. The required 1:2:1 current distribution as explained in Section 4.4 is satisfied.

The Four Square can be switched in four quadrants. Atchley also developed a switching arrangement that made it possible to switch the array directivity in increments of 45°. The second configuration consists of two side by side cardioid arrays. This antenna is discussed in detail in Section 4.8.

4.7.1. Quarter-Wave-Spaced Square, Quadrature-Fed

The array elements are placed in a square, spaced λ/4 per side. All elements are fed with equal current magnitude. The back element is fed with the reference feed current angle of 0°, the two center elements with –90° phase, and the front element with –180° phase difference. All currents are of equal magnitude.

The direction of maximum signal is along the diagonal from the rear to the front element (an array always radiates in the direction of the element with the lagging current).

4.7.1.1. Array Data

Dimension of square side: λ/4

Feed currents:

I1 = 1 ∠–180° (front element)

I2 = I4 = 1 ∠–90° (center elements)

I3 = 1 ∠0° (back element)

Ground: ρ = 5 mS, ε = 13

Radial loss: 2 Ω

Gain: 5.75 dBi

3-dB beamwidth: 100°

RDF = 10.52 dB

DMF = 21.02 dB

Feed-point impedances:

Z1 = 61.2 + *j* 56.1 Ω

Z2 = Z4 = 42.8 – *j* 19.3 Ω

Z3 = 0.6 – *j* 17.2 Ω

Modeling file: *Ch11-4sq-0-90-180-1-1-1-1.ez.*

Fig 11-130 shows the radiation patterns for the quadrature-fed Four Square array. Note the large high angle back lobe peaking at a 60° wave angle, where the F/B is only approximately 18 dB.

4.7.1.2. Feeding the Quadrature-Fed Four Square with the Hybrid Coupler Feed System

As the antenna is fed in quadrature, a hybrid feed system (see Section 3.4.6.) is possible, and is the indicated system provided one of the optimization systems is applied as described in Sections 3.4.6.5 and 3.4.6.6. All the details and step-by-step procedures for these optimized hybrid feed

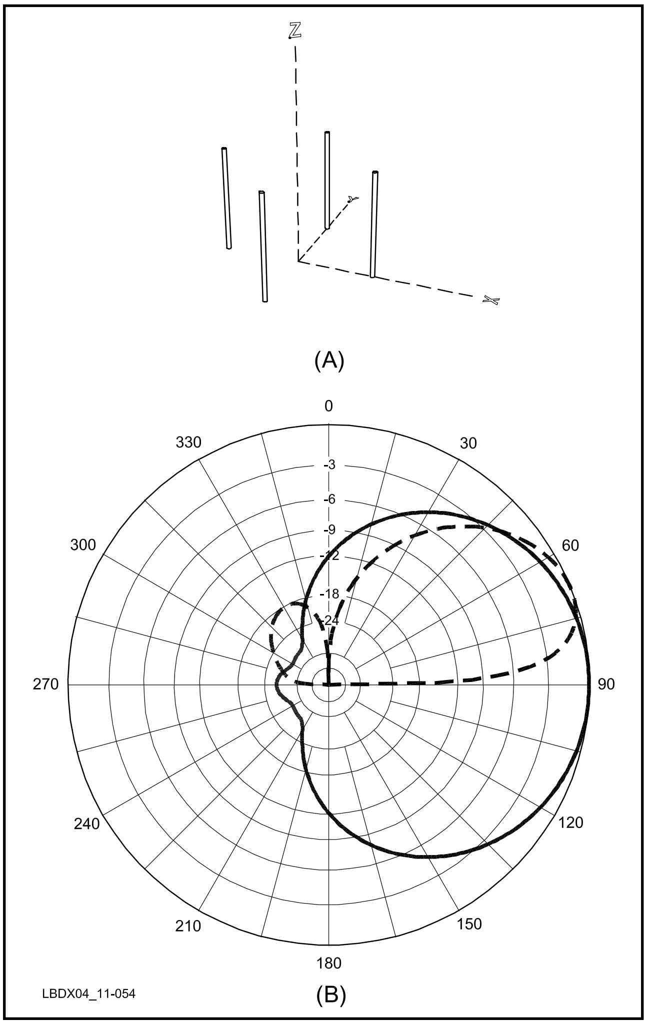


Fig 11-130 — Radiation patterns (horizontal at a 20° elevation angle) for a typical quadrature-fed Four Square array. Notice the important back lobe at relatively high elevation angles (about 60°).

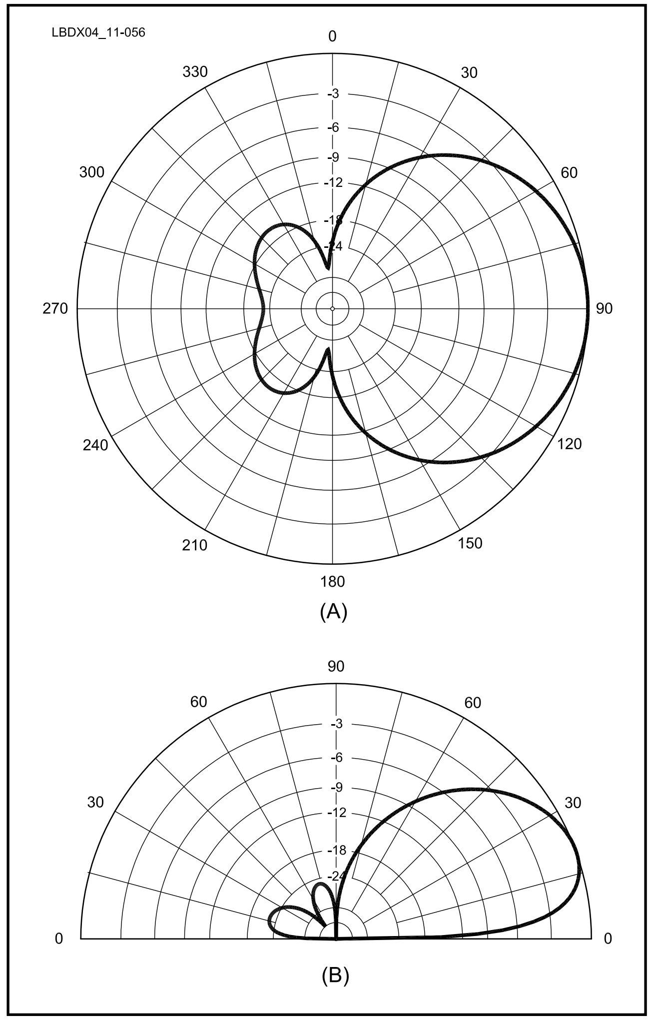


Fig 11-131 — Radiation patterns for the WA3FET-optimized Four Square, where the high-angle back lobe has been reduced substantially. Net result is 0.7 dB more gain and increased directivity.

systems are covered in these sections.

Note that this optimized hybrid coupler feed system also can be used for a Four Square configuration where the center element is fed with a current magnitude different from what we have in the front and the back element (Sections 3.4.6.5.10 and 3.4.6.6.7).

Remember that this system will give you abut 300 kHz of operational bandwidth on 80 meters (F/B >20 dB in that range). None of the other feed systems can reach this excellent broadband performance.

4.7.1.3. Feeding the Quadrature-Fed Four Square with the Lewallen Feed Method

Section 3.4.5.6 covers the detailed calculation of the Lewallen feed system (LC-network) using the *Lahlum-Lnetwork.xls* spreadsheet. The operational bandwidth for this feed system is not more than approximately 80 kHz on 80 meters, which is quite a difference from what can be achieved with the optimized hybrid coupler systems.

The Lewallen feed method for this array is also worked out in great detail in *The ARRL Antenna Book*, and L-network values are listed for a range of feed-line impedances and ground systems.

4.7.1.4. Opposite Voltage Feed System (OH1TV)

This novel feed system is described in detail in Section 3.4.9 and several Four Square applications are covered more in detail in Section 4.7.6. Note that this system has a much better bandwidth than the L-network system (typically 200 kHz vs 80 kHz), but cannot beat the bandwidth performance of the optimized hybrid coupler system.

4.7.2. The WA3FET Optimized Four Square Array

Jim Breakall, WA3FET, optimized the quarter-wave-spaced Four Square array to obtain higher gain (0.6 dB increase) and better directivity.

In Fig 11-130 we saw that the original Four Square exhibits a very major high-angle back lobe (–15 dB). By changing the feed current magnitude and angle to the various elements one can change the size and the shape of the back lobes as well as the width of the front lobe (**Fig 11-131**). Full optimization is a compromise between optimization in the elevation and the azimuth planes. With Breakall's optimization, the gain of the array went up by 0.54 dB. At least as important is a significant gain in directivity (RDF and DMF).

4.7.2.1. Array Data

Dimension of square side: λ/4
Feed currents:
  I1 = 0.872 ∠–218° (front element)
  I2 = I4 = 0.9 ∠–111° (center elements)
  I4 = 1 ∠0° (back element)
Gain: 6.29 dBi
3-dB beamwidth: 87°
RDF = 11.4 dB
DMF = 24.4 dB
Calculated feed-point impedances:
  Z1 = 36.6 + *j* 58.7 Ω (front element)
  Z2 = Z3 = 32 – *j* 7 Ω (center elements)
  Z4 = 5.8– *j* 4.5 Ω (back element)
Modeling file: *Ch11-4sq-WA3FET.ez.*

4.7.2.2. Feeding the WA3FET Four Square with the Lahlum/Lewallen Feed System

In Section 3.4.5.8 we covered in detail the design of the Lahlum/Lewallen feed system for this array. Note that in Section 3.4.5.8 we lengthened all array elements an equal amount to obtain a non-reactive impedance in the center two elements, which results in slightly different component values and impedances than those shown in **Fig 11-132**.

4.7.3. The 1/8-λ Spaced Four Square

4.7.3.1. Array Data

Using λ/8 as the side dimensions of our Four Square, we found the best directivity (assuming equivalent radial loss resistance = 2 Ω, ρ = 5 mS and ε = 13) using the following feed currents in our model (*Ch11-4sq-one-eighth-wave-spacing.ez)*:
  I1 = 1 ∠–270° (front element)
  I2 = I3 = 1 ∠–135° (center elements)
  I4 = 1.15 ∠0° (back element)
Gain: 4.91 dBi
3-dB beamwidth: 88°
RDF = 11.4 dB
DMF = 25.0 dB
Feed-point impedances:
  Z1 = 3.4 – *j* 12.3 Ω (front element)
  Z2 = Z3 = 17.8 – *j* 3.5 Ω (center elements)
  Z4 = –13 + *j* 16.6 Ω (back element)

This small footprint Four Square trades in 1.3 dB of gain compared to its optimized big brother, but it has every bit as good or even better directivity (**Fig 11-133**).

The main disadvantage of this design is the much narrower bandwidth. Note that the reduction in gain is to a large extent due to the lower impedances of the elements, taking into account that we inserted an equivalent radial loss resistance of 2 Ω at the base of all elements.

Running the modeling file for this array, one can notice that the slightest change in k or θ has a great influence on the directivity.

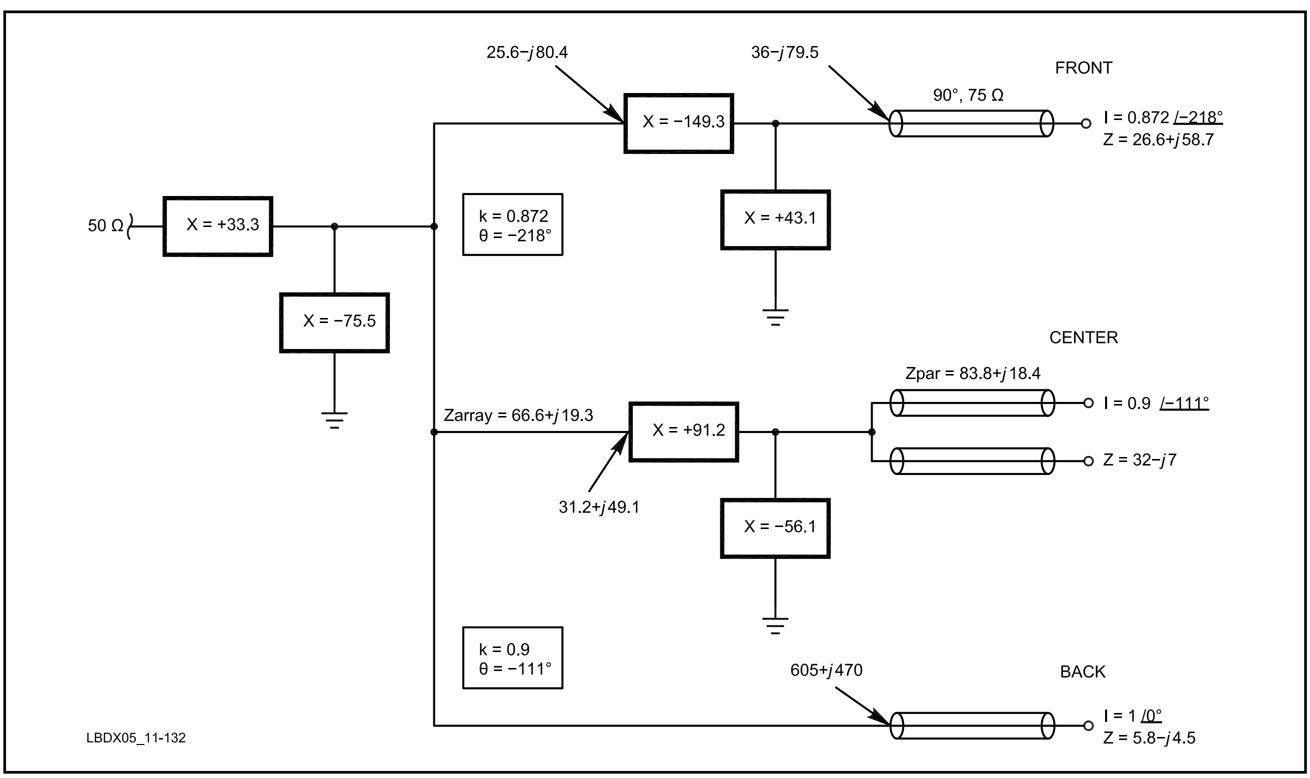


Fig 11-132 — Lahlum/Lewallen feed circuit for the WA3FET style Four Square, with optimized phase angle and drive current magnitudes. In Fig 11-25 slightly different element impedances were used. Note that the variation of the L-network components are well within normal "tuning range." The feed impedances are also within a few percent of one another.

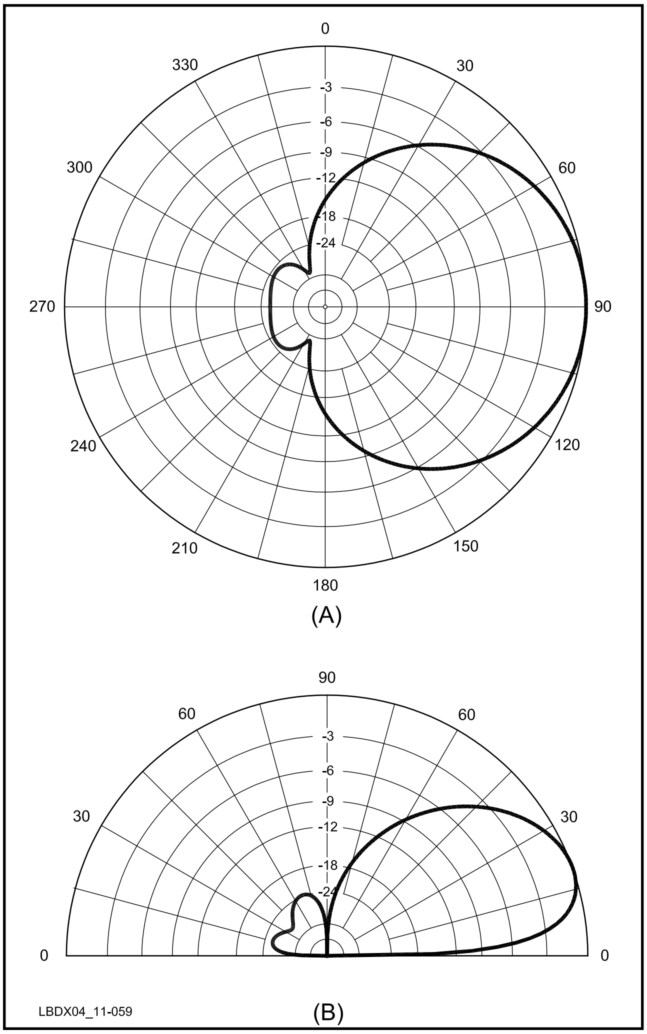


Fig 11-133 — Radiation patterns for the reduced size (λ/8 side) Four Square, which exhibits even better directivity than the larger varieties, but with slightly less gain and less bandwidth.

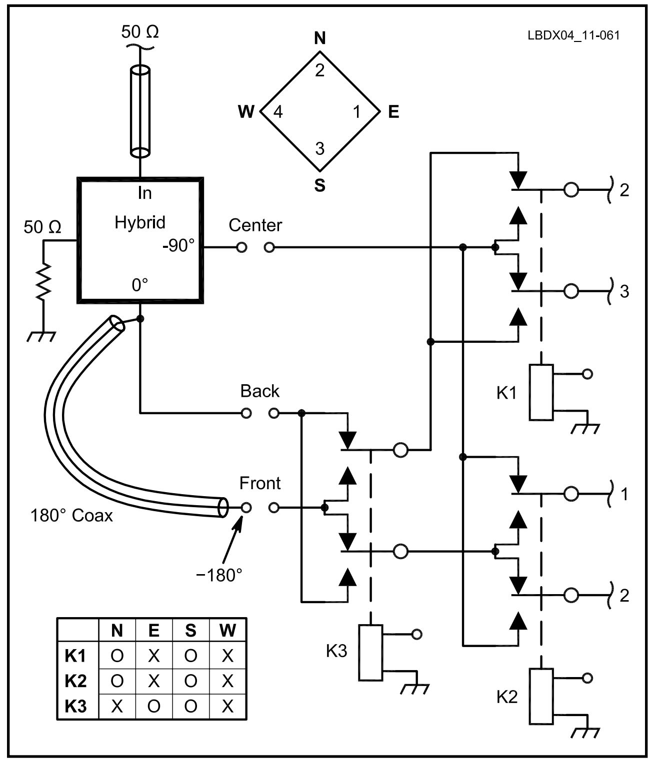
All of this means that such reduced-size arrays should really only be considered if a very good ground system can be installed (or over saltwater), and if a narrow operational bandwidth is sufficient.

4.7.3.2. Feeding the "Small" Four Square

Because of the very low impedances involved, feeding this array is tricky and at best the bandwidth will be narrow. For those reasons I have chosen to no longer publish a feed network. In any case, the feed network for this array would be more theoretical than practical, and very difficult to adjust.

4.7.4. Direction Switching for the Four Square Arrays

Fig 11-134 shows a direction-switching system that can be used with all the Four Square arrays. The "front" element (in the direction of firing) will of course be fed with the most lagging feed angle (–180° in case of quadrature feeding) and the back element with the zero (reference) feed angle. "Center," Back" and "Front" go to the corresponding points in the feed circuits in case of a Lahlum-Lewallen circuit,



	N	E	S	W
K1	O	X	O	X
K2	O	X	O	X
K3	X	O	O	X

Fig 11-134 — The direction switching system shown can be used with all Four Square arrays, whatever phasing circuitry is used. The 180° phasing line can of course be replaced by a 1:1 phase inverter transformer.

In case of a quadrature-fed array, a hybrid can be used, wired as shown in the figure. The Comtek and the DX Engineering hybrid couplers include a 180° phase-inversion transformer which replaces the half-wave (180°) coaxial line.

4.7.5. Opposite Voltage (Voltage-Forcing) Feed System for the Four Square Array

This novel array feed system is an attractive candidate for feeding a typical Four Square array. With the system you can use feed conditions that deviate considerably from the quadrature condition.

As an example we have designed five different Four Square

Table 11-23
Data for Four Square Arrays Using Opposite Voltage Feed System

	1∠0, 1∠–0, 1∠–180	*1∠0, 1∠–103, 1∠–260*	*1∠0, 1∠–120, 1∠–240*	*1∠0, 0.9∠–111, 0.879∠–218*	*1∠0, 0.85∠–90, 1∠–180*
Gain (dBi)	5.22	5.49	5.5	5.55	6.24
F/B (dB)	25.7	29.6	32.1	26.7	27.4
RDF (dB)	10.76	11.34	12.01	11.61	10.84
A	2.62 µH	1.91 µH	1.16 µH	1.44 µH	2.10 µH
B	0.82 µH	0.79 µH	0.49 µH	0.53 µH	0.58 µH
A'	1.84 µH	2.17 µH	2.42 µH	2.35 µH	1.39 µH
B'	1022 pF	1141 pF	1367 pF	1139 pF	1247 pF
match L	0.81 µH	0.67 µH	0.57 µH	0.66 µH	0.85 µH
match C	1600 pF	1887 pF	2384 pF	1965 pF	1749 pF

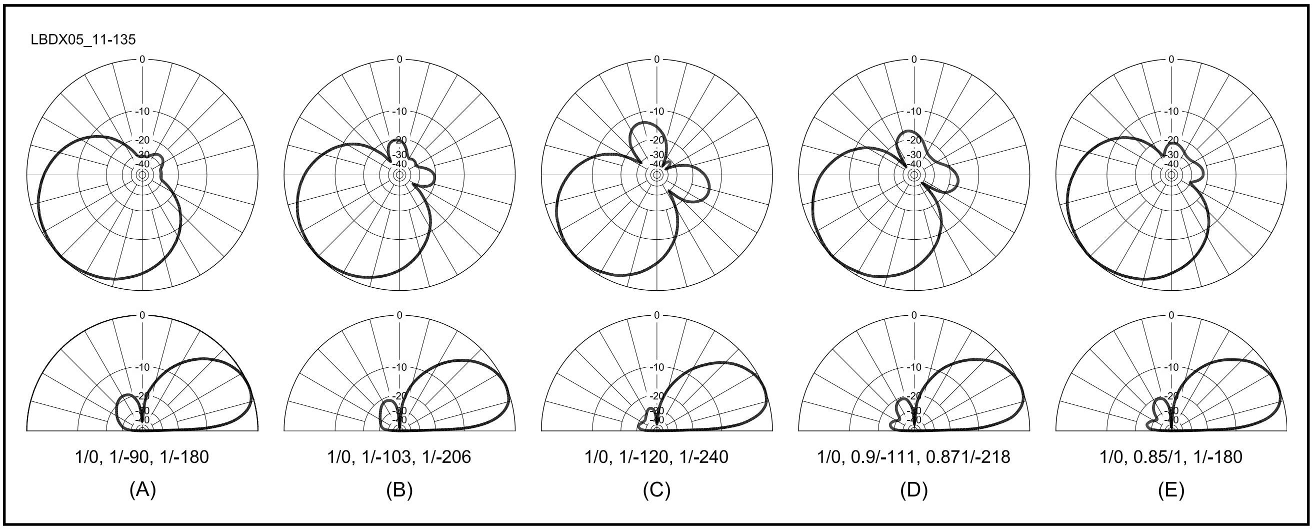


Fig 11-135 — Horizontal and vertical radiation pattern (at design frequency) for five different Four Square arrays, using loading elements (and L-network) calculated with the *4-sq-voltagefeed-calculator.xls*. The modeling files are available on this book's CD.

arrays, all fed with the opposite voltage feed system. **Table 11-23** shows the calculated load element values (procedure outlined in Section 3.4.9) as well as the performance results (gain, F/B and RDF). This exercise proves that with this system we can develop feed systems with variable k and θ specifications. **Fig 11-135** shows the vertical and horizontal patterns for these five designs (all on their design frequency, 3.65 MHz).

The next aspect to be evaluated is the operational bandwidth. I calculated the performance data vs frequency for three different designs: pure quadrature, quadrature with center element k = 0.85, and the famous WA3FET configuration. The results are listed in **Table 11-24**

With all three designs we obtain an operational bandwidth between 150 and 200 kHz, obviously limited by our 20 dB minimum F/B requirement. This is almost 2.5 times wider than what we can obtain with the classic L-network (Lewallen) feed system, but approximately 30% to 50% less than the operational bandwidth delivered by the optimized hybrid feed system. Note also that the feed system delivers a very flat SWR curve.

Perhaps the main disadvantage of the system is that it requires four half-wave long feed lines, which is quite a bit of cable. I would strongly warn against using cheap cable, as we do want to keep the cable losses as low as possible. In the different modeling files on this book's CD (*Ch11-4sq-voltage-feed.ez*, *Ch11-4sq-volt-90-180.ez, Ch11-4sq-volt-0.85_-90-1-180.ez* and *Ch11-4sq-volt-WA3FET.ez*) you can specify the losses of those cables. A little modeling exercise tells us that going from RG-213 cable to ½ inch hard line saves about 0.5 dB in total antenna gain, which is considerable.

Direction Switching

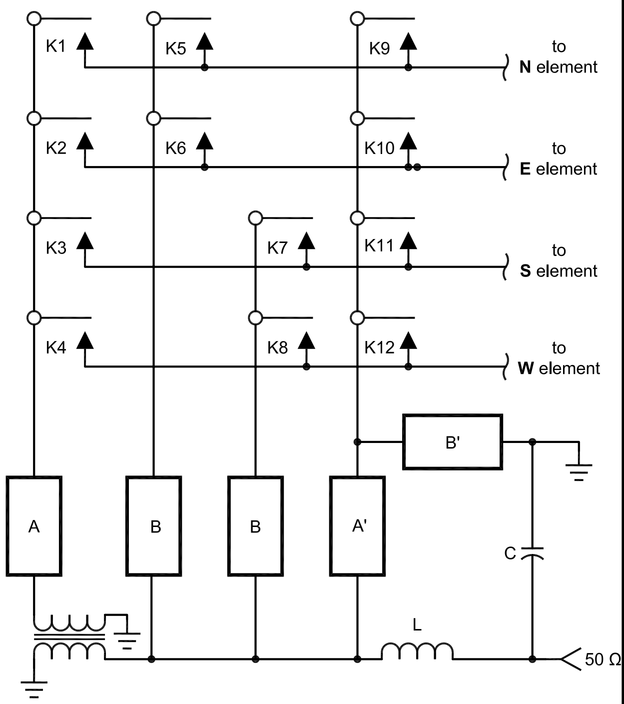
Single pole, normally open contact (SPNO) relays are ideal for constructing a switching matrix to switch this antenna. **Figs 11-136** and **11-137** show a matrix of 3 × 4 SPNO 30 A relays in an antenna switching matrix at ON4UN. Vacuum relays as shown in Fig 11-150 later in this chapter, can of course also be used.

The values of A, B, A' and B' are calculated using the *4sq-voltagefeed-calculater.xls* spreadsheet. L and C make in input 50 Ω matching L-network. The 180° phase-inversion transformer (see Section 3.4.6.3) can also be replaced by another λ/2 feed line. **Fig 11-138** shows OH1NM's array built with the opposite voltage feed system.

Table 11-24
Operational Bandwidth for 80 Meter Four Square Arrays

In all cases, the F/B limited bandwidth is between 150 and 200 kHz.

	Freq (MHz)	*3.5*	*3.55*	*3.6*	*3.65*	*3.7*	*3.75*	*3.8*
Quadrature	Gain (dBi)	5.17	5.27	5.27	5.22	5.13	5.01	4.91
	F/B (dB)	19.2	24.5	37.5	25.7	20.2	17.3	15.5
	RDF (dB)	11.25	11.07	10.91	10.76	10.62	10.49	10.38
	SWR	1.16	1.05	1.00	1.00	1.10	1.20	1.35
1∠0, 0.85∠–90	Gain (dBi)	5.13	5.27	5.29	5.24	5.16	5.04	4.90
1∠–180	F/B(dB)	15.7	17.3	21.0	27.4	28.1	22.4	19.0
	RDF (dB)	11.46	11.19	11.01	10.88	10.59	10.52	9.67
	SWR	1.20	1.10	1.05	1.00	1.05	1.15	1.30
WA3FET	Gain (dBi)	4.57	5.10	1.25	5.54	5.57	5.51	5.38
	F/B (dB)	11.80	14.80	19.30	26.60	27.10	20.90	17.20
	RDF (dB)	11.46	11.19	11.01	10.88	10.59	10.52	9.67
	SWR	1.20	1.10	1.05	1.00	1.05	1.15	1.30



	K1	K2	K3	K4	K5	K6	K7	K8	K9	K10	K11	K12
N	X					X		X			X	
E		X			X		X					X
S			X			X		X	X			
W				X	X		X			X		

LBDX05_11-136

Fig 11-136 — Opposite voltage feed system including the matrix relay direction switching system.

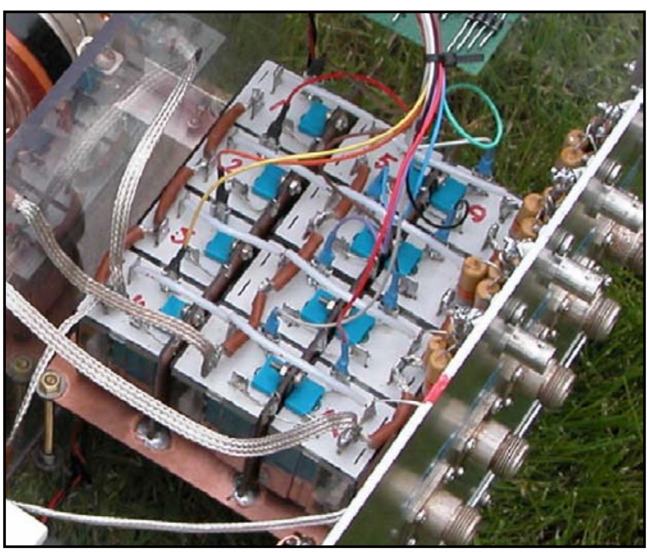


Fig 11-137 — Switching matrix (3 by 4) with standard open-frame 30-A single pole, normally open relays, used for direction switching of the Four Square array at ON4UN. (See also Fig 11-97.)

Fig 11-138 — Two-element 80 meter end-fire array at OH1NM, fed according the opposite voltage feed system. This is not the shore of a lake, but the shore of the Baltic Sea.

Conclusion

The opposite voltage feed system is not as broadband as the optimized 90° hybrid system. Its advantage, though, is that it can provide phasing angles deviating from 90° increments.

4.8. Four Square Array with Eight Directions

With a Four Square having a –3 dB forward lobe beamwidth of 85 to 100° (depending on spacing and phasing), four directions seem to quite adequately cover the azimuth. Right in between two adjacent main directions you can lose from 2.5 to 3 dB in signal strength, though. Adding an intermediate direction, covered by two side-by-side end-fire arrays does not give you back these 3 dB as the side-by-side end-fire array has a gain that is almost 1.5 dB lower than for the Four Square that covers the main directions. This means that right in between the main direction you can gain approximately 1.5 dB. Looking at this issue from a statistical point of view, on average the gain will be about half that figure, which means less than 1 dB. Is this worth the effort?

Admittedly, more than four directions may be advantageous because the nulls in the back can be moved around to null out QRM and noise from certain directions. This advantage would only be real if both arrays are fed in an optimal way, and if you do not use separate receiving antennas.

Personally I question whether the effort is worth the result. Having been a user of a Four Square with just four directions for nearly 20 years now, I have never felt the urge to add four more directions on transmit! If I can hear the DX on one of my 12 Beverages, I can work it, even if it is right in between two adjacent main directions.

In principle we could design a 90° hybrid-based feed system for an eight-direction array where everything is fed in quadrature. The impedances coming from the Four Square and

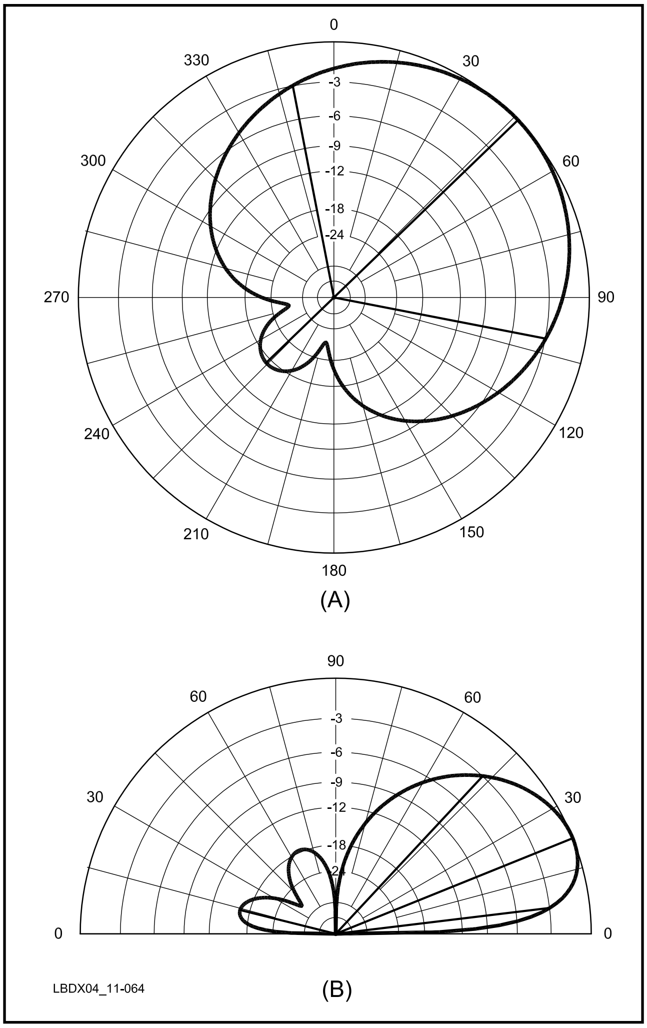


Fig 11-139 — Radiation patterns (horizontal at 20° wave angle) for the optimized side-by-side end-fire cells, showing improved high angle rejection off the back due to the 105° phase shift feed.

from the two side-by side end-fire elements are totally different, so we will have to build separate hybrid couplers, both with their own optimization circuits. This becomes fairly complex. Using a single off-the-shelf hybrid coupler (without optimization) will yield poor directivity results. Another problem is that the input impedance of the hybrid coupler will be very different when used in the full or in-between directions.

If you really want the extra directions, go all the way, and design the array around the WA3FET Four Square configuration and use two separate Lewallen feed systems.

If, for λ/4 spacing between the elements, we increase the phase shift to 105°, we get somewhat higher gain and directivity for these directions. **Fig 11-139** shows radiation patterns for the optimized side-by-side end-fire cells.

4.8.1. Data, Optimized Eight-Direction Four Square

The data below are for the intermediate direction, to be used together with the WA3FET configuration for the main direction. The data for the main directions are in Section 4.7.2.

Side of square: λ/4

Ground: ρ = 5 mS, ε = 13

Radial equivalent ground loss resistance: 2 Ω

Feed currents:

I4 = I3 = 1 ∠–105° (front elements)

I1 = I2 = 1 ∠0° (back elements)

Gain: 5.01 dBi (WA3FET optimized regular Four Square configuration: 6.29 dBi)

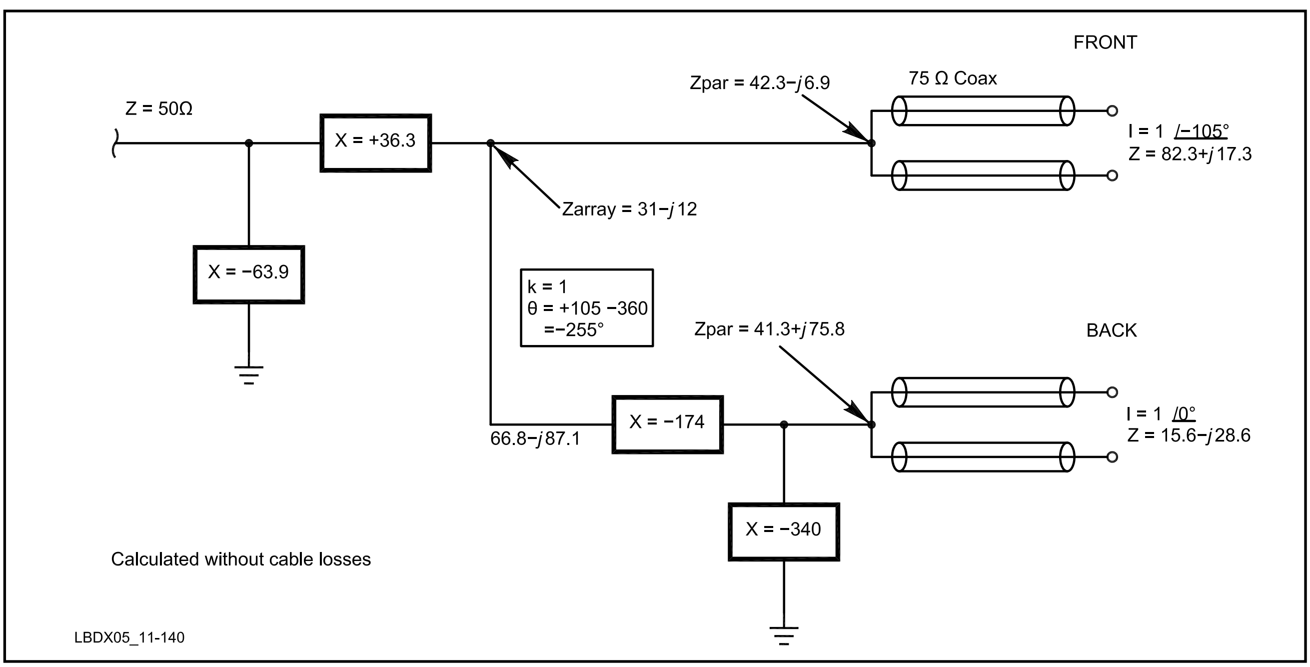


Fig 11-140 — Lahlum/Lewallen feed circuit for the Four Square working as two closely spaced end-fire cells, with optimized phasing, shooting along the directions of the side of the square.

3-dB beamwidth: 113°
RDF = 9.67 dB
DMF = 16.4 dB
Feed-point impedances:
Z4 = Z3 = 82.3 + j 17.3 Ω (front elements)
Z1 = Z2 = 15.62 – j 29.6 Ω (back elements)
Modeling file: *Ch11-4sq-intermediate-dir-optim.ez*

4.8.2. Feed System, Optimized Eight-Direction Four Square

Fig 11-140 shows the Lahlum/Lewallen feed system for the Four Square in its intermediate directions. The circuit includes the L-networks to match the input impedance to the 50-Ω transmission line.

We could also feed the back elements directly and the front element through an L-network to obtain a 105° phase shift but this solution would result in a much lower input impedance.

4.8.3. Direction Switching, Optimized Eight-Direction Four Square

Fig 11-141 shows a possible direction switching method, using a small relay matrix of 12 SPST relays. The truth table shown in Fig 11-141 is translated to a matrix diode switching system as shown in **Fig 11-142**.

4.9. The Broadside/End-Fire Array

In **Fig 11-143** we see how a broadside array with large spacing (0.5 to 0.625 λ) produces good gain and a narrow

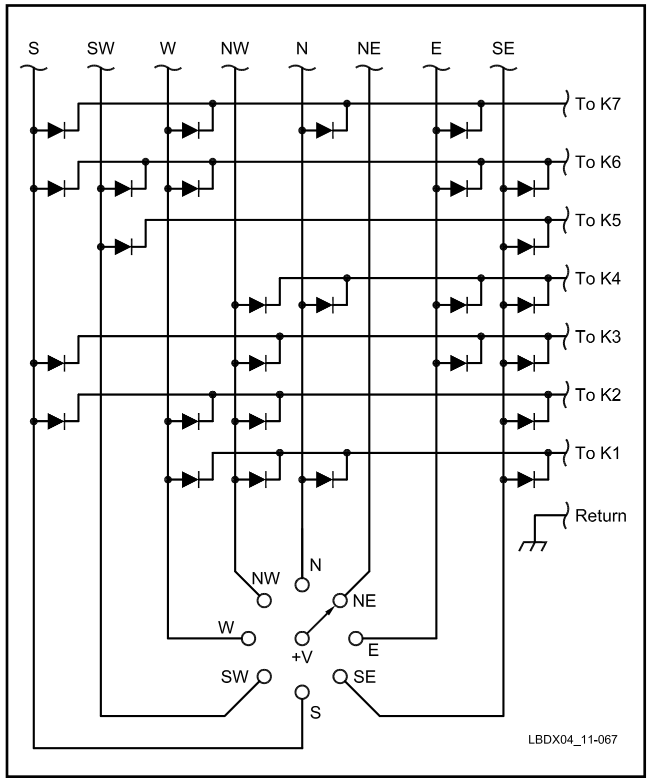
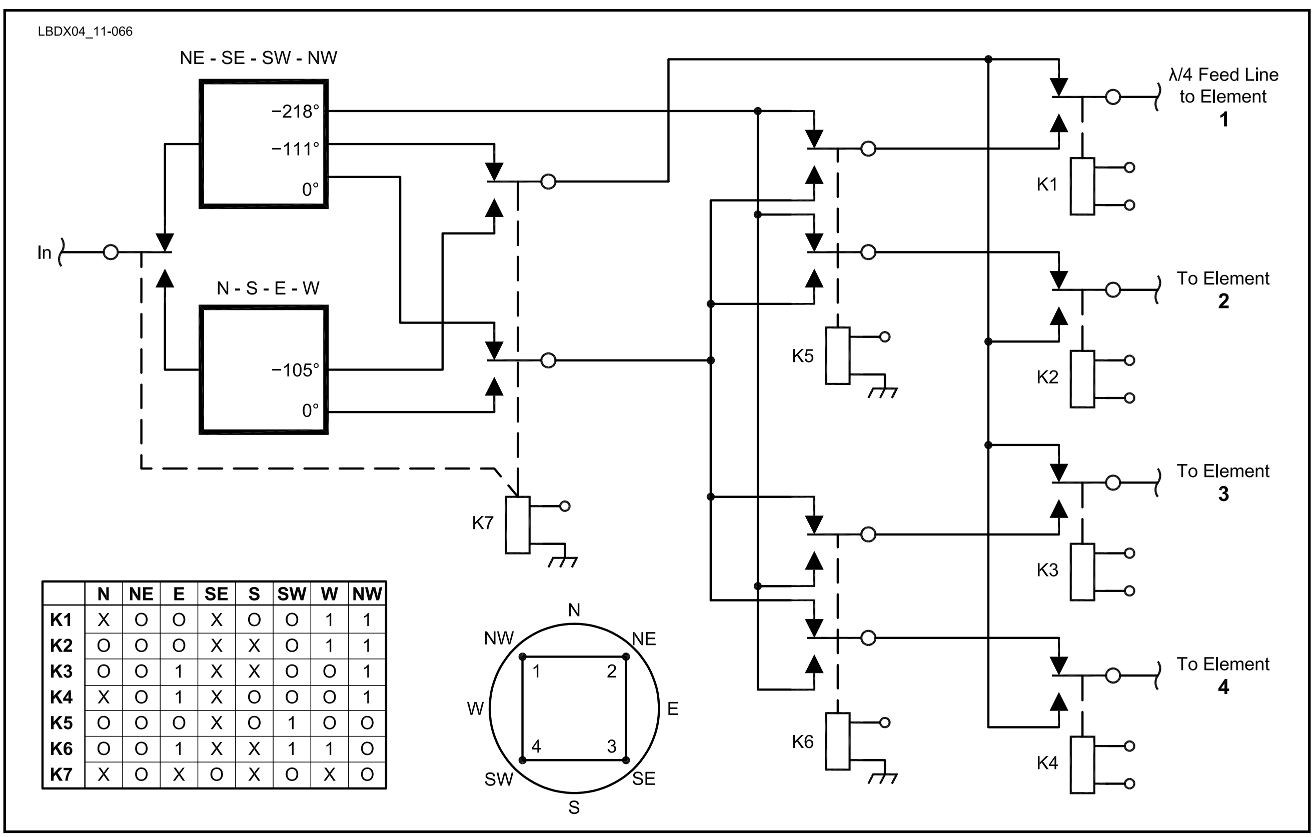


Fig 11-142 — Diode matrix for switching the Four Square in eight directions. This circuit can be applied to 11-141.



	N	NE	E	SE	S	SW	W	NW
K1	X	O	O	X	O	O	1	1
K2	O	O	O	X	X	O	1	1
K3	O	O	1	X	X	O	O	1
K4	X	O	1	X	O	O	O	1
K5	O	O	O	X	O	1	O	O
K6	O	O	1	X	X	1	1	O
K7	X	O	X	O	X	O	X	O

Fig 11-141 — Direction switching for the optimized Four Square (WA3FET) with intermediate direction. Two feed networks are used, and are selected by relay K7. The feed network for the main directions (NE, SE, SW and NW) is shown in Fig 11-132; the network for the intermediate directions is in Fig 11-140.

(bidirectional) lobe. We also know that the 2-element end-fire array can produce a good F/B but has a wide forward lobe. Combining these two principles in one array is using the best of both worlds.

This 160-meter array has a spacing in the end-fire cells of 45 meters and a broadside spacing of 90 meters. The broadside spacing should be at least 80 meters and can be as much as 125 meters, achieving better rejection at high wave angles but poor directivity at low angles.

The end-fire cells could also have much smaller spacing, in which case a different phasing would be required, for example 135° to 145° for λ/4 spacing. The bandwidth would however suffer from the small spacing. The array that was calculated is fed with 90° phase shift.

These end-fire broad-side combinations are also used in the Eight Circle array (Section 4-12).

4.9.1. Data, Broadside/End-Fire Array

Configuration data:
Ground: ρ = 5 mS, ε = 13
Radial equivalent ground loss resistance: 2 Ω
Broadside spacing: 0.55 λ

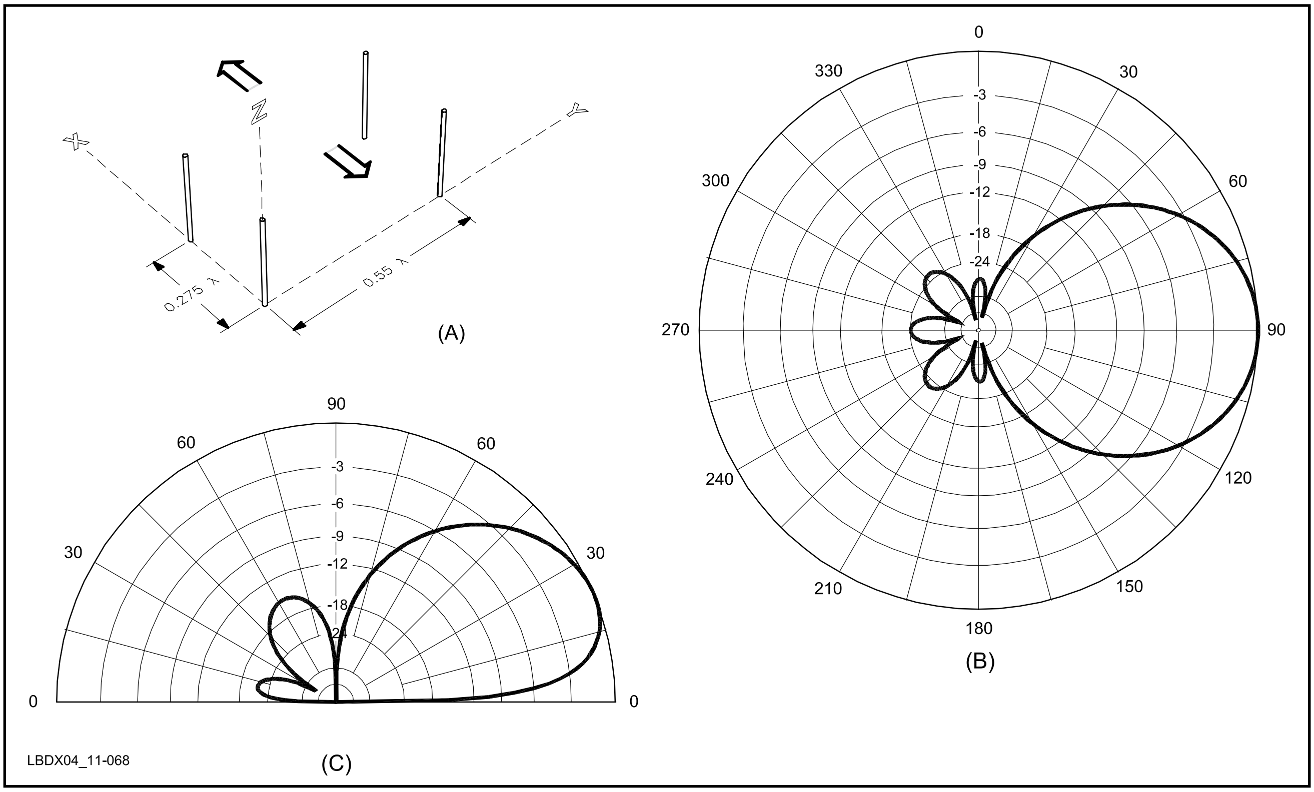


Fig 11-143 — Layout and radiation pattern (horizontal pattern at 20° wave angle) for a broadside end-fire array with a broadside spacing of 0.55 λ and side-by-side spacing of 0.275 λ.

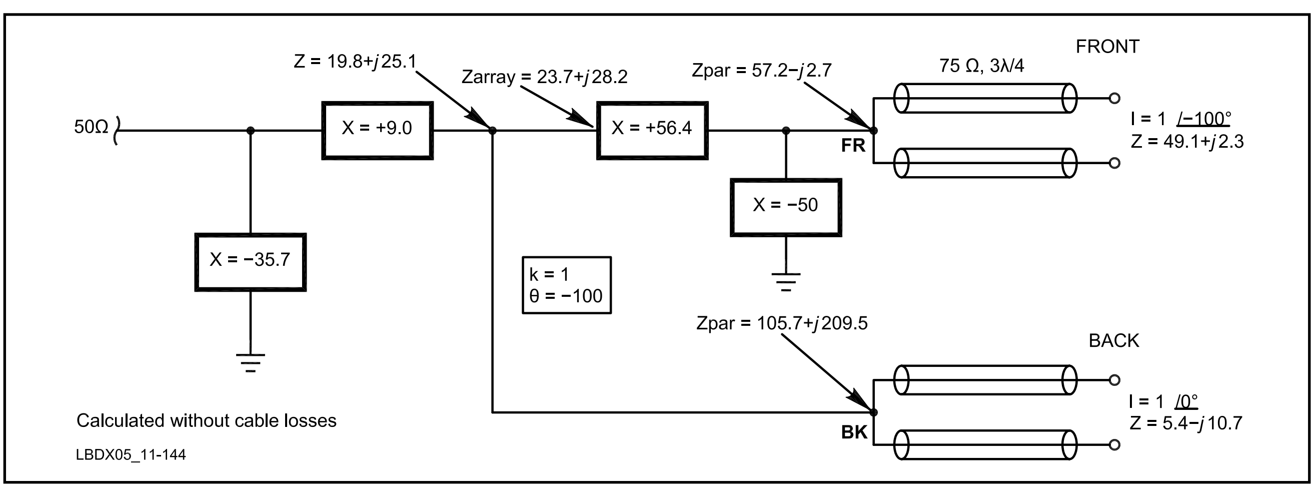


Fig 11-144 — Lewallen feed system (L-network) for the array. An L-network is included to match the feed system input impedance to 50 Ω. The alternative where we feed the front elements directly and the back elements via an L-network result in "extreme" L network components values.

End-fire cell separation: 0.275 λ
(modeling file: *Ch11-4el-broadside-endfire.ez*)
Feed currents:
El 1 and El 2 (back): I = 1 ∠ + 0°
El 3 and El 4 (front): 1 ∠–95°
Gain = 7.71 dBi
3-dB forward angle: 58°
RDF = 12.29 dB
DMF = 20.8 dB
Feed impedances:
Z(el 1) = Z(el 2) = 5.4 – *j* 10.7 Ω (back)
Z(el3) = Z(el 4) = 49.1 + *j* 2.3 Ω (front)

4.9.2. Feed System, Broadside/End-Fire Array

The obvious feed system is the Lewallen LC-network type coupler (Section 4.1), which has the advantage of being able to "tune" the array. The network was designed around 75-Ω current-forcing feed lines which results in higher impedances than when using 50-Ω lines. Note that we need to use 270° long feed lines because of the physical separation of the two end-fire cells. We could have either a –100° phase shift L-network feeding the parallel feed lines to the front element, or a –360 + 100 = –260° phase shift L-network feeding the parallel feed lines to the back elements. In view of the very low feed impedances of the back elements, it is much better to opt for the first solution.

Direction switching is very simple. All you need is a single DPDT relay to invert the paired feed lines to the front and to back (points indicated as FR and BK in **Fig 11-144**).

4.10. Five Square Array

The Five Square is a modified Four Square as shown in **Fig 11-145**. The side of square is 0.3 λ. The array can be made to cover eight directions. Shooting along the X and the Y axis the array has one reflector (element #5), one director (element #4) and three elements (1, 2 and 3) fed in phase be it with slightly different current magnitudes.

4.10.1. Data for the "Diagonal" Operation of the Five-Square Array

Configuration data:
Side of square: 0.3 λ
Ground: ρ = 5 mS, ε = 13
Radial equivalent ground loss resistance: 2 Ω (model: *Ch11-5sq-array.ez*)
Feed currents:
El 1 (back): I = 1 ∠0°
El 2 = El 3 (outside center): 0.8 ∠–125°
El 4 (front): I = 1 ∠–255°
El 5 (center): I = 0.8 ∠–125°
Feed impedances:
Z(el 1) = 5.1 – *j* 5.2 Ω (back)
Z(el 2) = Z(el 3) = 40.1 + *j* 0 Ω (outside center)
Z(el 4) = 23.5 + *j* 60.9 Ω (front)
Z(el 5) = 47.3 – *j* 2.9 Ω (center, middle)
Gain = 6.68 dBi
3-dB forward angle: 78°
RDF = 11.8 dB
DMF = 22.9 dB

In this configuration the main radiation is along the diagonal line going through the center element and connecting two opposite corners of the square. Note that the excellent directivity is mainly derived from its relatively narrow forward lobe, typically 20-25° less than for a quadrature-fed Four Square.

4.10.2 Data for the Intermediate Angles of the Five Square Array

Using the same physical layout shown in Fig 11-145 we can shoot along the X or the Y axis, adding another four directions to the array. In this configuration we have two directors and two reflectors (the elements at the corners of the square). See modeling file *Ch11-5sq-array-half-dir.ez.*
Feed currents:
El 1 = El 2 (back): I = 1 ∠ 0°
El 3 = El 4 (front): I = 1.1 ∠–230°
El 5 (center): I = 3.4 ∠–110°
Feed impedances:
Z(el 1) = Z (el 2) = 6.7 – *j* 39.4 Ω (back)
Z(el 3) = Z(el 4) = 17.4 + *j* 74.5 Ω (front)
Z(el 5) = 24.6. + *j* 4.9 Ω (center)
Gain = 5.14 dBi
3-dB forward angle: 106°
RDF = 10.52 dB
DMF = 18.5 dB

In the main directions (shooting diagonally across the square) the performance is substantially better (1.5 dB more gain and 4 dB better DMF). All of this is the result of a substantially narrower 3-dB forward beamwidth (78° vs 106°). See **Fig 11-146**.

In both configurations, the array should be fed using the Lahlum/Lewallen feed system.

The nice thing about this Five Square is that you don't really need five towers. You can hang the center element from some nylon or Dacron cables strung between the four towers! If the center element is slightly shorter because of the sag of the support cables, just top load it with four cross wires, running in the direction of the support cables.

4.10.3. Feed Systems for the Five-Square Array

Fig 11-147 and **Fig 11-148** show two feed systems developed according to the Lahlum/Lewallen system, using the *Lahlum-Lnetwork.xls* spreadsheet program. If you want to make the array cover eight directions, you will have to install both networks and switch them in and out of the circuit according to the direction used.

Note that to reach the center of the square you will need to use λ/4 current-forcing feed lines using foam dielectric with a VF of 0.85, or just slightly reduce the size of the square. Otherwise you will need to use ¾ λ feed lines.

4.10.4. Switching Directions, Five Square Array

Fig 11-149 shows the direction switching, accomplished with a small relay matrix. According to the selected direction the appropriate drive network is selected with relay K13. **Fig 11-150** shows a small vacuum relay suited for array switching.

Table 11-25 shows the relay truth table. From this table you can design a diode switching matrix as shown in Fig 11-142.

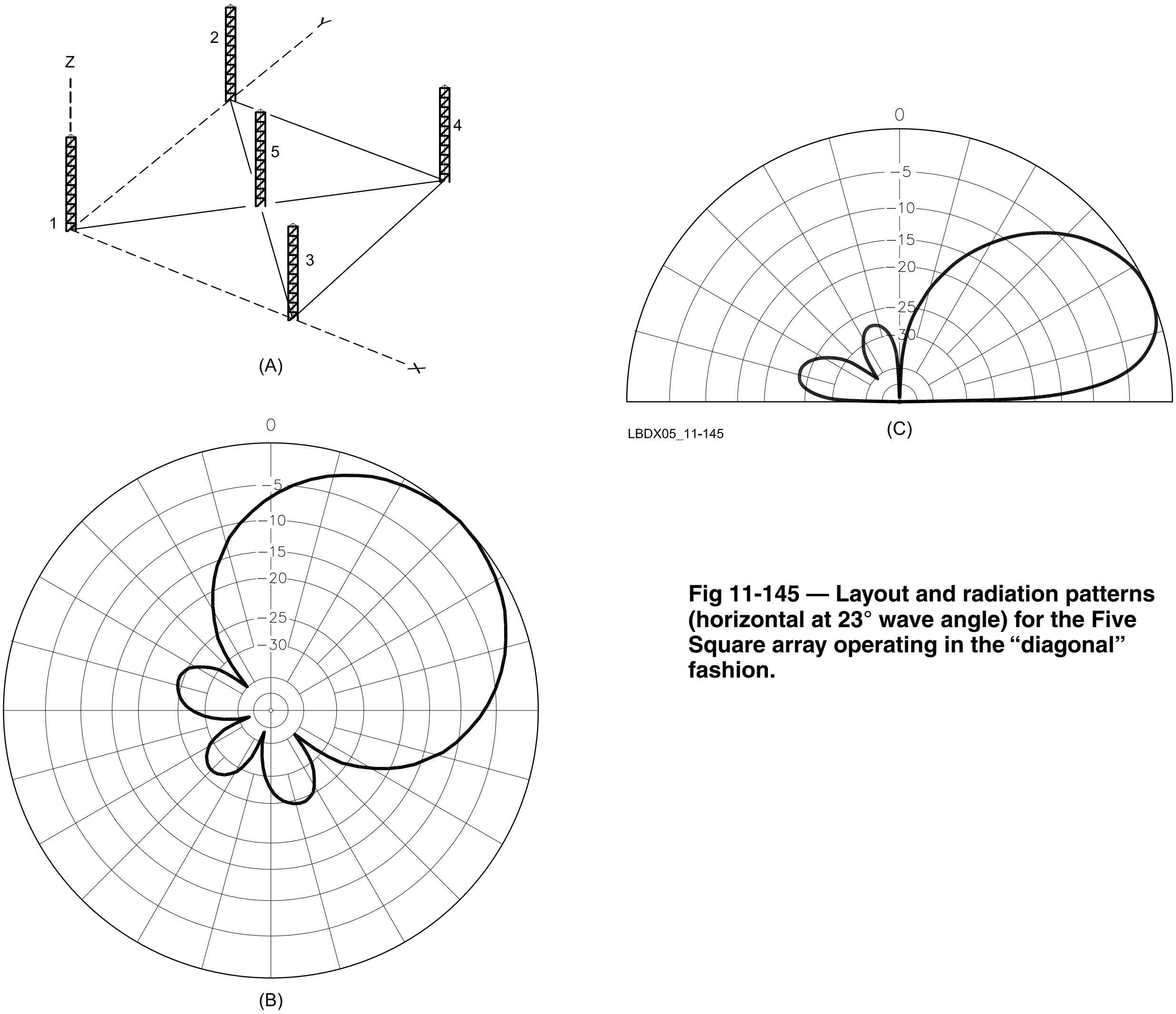


Fig 11-145 — Layout and radiation patterns (horizontal at 23° wave angle) for the Five Square array operating in the "diagonal" fashion.

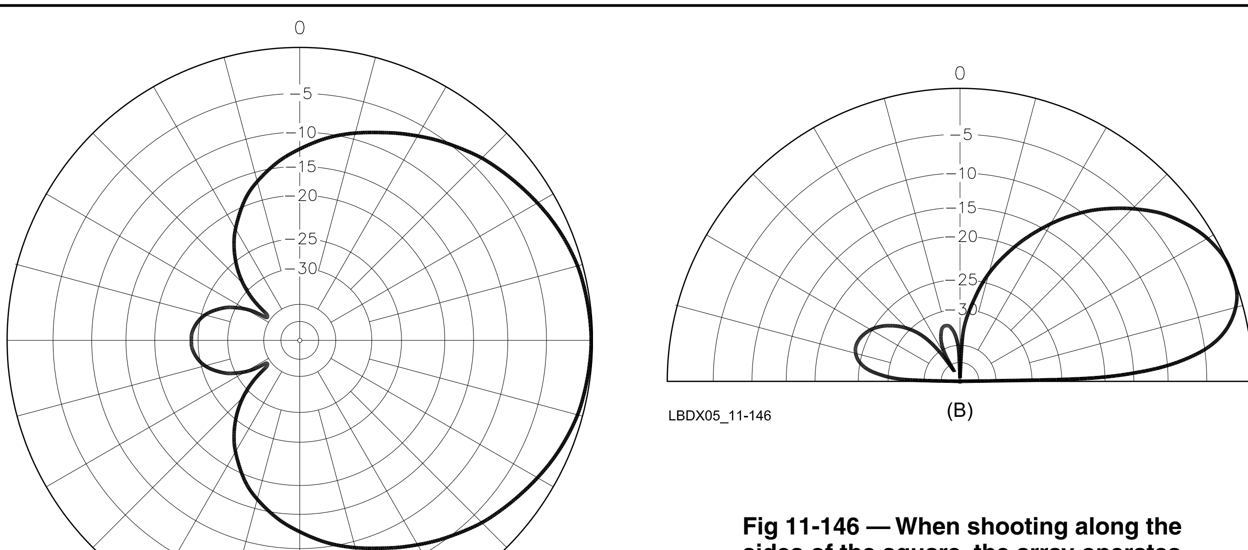


Fig 11-146 — When shooting along the sides of the square, the array operates with one center element with two directors and two reflectors.

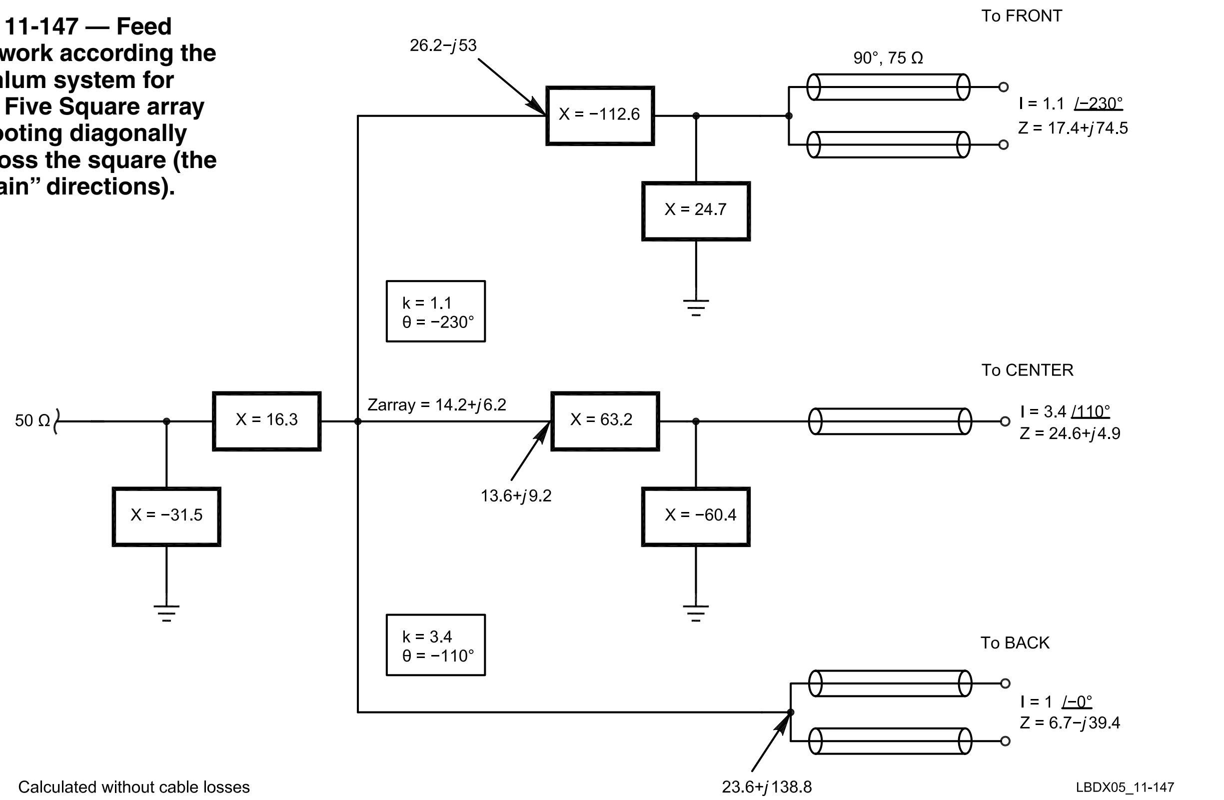


Fig 11-147 — Feed network according the Lahlum system for the Five Square array shooting diagonally across the square (the "main" directions).

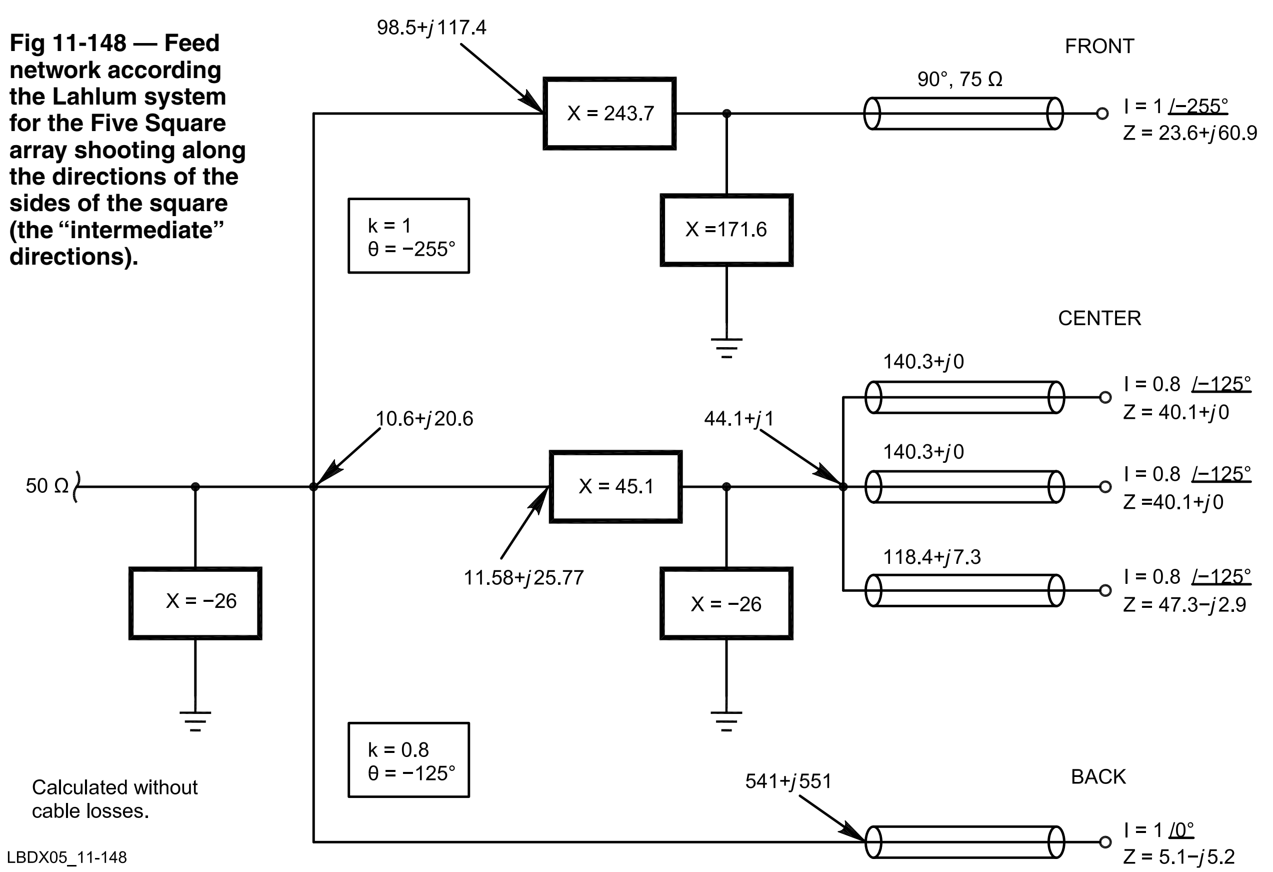


Fig 11-148 — Feed network according the Lahlum system for the Five Square array shooting along the directions of the sides of the square (the "intermediate" directions).

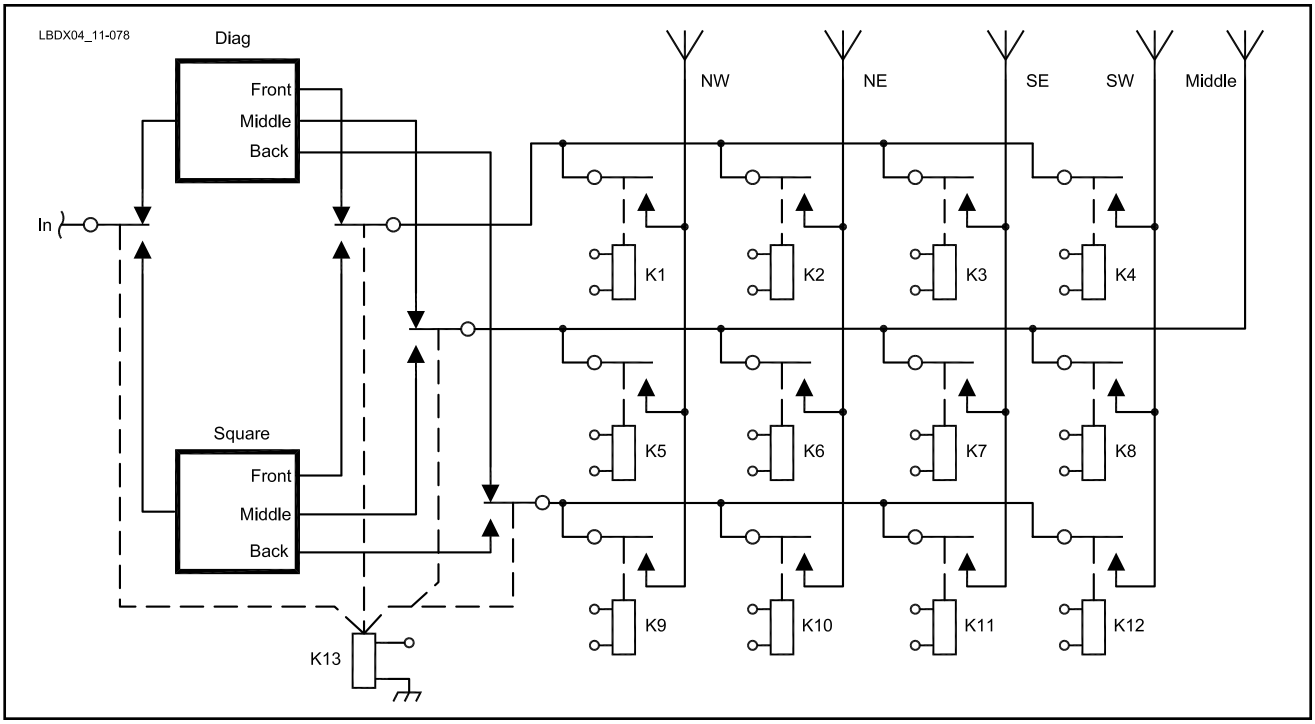


Fig 11-149 — Direction switching system for the Five Square array.

Table 11-25
Truth Table of Relay Matrix for Five Square Array

	K1	*K2*	*K3*	*K4*	*K5*	*K6*	*K7*	*K8*	*K9*	*K10*	*K11*	*K12*	*K13*
N	X	X	O	O	O	O	O	O	O	O	X	X	X
NE	O	X	O	O	X	O	O	X	O	O	O	X	O
E	O	X	X	O	O	O	O	O	X	O	O	X	X
SE	O	O	X	O	O	X	O	X	X	O	O	O	O
S	O	O	X	X	O	O	O	O	X	X	O	O	X
SW	O	O	O	X	X	O	X	O	O	X	O	O	O
W	X	O	O	X	O	O	O	O	O	X	X	O	X
NW	X	O	O	O	O	X	O	X	O	O	X	O	O

Fig 11-150 — Russian-made small vacuum relays with a single make/break contact are ideally suited for a relay matrix.

4.10.5. Conclusion, Five-Square Array

Including 40-meter long radials this 160-meter array requires a terrain measuring 107 by 107 meters (about 3 acres). When using 40-meter long radials, the antenna has a footprint that is only 20% larger than for the classic Four Square with λ/4 sides, yet produces 2 dB more gain and has the possibility of switching in eight directions. Clearly a winner! If you want to use the eight directions, you will of course need two different sets of L-networks to establish the required phased shifts.

With 120 twenty-meter long radials on all elements the footprint is approximately 2 acres, and the trade-off for gain will be marginal (a fraction of a dB).

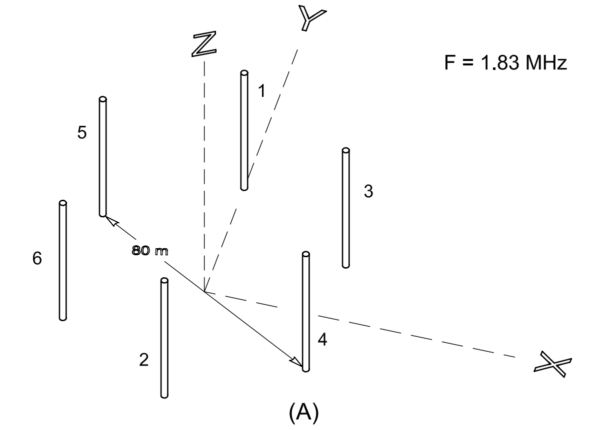
4.11. The Six Circle Array

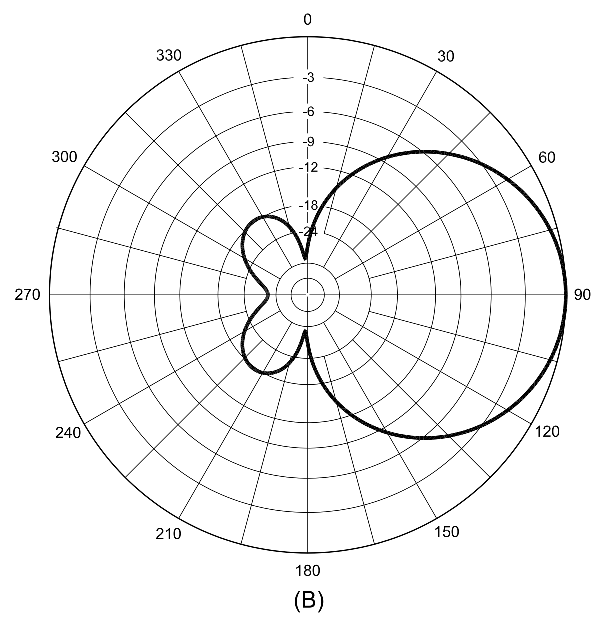
The Six Circle is described in Chapter 7 (Section 1.29) as a receiving antenna. I developed two transmit versions, one having a circle diameter of 80 meters (**Fig 11-151**) and a smaller version 60 meters in diameter (both for a design frequency of 1.83 MHz). Still smaller versions (30 meter diameter) work very well as receiving arrays, but they have low feed impedances and narrow bandwidth when used as a transmit antenna.

4.11.1. Data, Six Circle Array (Diameter = 80 Meters)

Note: The array uses six ¾ wave long 50-Ω feed lines, made of RG-213 (~0.8 dB per 100 meters flat line loss at 1.83 MHz). Six ¾ λ feed lines also represent 720 meters of coax, a considerable quantity from any point of view.

Configuration data:
Array design frequency: 1.83 MHz
Circle diameter: 80 meters
Ground: ρ = 5 mS, = 13
Radial equivalent ground loss resistance: 2 Ω
Model: *Ch11-6circle-HEX-large-quadrature.ez*
Gain = 6.45 dBi (the array loses
1.3 dB in the ¾ wave long feed





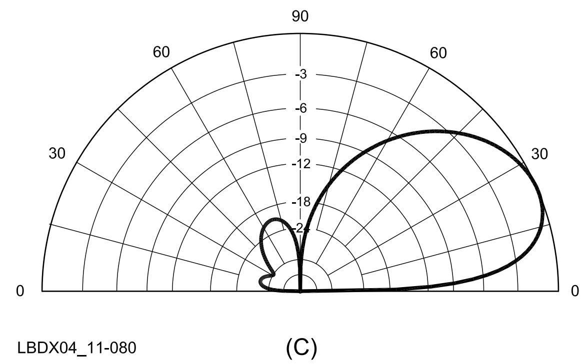


Fig 11-151 — Layout and radiation patterns (horizontal pattern at 20° wave angle) for the Six Circle array. The patterns remain almost identical for different sizes (diameters) of the array. Larger arrays will show higher impedances, and be somewhat easier to feed, and have somewhat more bandwidth. Larger arrays will require 3/4 λ current-forcing feed lines, which has a definite disadvantage when it comes to bandwidth. If you drop the diameter to 60 meters, λ/4 feed lines will reach if you use foam dielectric coax with VF ~ 0.8.

lines, as compared to a calculation
with theoretical lossless lines)
3-dB forward angle: 72.4°
RDF = 11.77 dB
Feed currents:
El 5 and El 6 (back) : 1 ∠0°
El 1 and El 2 (center): I = 2 ∠–90°
El 3 and El 4 (front): I = 1 ∠–180°

4.11.2 Data, Six Circle Array (Diameter = 60 Meters)

The smaller (60 meter diameter) version can work with ¼ λ current-forcing feed lines (providing you use foam dielectric coax with VF ~ 0.82), which helps both on the money budget as well as the dB loss budget. Whereas this smaller HEX array has 6.76 dBi gain (calculated with lossless cables, which is 1 dB down from the 80-meter diameter array), when we include the losses in the feed lines the difference drops to 0.22 dB, which is "negligible. This is why I decided to elaborate further on this smaller (cheaper) array that performs almost equally well.
Feed currents:
El 5 and El 6 (back): 1 ∠0°
El 1 and El 2 (center): 1.75 ∠–90°
El 2 and El 3 (front): 1 ∠–180°
Gain = 6.32 dBi
3-dB forward angle: 87.7°

4.11.3 Feed System for the Quadrature-Fed Six Circle (Diameter = 60 Meters)

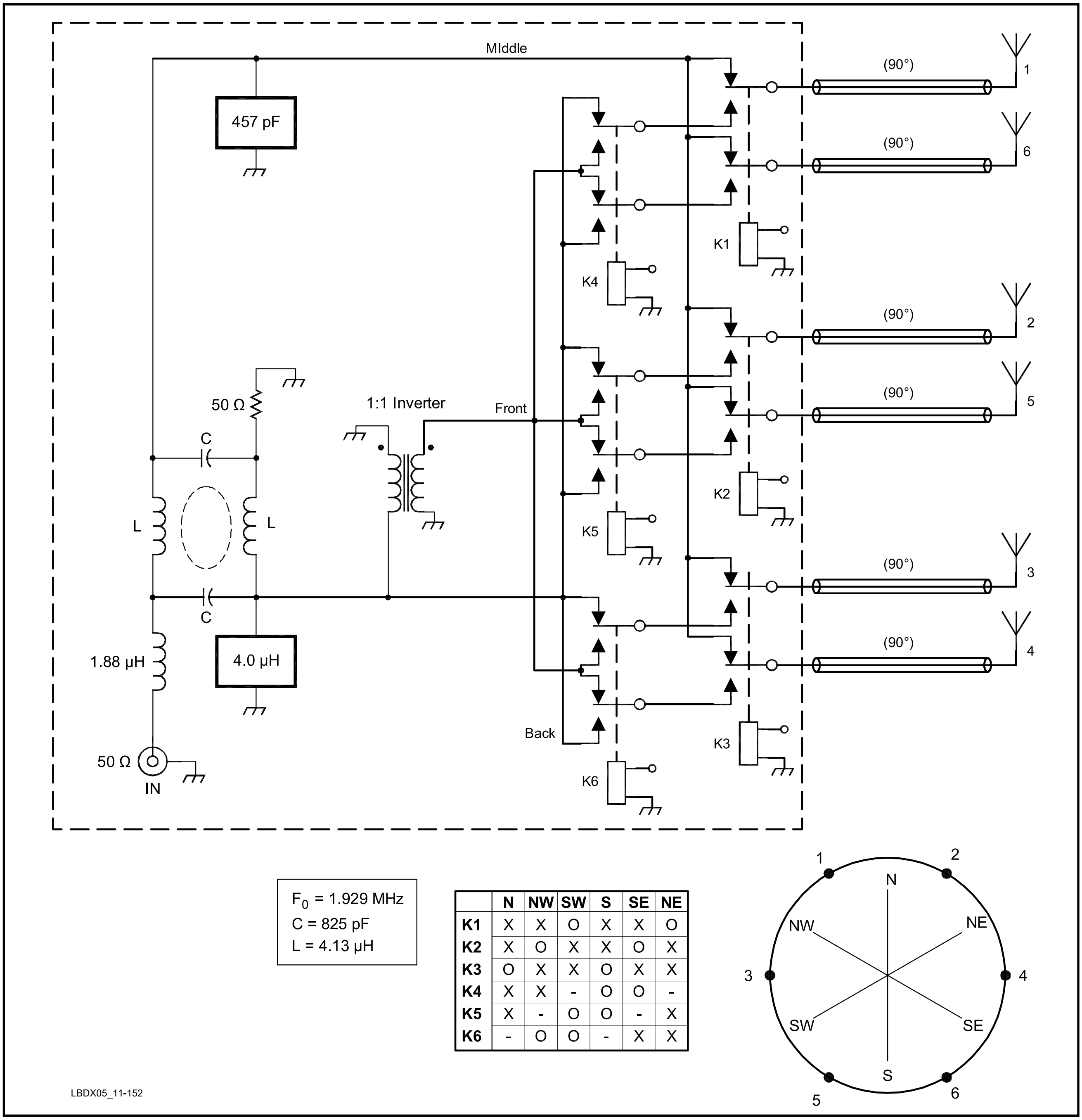
Both the W1MK phase compensation hybrid feed systems and the single-shunt compensated hybrid coupler feed system were described and calculated in Section 3.4.6.10.3.

4.11.4. Direction Switching, Six Circle Array

Fig 11-152 shows the W1MK compensation hybrid coupler feed system (see also Section 3.4.6.10.3) and the direction switching circuitry for the Six Circle array. Six DPDT relays are required.

4.12. The Eight Circle Array

The Eight Circle array (**Figs 11-153** and **11-154**) was, to my knowledge, first built by Tom, W8JI, as a receiving only antenna (see Chapter 7, Section 1.30). In this array we only use four elements at a time. The Eight Circle consists of a wide-spaced broadside array, each cell consisting of a 2-element end-fire array. Using side by side separation of 0.65 λ, one obtains the narrowest possible forward lobe be it at the sacrifice of rather important side lobes (only 11 dB down). For



	N	NW	SW	S	SE	NE
K1	X	X	O	X	X	O
K2	X	O	X	X	O	X
K3	O	X	X	O	X	X
K4	X	X	-	O	O	-
K5	X	-	O	O	-	X
K6	-	O	O	-	X	X

Fig 11-152 — Hybrid feed system with W1MK phase compensation plus direction switching for the quadrature-fed Six Circle array.

F = 1.83 MHz, this results in a circle diameter of 115.8 meters and a separation between the two elements of the end-fire cell of 44.3 meters.

For this edition of the book I made the circle diameter somewhat smaller (104.5 meters), which reduced the important side lobes from –11 dB to –17 dB.

This array is quite simple, but a very good performer. The drawback is its size, and the huge amount of coaxial feed line required. Eight 3λ/4 feed lines on 160 meters (assuming VF = 0.82) represent 800 meters of cable! The nice thing is that we can use quadrature feeding, which makes it possible to use the optimized hybrid coupler feed system. That guarantees the widest possible operational bandwidth.

On the negative side: We cannot use small and cheap feed line for the eight ¾λ feed lines as this is a transmit antenna. Using ½-inch Hardline we will lose 0.4 dB of the theoretical gain in the feed lines With RG-213 we would lose 1.3 dB. Using ⅞-inch 50-Ω Hardline would yield a mere 0.2 dB loss.

4.12.1. Data, Quadrature Fed Eight Circle Array

The data are for the end-fire cells using 90° phase shift (modeling file: *Ch11-Eight Circle-90-deg-cell-104mdiam-incl-feedl.ez*), using 100-meter long 50-Ω feed lines to each element (¾ λ).

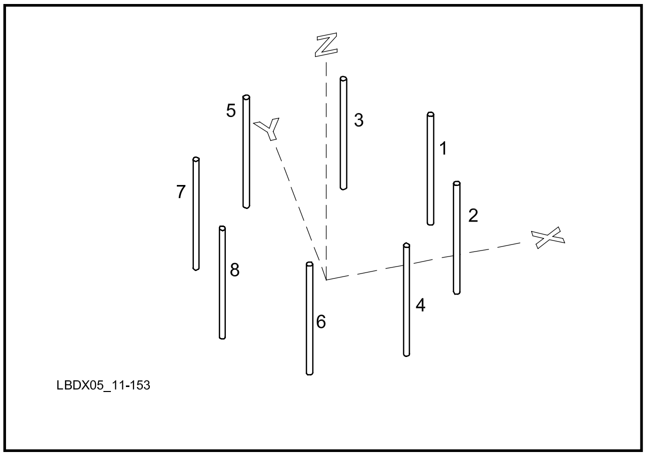


Fig 11-153 — Layout of the Eight Circle array.

Gain = 8.41 dBi
3-dB forward angle: 52.8°
RDF = 12.65 dB
DMF: 21.9 dB
Feed currents:
El 5 and El 6 (back) : 1 ∠0°
El 3 and El 4 (front): I = 1 ∠–90°
Feed impedances, two cells connected in parallel:
Z2 = Z(el 4) in parallel with Z(el 3) = 67.3 + *j* 51 Ω (front)
Z3 = Z(el 5) in parallel with Z(el 6) = 29.1 – *j* 1 Ω (back)

4.12.2. Optimized Hybrid Coupler Feed System

Let's work out the feed system according to the W1MK two-shunt phase compensation system (see Section 3.4.6.6) for f_a = 1.83 MHz.

If we plug the above impedances the *two-shunt-hybrid-comp.xls* spreadsheet as Z3 and Z2, we calculate the required shunt impedance in the 0° leg (port 3) as 847 Ω (73 µH on 1.83 MHz) giving Z3' = 19.13 Ω. In practice we can leave out this coil, as Z2 only has a very small reactive part in the impedance. The shunt element across Z2 is a capacitor of 622 pF, and Z2' becomes 105.95 Ω. Using Z_0 = 50 Ω, the port 4 dump power is down 25.7 dB, which is excellent. The input impedance of the hybrid (24.21 – *j* 17.75) can best be matched to a 50-Ω feed line with a series inductor having a reactance of +17.75 Ω, followed by a 2:1 broadband impedance transformer.

Applying the same Z2 and Z3 impedances to the *single-shunt-hybrid-comp.xls* software, we really find the same solution as in the above case. The compensation was also reduced to single element compensation. The calculation using this single-shunt element method is more correct if we do not use the 73 µH shunt element across Z3 as calculated above. With the single-shunt element software we calculate a shunt capacitor of 678 pF across Z2, a port 4 power loss of 25.1 dB, and a input impedance of 31.6 – *j* 21.9 Ω. If we can live with a 1.2:1 SWR at the design frequency, we can apply the same method as explained above for matching the 50-Ω feed line impedance (a series coil with ZL = +21.9 Ω and a 2:1 broad-band matching transformer).

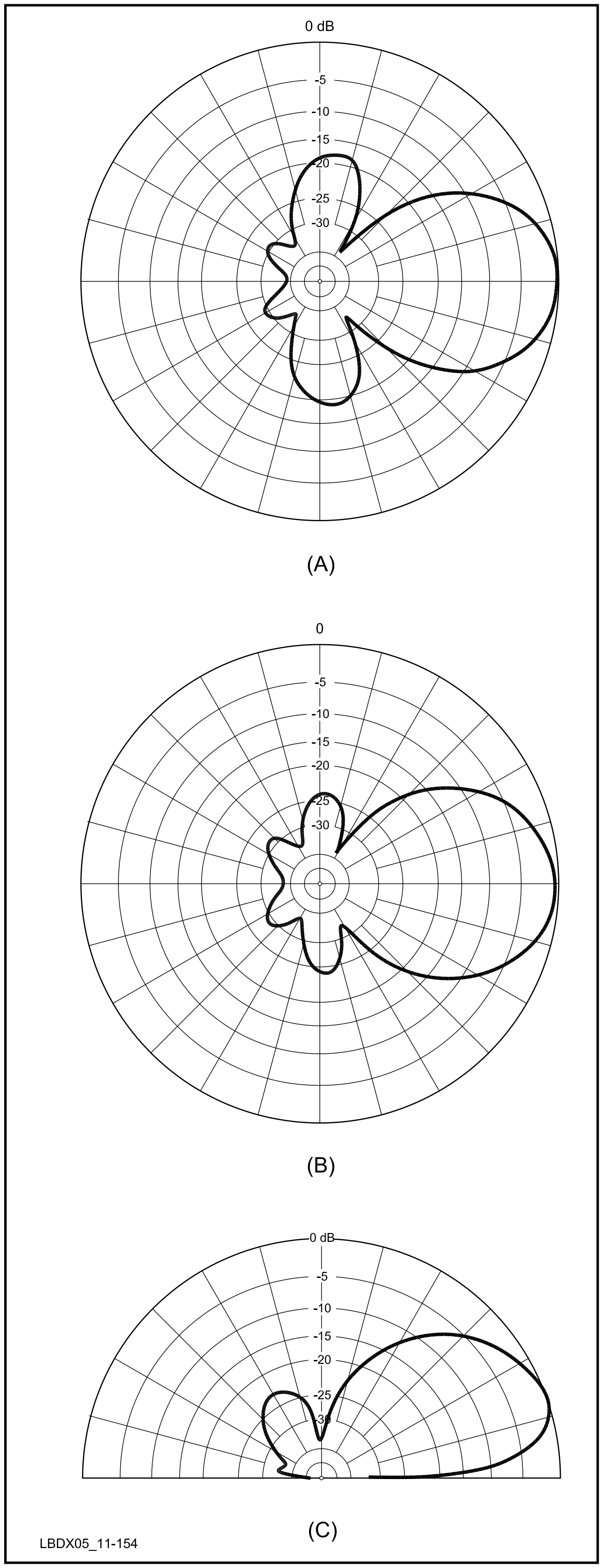


Fig 11-154 — Top: the Eight Circle with a circle diameter of 115.8 meters. Note the important sidelobes. Middle: The same quadrature-fed array with slightly reduced diameter (104.5 meters). In both cases the vertical pattern remains the same as the cells are identical (90° phase shift with k = 1).

4.12.3. Direction Switching System, Eight Circle Array

Fig 11-155 shows the direction switching system for the Eight Circle array.

4.12.4. Discussion, Eight Circle Array

Of all the big high performance arrays, this is certainly the easiest one to build, as there are only two phase angles involved. The directivity is excellent, especially the RDF, because of the narrow forward lobe. With this array you will likely not need separate receiving antennas. Unfortunately you will need about 10 acres of real estate for this array, and almost 1 km of feed line!

4.13. The Nine Circle Array

The Nine Circle transmit array was originally developed by John Brosnahan, WØUN. To my knowledge only two transmit arrays were ever built, one for 160 meters at K9DX (Ref 989), another one for 80 meters at K4JA (unfortunately no longer up). A couple of receive-only versions have been built as well (see Chapter 7, Section 1.33). See **Figs 11-156** and **11-157**.

4.13.1 Data, Nine Circle Array

Configuration data:
Frequency: 1.83 MHz
Circle diameter: 128 meters
Ground: $\rho = 5$ mS, $\varepsilon = 13$
Radial equivalent ground loss resistance: 2 Ω
Model: *Ch11-Nine Circle-big.ez*
Feed currents:

El 6 (back, tip): $I = 1 \angle 0°$
El 5 and El 7 (back, side by side): $I = 1.66 \angle{-90°}$
El 4 and El 6 (middle, outer): $I = 1 \angle{-180°}$
El 1 (middle, center): $I = 3 \angle{-180°}$
El 3 and El 9 (front, side by side): $I = 1.66 \angle{-270°}$
El 2 (front, tip): $I = 1 \angle{-360°}$

Feed impedances:

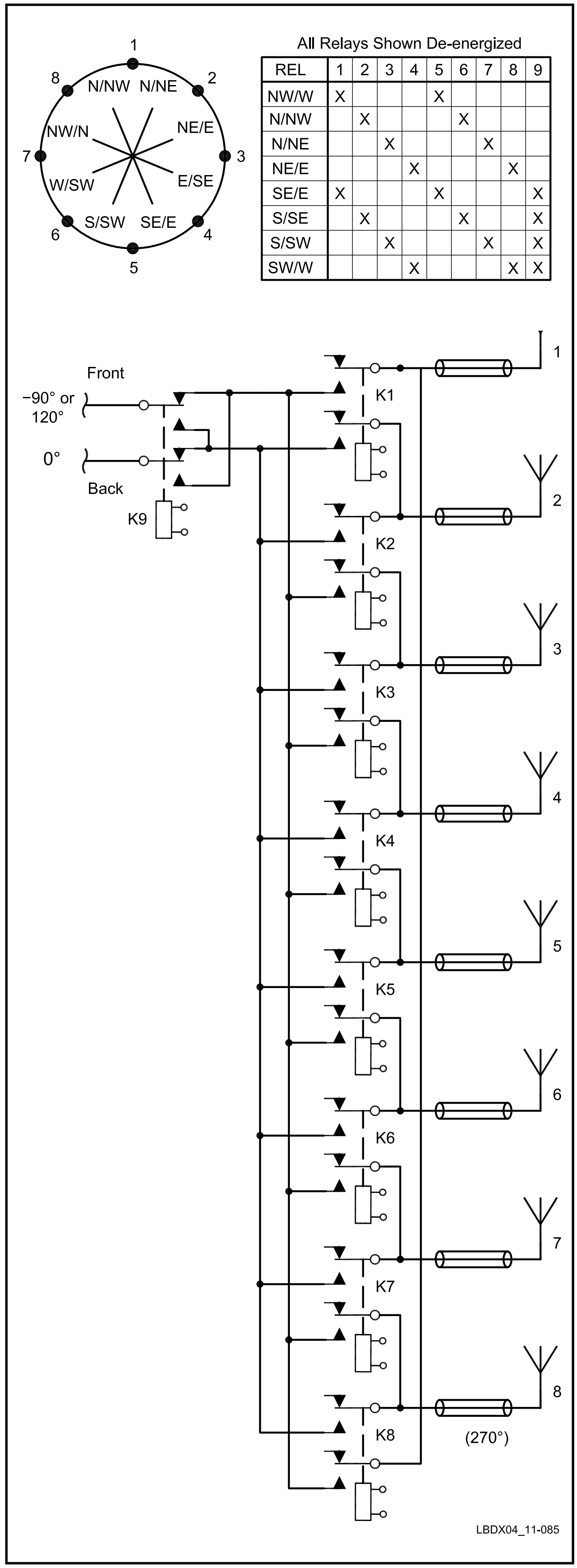
Z(el 6) = $-27.4 + j\,2.2$ Ω (back)
Z(el 5) = Z(el7) = $12.3 - j\,12$ Ω (back, side by side)
Z(el 4) = Z(el 8) = $58.7 - j\,26$ Ω (middle, outer)
Z(el 1) = $36.7 + j\,3.1$ Ω (middle, center)
Z(el 3) = Z(el 9) = $75.9 - j\,2.3$ Ω (front, side by side)
Z(el 2) = $112.5 + j\,157$ Ω (front, tip)

Gain = 9.1 dBi
3-dB forward angle: 58.5°
RDF = 12.99 dB
DMF = 31.7 dB

Note that the impedances include 2 Ω equivalent loss resistance in each element.

I also modeled a smaller version of the Nine Circle, having a diameter of 80 meters vs 128 meters for the "big" one above which is even 10% larger than the "big" Eight Circle described in Section 4.12.

As expected the gain is down somewhat (0.7 dB). The 3-dB beamwidth is a little wider (69.6° vs 58.5°), which results in a somewhat lower RDF (12.39 vs 12.99). The behavior in the back is almost as spectacular as for its larger brother, resulting in a DMF of not less than 31 dB!



REL	1	2	3	4	5	6	7	8	9
NW/W	X				X				
N/NW		X				X			
N/NE			X				X		
NE/E				X				X	
SE/E	X				X				X
S/SE		X				X			X
S/SW			X				X		X
SW/W				X				X	X

Fig 11-155 — Switching system for the Eight Circle array.

Fig 11-156 — Aerial picture of the K9DX Nine Circle array as constructed by John, K9DX. The circular array, including the radials that extend 40 meters from the array itself, has a diameter of 208 meters (two soccer fields long!). The thick stripes are the feed line trenches that had not grown over when the picture was taken.

4.13.2. Data, Small Nine Circle Array

Configuration data:
Frequency: 1.83 MHz
Circle diameter: 80 meters
Ground: ρ = 5 mS, ε = 13
Radial equivalent ground loss resistance: 2 Ω
Model: *Ch11-Nine Circle-small.ez*
Feed currents:
El 6 (back, tip): I = 1 ∠0°
El 5 and El 7 (back, side by side): I = 1.4 ∠–100°
El 4 and El 6 (middle, outer): I = 1 ∠–200°
El 1 (middle, center): I = 3 ∠–200°
El 3 and El 9 (front, side by side): I = 1.4 ∠–300°
El 2 (front, tip) I = 1 ∠–40°
Feed impedances:
Z(el 6) = –18 – *j* 7.5 Ω (back)
Z(el 5) = Z(el 7) = 10.9 – *j* 20.1 Ω (back, side by side)
Z(el 4) = Z(el 8) = 54.9 – *j* 11.1 Ω (middle, outer)
Z(el 1) = 27.6 + *j* 2.2 Ω (middle, center)
Z(el 3) = Z(el 9) = 54.2 + *j* 48.8 Ω (front, side by side)
Z(el 2) = –77 + *j* 104.7 Ω (front, tip)
Gain: 8.15 dBi
3-dB forward angle: 69.6°
RDF = 12.39 dB
DMF = 31.0 dB

Calculations are done including an equivalent ground loss resistance of 2 Ω in each element.

4.13.3. The K9DX Nine Circle Near Chicago

John, K9DX, built his Nine Circle using 27 meter long elements (Titanex 160HD). Using 120 quarter-wave long radials on each element, the tradeoff caused by these shorter elements is nil. John tuned the element with a high-Q coil at the bottom of each element (see **Fig 11-158**). Of course, the feed impedances are different from those shown above.

The impedances (including the loading coil which has an inductance of 120 Ω) are (data obtained from K9DX):
Z(el 6) = –16.6 – *j* 1.0 Ω (back)
Z(el 5) = Z(el7) = 5.9 – *j* 7 Ω (back, side by side)
Z(el 4) = Z(el 8) = 30.3 – *j* 15 Ω (middle, outer)
Z(el 1) = 18.7 + *j* 0 Ω (middle, center)

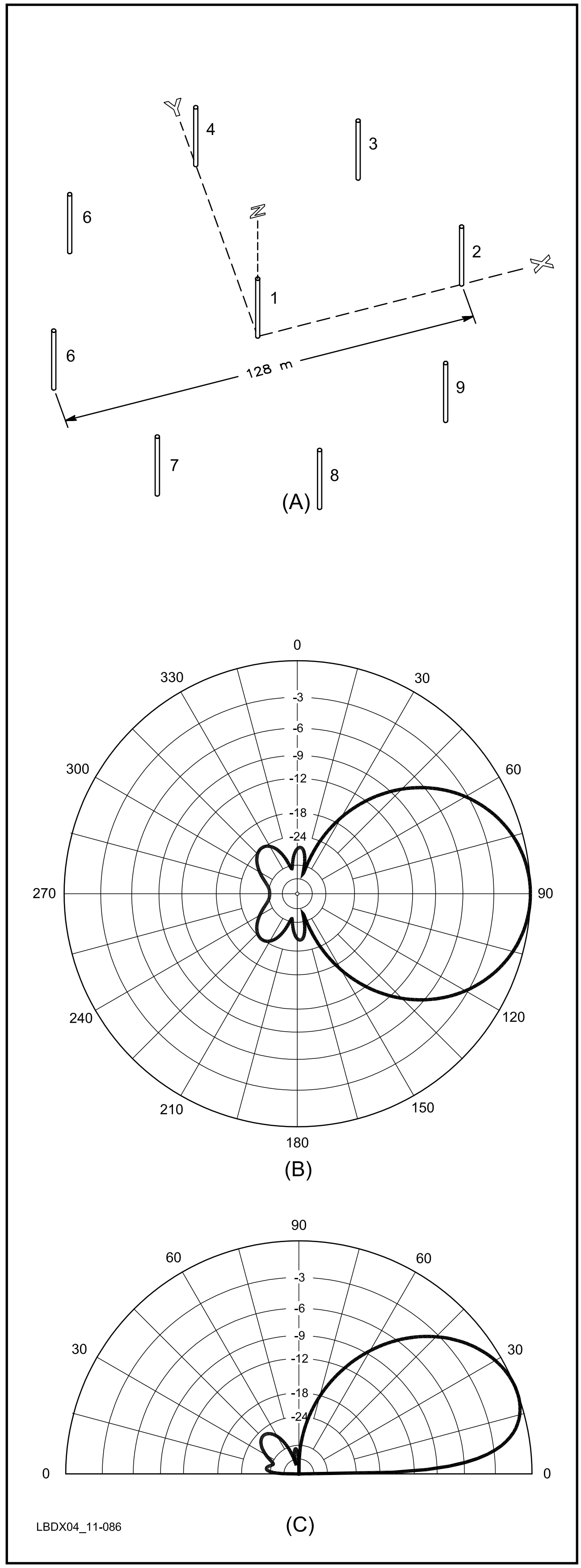


Fig 11-157 — Layout of the Nine Circle, along with the horizontal (at a 20° elevation angle) and vertical radiation patterns obtained with the larger Nine Circle, which measures 128 meters in diameter.

Fig 11-158 — Base of one of the elements of the K9DX Nine Circle. Note the high-Q loading coil and the ring (1 meter diameter) made of 10 mm copper, to which all of the 120 quarter-wave radials are connected.

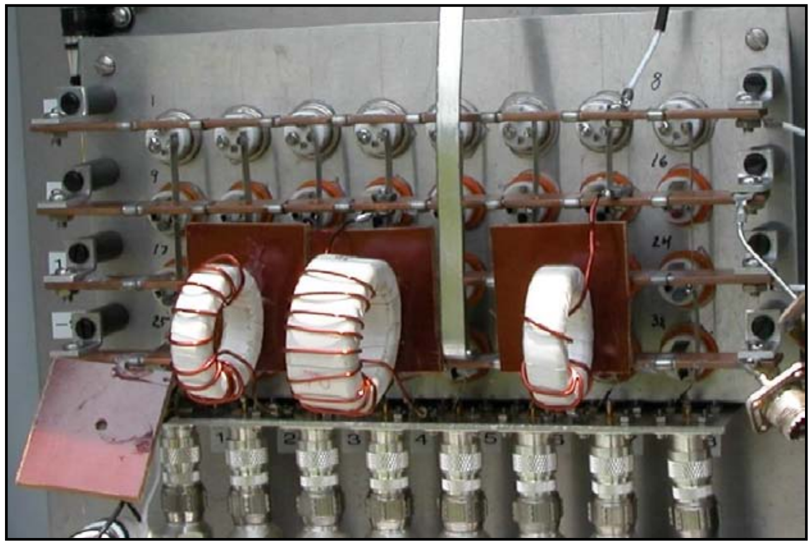


Fig 11-160 — K4JA used a matrix of 8 by 4 small vacuum relays to do direction switching. This has inherently more inductance, but if you use the adjustable L-networks it should be possible to tune out the effect of the stray inductances in the wiring.

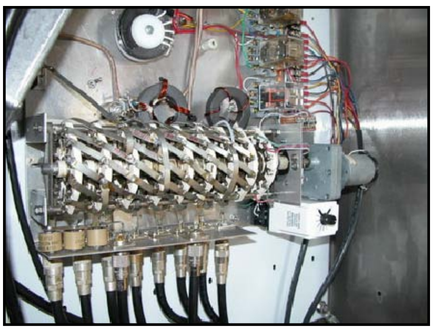


Fig 11-159 — A motor driven rotary switch is used for direction switching at K9DX.

Z(el 3) = Z(el 9) = 52.2 + j 48.8 Ω (front, side by side)
Z(el 2) = 53.7 – j 127 Ω (front, tip)

4.13.4. Feed System, Nine Circle Array

Modeling an impressive array like a Nine Circle is one thing; building it and making it work like the model is a totally different thing! As for designing a feed system, there are many roads that lead to Rome.

John, K9DX, uses a motor driven rotary switch with nine heavy contact ceramic wafers (see **Fig 11-159**). Each wafer is connected to one element. This system ensures minimum inductance, but is expensive if you need to buy the switch new. It can be bought from Multi-Tech Industries (**multi-tech-industries.com**). For his 80-meter array, K4JA used a matrix of 32 small vacuum relays as shown in **Fig 11-160**.

In the design shown in **Fig 11-161** I made use of two L-networks and three transformers, one 9:1 (impedance ratio) transformer, and two 1:1 180° phase-inversion transformers. Note that this is not the feed system John is actually using.

The feed systems shown in Fig 11-161 and **Fig 11-162** use only two L-networks and *no* additional coaxial cables for obtaining the required phasing angles. The center element (with a relative feed current of 3) is fed directly from the input terminals via a 180° phase-reversal transformer (see also Section 3.4.6.3). This is better than a 180° long piece of coax because it much more frequency-independent. If well made, the transformer has little loss and a phase delay of only a few degrees more than 180° (see Section 3.4.6.3).

The two elements that are fed in phase with the center element, but with ⅓ of the current magnitude, are also fed via an identical 180° phase-inversion transformer and through a 9:1 (impedance ratio) transformer (a 3:1 voltage ratio = 3:1 turns ratio). This ensures that these elements get three times less feed current compared to the center element.

The front and the back element are 360° out of phase, which means that they are in phase. As they have a feed current magnitude of 1, they can be connected directly to the output (low Z side) of the 9:1 transformer.

So far we have the three center elements fed at –180°, the front element at –360° and the back element at 0° Very simple so far, and only broadband components (the transformers) are used.

We will now design two L-networks that take care of the proper phase angle and magnitude for feeding the remaining elements two directors and two reflectors, all fed at a relative current magnitude of 1.66. For this we use the *Lahlum.exe* spreadsheet program. The inputs of the L-networks are connected to the "3 V rail" (the line that feeds the center element), which has a phase angle of –180°. The two directors require a phase angle of –270°, so will require a θ of –90°. The k factor is 1.66/3 = 0.55. The two reflectors require a phase angle of –90°, which means that for this L-network θ = –90 – (–180) = + 90° – 360° = –270°. The same k factor applies (0.55).

Fed this way, the array appears to have a total array feed impedance of approximately 19 Ω which is quite acceptable. Using 75-Ω cables (Fig 11-162) for the current-forcing feed lines we obtain a feed impedance of 40.4 + j 19.7 Ω. With just a parallel capacitor (reactance = –102.6 Ω), the feed impedance turns into exactly 50 Ω (lucky strike!). With "real" cables the

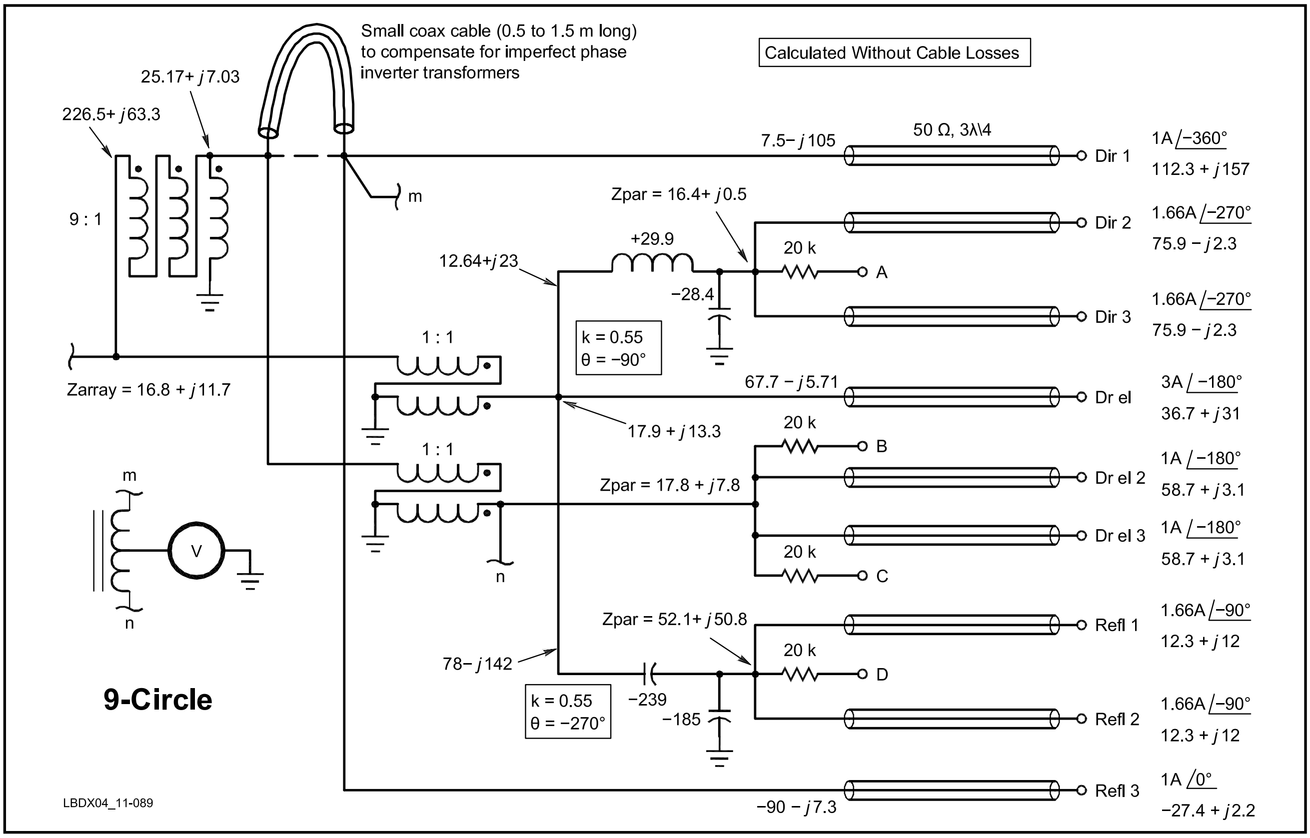


Fig 11-161 — Feed system for the quadrature-fed Nine Circle array based on 50-Ω feed lines. The calculations were done without taking cable losses into account. See text for details.

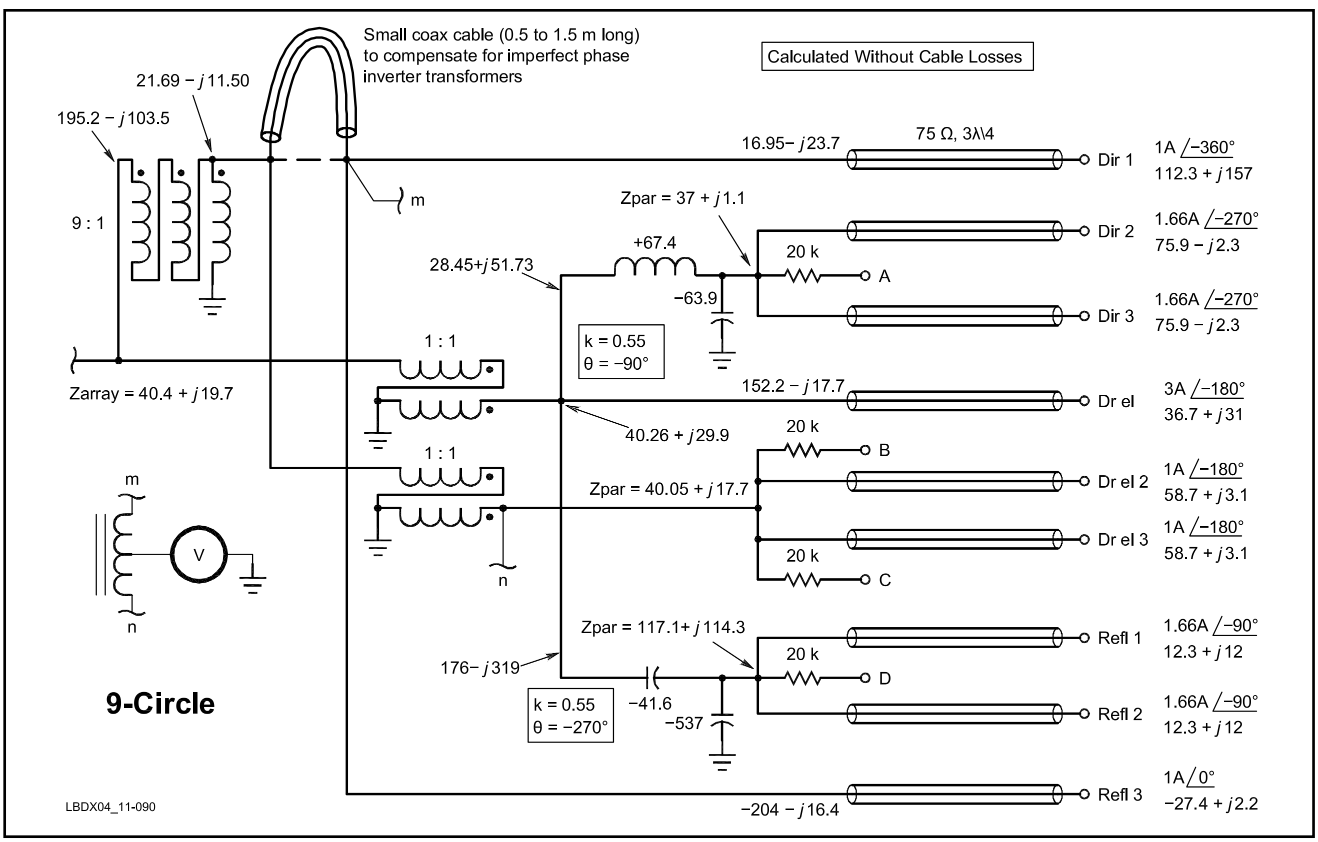


Fig 11-162 — Feed system for the quadrature-fed Nine Circle array based on 75-Ω feed lines. The calculations were done without taking cable losses into account. With real lines (including line losses) the array feed impedance will be very close to 50 Ω.

impedance will be a little higher, maybe 52 to 55 Ω, depending on cable losses.

It turns out that the values calculated for the two L-networks are very normal and quite manageable. To adjust the components to obtain the right phase angle and feed current magnitude, we can use the hybrid-coupler adjustment system as developed by W1MK (see Section 3.6.1).

In Figs 11-161 and 11-162 we see four voltage divider resistors (20 kΩ) installed. All we need to do is connect one hybrid coupler between points A and B and another one between C and D. As the required phase angle difference is 90° we need no extra lengths of coax to correct for non-quadrature phase angles. We will have to provide some attenuation in the legs going to the points B and C though. The voltage ratio is 3/1.66 = 1.81, or in dB: 20 log 1.81 = 5.14 dB. We need two 50-Ω attenuators of 5.14 dB. A T-attenuator using 15-Ω resistors in the series branches and 82 Ω in the parallel branch will be very close.

If we use the detector/wattmeter designed by W1MK, we can adjust the values of the four components until we get a "good" null on the summed output. Bingo!

4.13.5. Broadband Transformer Construction

John, K9DX, has gained a lot of first hand experience in building RF transformers as required in this array. John winds most of his transformers with RG-303 single shield Teflon coax. For all of his transformers John uses ferrite core made of 61 material, permeability of 125, 2.4 inch OD, A_L = 171 (Amidon FT-240-61).

4.13.5.1. The 180° Phase Shift Transformers

The design that John used for his 180° phase shift transformer is covered in detail in Section 3.4.6.3. To compensate for the extra phase shift created by the length of coaxial cable wound on the core, he introduced a short piece of coaxial cable creating a similar amount of "extra phase" shift (3 to 5°) in the branches not fed via the 180° transformer(s). This compensates for the imperfect phase reversal transformer (see Figs 11-161 and 11-162).

To adjust the short coax length in order to obtain 180° phase delay in the system, we can use a small push-pull (balanced) transformer. Connect the balanced inputs (m and n, as shown on the insert of Figs 11-159 and 11-162), and adjust the length for minimum voltage between the center tap and ground. You can use the detector-wattmeter described in Section 3.6.1.1 as RF voltmeter.

John also commented on this issue that "... this 'over 180 degree problem' is one of the reasons I will probably stick to 180 degree coaxial lines in my 80-meter system. The lines can be adjusted to hit the delay right on the nose. Of course the downside is that their length must be changed between phone and CW which adds more complexity."

4.13.5.2. The 9:1 Impedance Ratio Transformer

For the 9:1 transformer, John takes the shield from another piece of coax and slides it over the Teflon insulation to get three turns (see Figs **11-163** and **11-164**).

4.13.6. Bandwidth

A major issue in designing a feed network is bandwidth. It is relatively easy to adjust the phase and magnitude of the antenna currents at one frequency, but depending on the feed system used, things can fall apart rapidly when the frequency is changed.

Fig 11-163 — A 9:1 transformer made by John, K9DX. The transformer counts 4 trifilar turns. The trifilar wiring is made by using a small Teflon coax equipped with a second shield. See text for details.

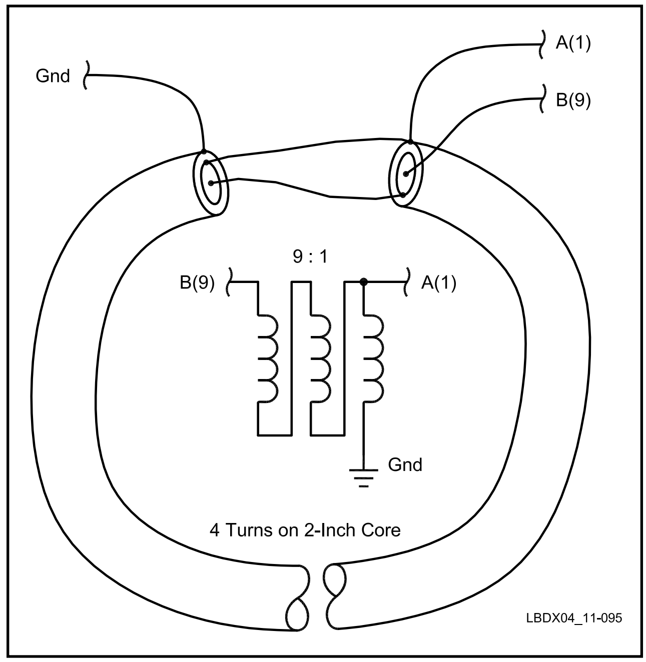


Fig 11-164 — K9DX's construction of a 9:1 transformer.

You can always make a feed system that does exactly what you engineered, on the frequency you engineered it for! If you set out to build an array like a Nine Circle, there is no room for compromises. Compromising to improved bandwidth is inevitably a losing battle. John, K9DX, says that his 160-meter Nine Circle holds perfect directivity (down 30 to 40 dB off the side and in the back) over approximately 30 kHz (1.8% bandwidth), which is adequate for that band.

On 80 meters, trying to cover the CW end and the phone end (8% relative bandwidth) in one network is just not possible. The antenna elements become the wrong lengths, the phase shifting networks (coaxial or L-networks) move far away from their design center, and the current-forcing feeds destroy the amplitude and phase relationships.

Note that this holds true not only for the Nine Circle but for all phased arrays that we want to operate on both the CW and the phone ends of the 80-meter band.

The only good solution on that band is to resonate the elements on both ends of the band (eg with a small extra loading coil on the CW end, or a series capacitor to tune a CW element to become a phone band element), and to adjust the ¾ λ (or λ/4 in case of smaller arrays) current-forcing lines to their exact length by inserting extra cable lengths when operating at the CW end of the band. See **Fig 11-165**.

With a quarter-wave current-forcing feed line, going from 3.5 to 3.8 MHz with a line cut for 3.8 MHz introduces a phase shift error of no less than 7° if the line is flat. With ¾ λ current-forcing lines, the phase error becomes three times that much (on a flat line). On a feed line with SWR (they all have SWR, sometimes very high such as on the "back" or the "front" elements of the array), the phase angle error for a ¾ λ line cut for 3.5 MHz and operating on 3.8 MHz can be as much as 100 or 150°! This makes it impossible to make the antenna work as intended on both ends of the band.

The only solution to cover both 3.5 and 3.8 MHz, is to lengthen the λ/4 feed line approximately 1.4 meters (exact value depending on cable velocity factor). Note that when using ¾ λ feed lines, you need to add almost 5 meters of cable! You should also use a different set of L-network component values for both band ends. In reality this is quite simple. Two DPST relays are all that is needed to switch in two different L-networks, one for each band-end.

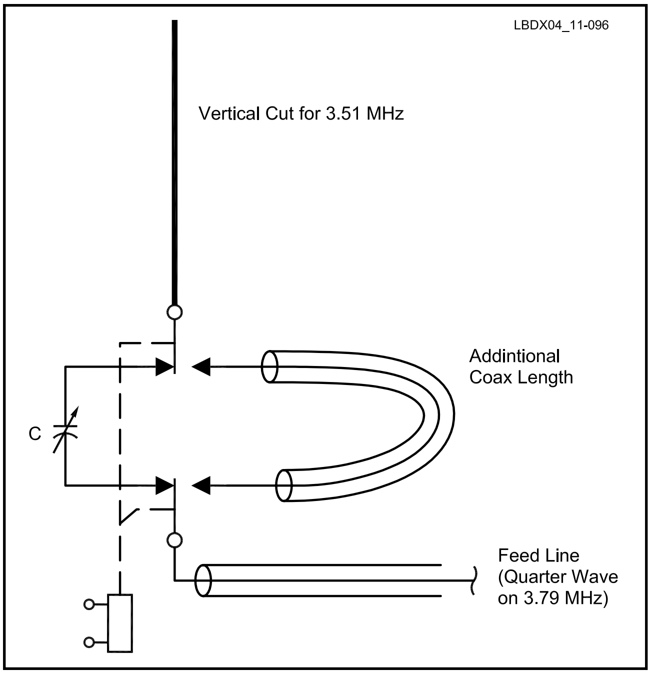


Fig 11-165 — To cover both the CW and phone ends on 80 meters, you must re-resonate the elements and change the length of the current-forcing feed lines.

4.13.7. Conclusion, Nine Circle Array

The Nine Circle is a low-band array most of us can only dream of. It's an interesting subject where you let your imagination go and design your own feed system. There are numerous alternatives, and they all have pros and cons.

Building and owning a Nine Circle array is not for everyone: a 160-meter version requires about 10 acres of real estate! It also requires above-average knowledge of antennas and electronics to design the feed system, to build it and to adjust it.

True, there is no free lunch, and definitely not in antenna matters.

4.14. A Final Overview

Table 11-26 gives a performance overview of the arrays covered in this chapter. Gain is given over "Average Ground,"

Table 11-26
Overview of the Performance Data for Arrays Covered In Section 4

Array Type	*Gain (dBi)*	*RDF (dB)*	*DMF (dB)*	*3-dB Beam-width (°)*	*Foot-print (1)*	*Foot-print (2)*	*Reference Section*
Single vertical	0.34	—	—	—	0.16	0.4	
2 el end-fire quadrature	3.36	8.12	12.3	178	0.32	1.0	4.1.1
2 el end-fire optimized (105°)	3.84	8.7	5	160	0.28	0.9	4.1.1
2 el end-fire 1/8 λ spacing	4.22	9.57	16.6	132	0.24	0.8	4.1.1
3 in line quadrature (λ/4 spacing)	5.31	9.27	17.9	143	0.50	1.3	4.4.1
3 in line optimized (70° spacing)	6.5	11.2	28.7	98	0.40	1.1	4.4.1
Triangle wide quadrature (off top)	4.3	8.8	13.8	147	0.85	1.4	4.6
Triangle optimized (off top)	4.6	9.2	14.4	137	0.80	1.3	4.6
Four Square-quadrature	5.75	10.5	21.0	100	0.64	1.4	4.7.1
Four Square WA3FET	6.29	11.4	24.4	87	0.64	1.4	4.7.2
Four Square λ/8 spacing	4.91	11.4	25	88	0.36	1.0	4.7.3
Four Square half direction optimized	5.01	9.7	16.5	113	0.64	1.4	4.8.2
Broadside-end fire (0.55 λ spacing)	7.71	12.3	20.8	58	1.2	2.1	4.9
Five Square	6.68	11.8	22.9	78	0.8	1.7	4.10.1
Five Square - half dir.	5.14	10.5	18.5	106	0.8	1.7	4.10.2
Six Circle 60-m diameter (1.83 MHz)	7.59	11.7	25.5	78	1.0	2.0	4.11.2
Six Circle 80-m diameter (1.83 MHz)	7.77	11.6	26.3	75	1.5	2.6	4.11.1
Eight Circle optimized (1.83 MHz)	9.20	13.3	21.8	46	2.4	3.8	4.12.2
Nine Circle 128-m diam (1.83 MHz)	9.05	13.0	31.7	70	2.8	4.3	4.13.1
Nine Circle 80-m diam (1.83 MHz)	8.15	12.4	31.0	9.9	1.5	2.5	4.13.2

Gain in dBi over good ground ($\varepsilon = 13$, $\rho = 5$ mS), including 2 Ω ground loss
Footprint (1): 160 m array footprint in ha with ⅛ wave long radials (1 ha ~ 2.5 acres)
Footprint (2): 160 m array footprint in ha with ¼ wave long radials

RDF and DMF are also listed. Interesting information is the footprint required, which is given for the array with 20 meter long radials as well as for 40 meter long radials. The figures apply for a 160 meter antenna. For an 80 meter array, footprints are four times smaller.

5. ELEMENT CONSTRUCTION

5.1. Mechanical Considerations

Self-supporting λ/4 elements are easy to construct on 40 meters. On 80 meters it becomes more of a challenge, but self-supporting elements are feasible even with tubular elements when using the correct materials and element taper. Tubular full size λ/4 elements for 80 meters, shown in **Fig 11-166**, are available commercially from Array Solutions (**www.arraysolutions.com**) as well as from Titanex (**www.titanex.de**). Lattice-type construction is more commonly used, with tapering-diameter aluminum tubing at the top. On Top Band, most quarter-wave vertical radiators will be guyed towers (Rohn 25 is often used). As it is highly recommended to series-feed the elements of an array, the elements must be insulated from ground. That poses extra mechanical constraints on the construction.

I used the "Element Stress Analysis" module of the *Yagi Design* software (see the chapter on software) to develop self-supporting elements for 40 and 80 meters that withstand high wind loads. As the element is vertical, there is no loading of the element by its own weight, which means that the same element in a vertical position will sustain a higher wind load than in a horizontal position. When using the "Element Stress Analysis" module, one can create this condition by entering a near-zero specific weight for the material used.

It will, however, be much easier if you plan to have at least one level where the vertical can be guyed. This will typically lower the material cost for constructing a vertical that will survive high winds by a factor of three or more. Finally, the element construction that is best for your project will be dictated to a large extent by material availability.

Needless to say, guying materials need to be electrically transparent guy wires (Kevlar, Phillystran, nylon, Dacron, etc) or metallic guy wires broken up into small nonresonant lengths by egg-type insulators. Refer to *The ARRL Antenna Book* (Chapter 22), which covers this aspect in great detail.

Fig 11-166 — Self supporting full-size 80 meter λ/4 element commercially available from Array Solutions.

All the array data in this chapter are for ¼-λ full-size elements. Longer elements are to be avoided at all times (elements are no longer fed at a current maximum and the design becomes more complicated). It is not necessary, however, to use full-size elements. Top-loaded elements that are physically ⅔ full-size length can be used without much compromise. Make sure, however, that all elements in the array use the same amount of top-loading. If guyed elements (aluminum tubing) are used, the top set of guy wires can be used to load the element (see Chapter 9 on vertical antennas). If the array must cover 3.8 MHz as well as 3.5 MHz, a small inductance can be inserted at the base of each vertical (make sure the loading coils are identical!) to establish resonance for all elements at 3.5 MHz. Or you can make the vertical to resonant on 3.5 MHz and insert a series capacitor to bring it to resonance on 3.8 MHz (see also Fig 11-165).

The main cause of failure with guyed aluminum tubing elements is buckling (see Chapter 9, Section 6.1.1) . This usually happens when these conditions are met:

- The distance between guying points is too long.
- Thin-wall flimsy aluminum material is used (easy bending).
- There is too much vertical load on the mast (too much guy pulling).
- There is too much wind (bending in between guying points, eventually turning into buckling).

5.2. Shunt Versus Series Feeding

Shunt feeding the elements of an all-fed array is to be avoided in just about all cases. The matching system (gamma match, omega match, slant-wire match, etc) introduces additional phase shifts that are difficult to model and control. Such phase shifts will mess up the correct feed current in the antenna elements.

Only with arrays where all the feed impedances are identical could shunt feeding be applied successfully. The feed impedances of all elements of an array will be identical only when all the elements are fed in phase (or 180° out of phase). Shunt feeding may be considered for such arrays if the vertical elements as well as the matching systems are identical (including the values of any capacitors or inductors used in the matching system).

If you feel tempted to use your tower loaded with HF antennas as an element of an array, be aware that you might be trying to achieve the impossible.

- The loaded tower may be electrically quite long, which could very well be a hindrance to achieving the required directivity (see Section 2).
- You will be forced to use shunt feeding, which is just about uncontrollable, especially if all elements are not strictly identical (which will rarely be the case with loaded towers).

Loaded towers are just great for single verticals, but are more than a hassle in arrays.

5.3. Loaded Elements

It is not always possible to use full size λ/4 elements, and provided you install a very low-loss ground system, full-size elements are not really required. On 160 meters, many arrays have been built with inverted-L elements, although T-loaded elements will produce much better directivity!

5.3.1. Inverted-L Elements

The inverted-L vertical is described in Chapter 9, Section 7. For a single vertical, where we really do not expect much directivity at all, the inverted-L is a good antenna. In an array where you really are mostly after gain *and* directivity, the horizontally polarized high angle component radiating from the flat top section of the inverted-L is a problem.

5.3.2. T-Loaded Vertical Elements

If the central tower is not high enough to support full-size quarter-wave verticals from the sloping support wires, these verticals can be top-loaded by a sloping top-wire (see also Chapter 9, Section 8). The top-loading wires can be part of the support system, as described in the next section. The vertical elements are loaded with sloping top-wires in order to show resonance at 3.8 MHz. The sloping support wires have the property of not producing any horizontally polarized signal, provided the lengths on both sides of the vertical are the same.

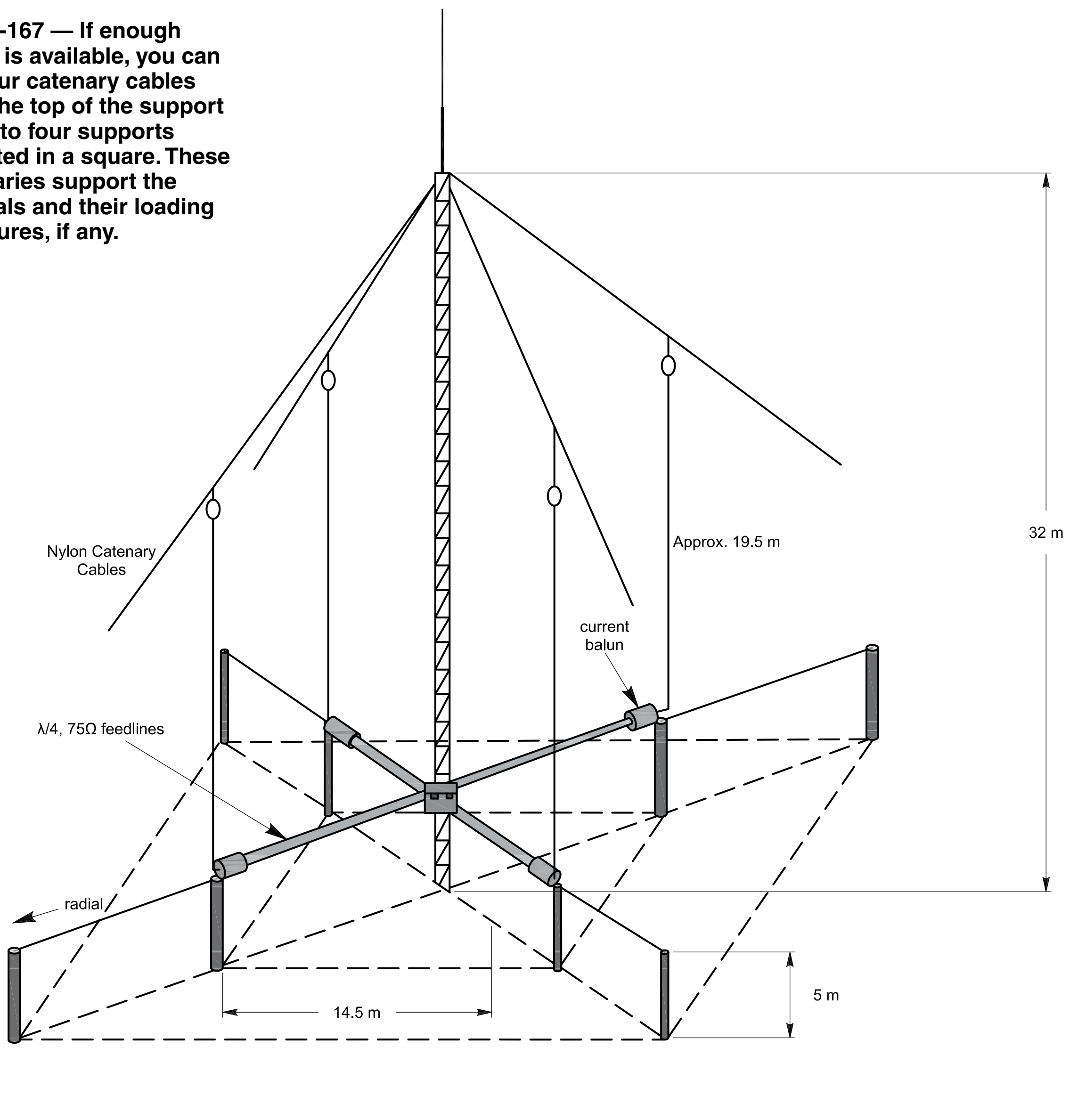
As long as the vertical wire is not shorter than ½ of full size (which means not shorter than λ/8), and provided the current return loss in the ground is small (many radials), the loaded verticals will produce almost the same results as the full-size verticals, with only some reduction in bandwidth. Section 7.5 describes a 160-meter Four Square with vertical elements that are not longer than 18.5 meters.

6. A FOUR SQUARE ARRAY WITH WIRE ELEMENTS

6.1. The Mechanical Concept

An 80-meter Four Square takes a lot of room to put up, not to talk about 160 meters! Almost 20 years ago I installed a somewhat special version of the Four Square around my full-size ¼-λ 160-meter vertical. This design has become very popular since it was first published. From the top of the vertical I run four 8-mm nylon ropes in 90° increments, to distant supports (poles). These nylon ropes serve as support cables from which

Fig 11-167 — If enough space is available, you can run four catenary cables from the top of the support tower to four supports mounted in a square. These catenaries support the verticals and their loading structures, if any.



I suspend the four verticals. A single radial is directed away from the center of the square (where the 160-meter vertical is located). See **Fig 11-167**.

If the central tower is not high enough to support full-size quarter-wave verticals from the sloping support wires, these verticals can be top-loaded by a sloping top-wire. The top-loading wires can be part of the support system. The sloping support wires have the property of not producing any horizontally polarized signal, provided the lengths on both sides of the vertical are the same.

In my particular case with the support being high enough, I managed full-size vertical elements with the feed point and the radial 5 meters above ground. In this setup the single radial serves three purposes:

1) It provides the necessary low-impedance connection for the feed line outer shield.
2) It helps to establish the resonance of the antenna (which is not the case with a large number of radials or buried radials, where the resonance is only determined by the length of the vertical member).
3) It provides some high-angle radiation. We can debate whether or not this is desirable, but in my particular case I wanted a fair amount of high-angle radiation as well, in order to be able to use the array successfully in contests where shorter range contacts are also needed.

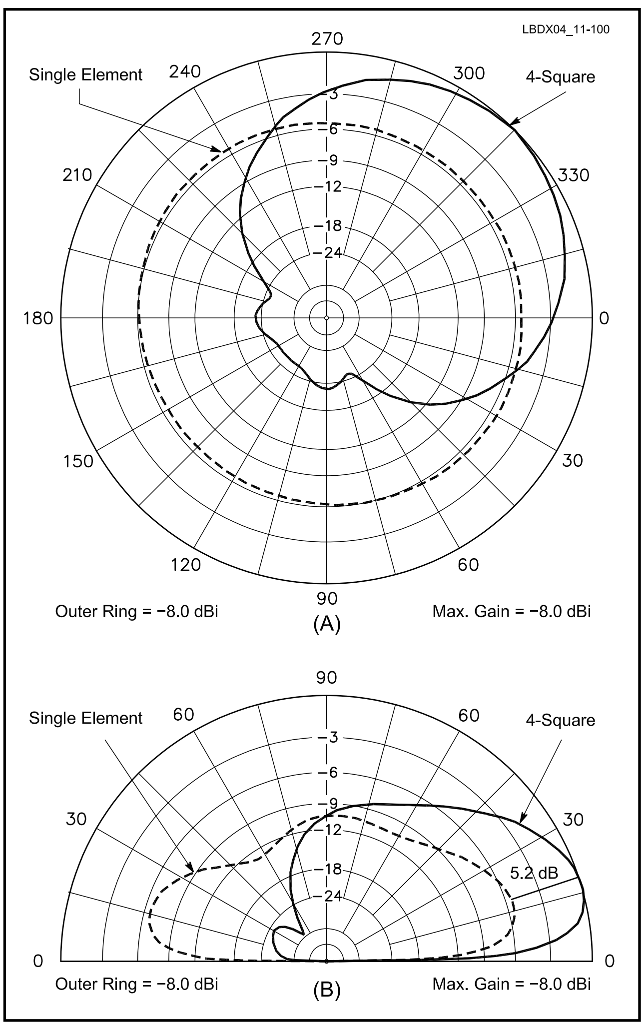


Fig 11-168 — Horizontal and vertical radiation patterns of the author's Four Square array with one elevated radial. Also shown is the pattern of a single vertical element. Both are modeled with a single radial per element, but over an extensive buried-radial system, 5 meters below the radial over very good ground. The buried radials are installed like spokes from the center of the square.

As long as the vertical wire is not shorter than ⅔ full size (approximately 15 meters), the loaded verticals will produce the same results as the full-size verticals, with only some reduction in bandwidth and gain.

Just a single radial, without an extra ground screen (on or in the ground) will make you lose up to 6 dB of maximum achievable forward gain, depending on the quality of the ground below. It will also greatly reduce the directivity of the array. I strongly advocate using a large number of radials on or in the ground, or else a ground screen under the verticals, extending as far as possible (see Chapter 9 , Section 2.2.13). In my particular case, there are some 250 radials (20 to 60 meters long) under the array, basically serving as the radial system for the 160-meter vertical that supports the array. With the extensive radial system, the array exhibits an acceptable degree of directivity (not matching what you can achieve with Beverages, however) and a very worthwhile gain (approximately 5 dB over a single element).

Fig 11-168 shows the modeled horizontal and vertical radiation patterns for the array as well as a single element over identical good ground (very good ground with 250 radials). Both the single vertical and the four elements of the Four Square use a single elevated radial. An *EZNEC* modeling file is available on this book's CD (*Ch11-on4un-wire-4sq-single-rad.ez*).

Array Data:
Design frequency: 3.775 MHz
Good ground: ρ = 30 mS, ε = 20 (very good ground)
Length of verticals: 18.7 meters (2-mm OD wire)
Length of radials: 21.2 meters
Height of feed point/radials: approximately 5 meters
Feed currents:
I1 = 1 $\angle 0°$
I2 = I3 = 1 $\angle -90°$
I4 = 1 $\angle -90°$
Gain: 5.2 dBi (calculated with *NEC-4* using high accuracy real ground method)
3-dB beamwidth: 96°
F/B: 25 dB
RDF: 9.08 dB
DMF: 13.3 dB

The bottom ends of the four vertical wires are supported by steel masts that are located on the corners of a square measuring 20 meters, with the 160-meter vertical (39 meters tall) right in the center of the square. The masts can be folded over for easy access to the element feed point. The vertical elements are 19.5 meters long.

6.2. Loading the Elements for CW Operation

An array making use of wire will exhibit less bandwidth than the same array where the elements are "fat" conductors, such as tower elements. Added to that, the single radial is very much resonant, while a large number of radials exhibit no resonance. All of this gives such a vertical limited operational bandwidth, typically 100 kHz in the 3.5 MHz band. To make the antenna cover the entire 80-meter band I use a stub (or linear loading section), inserted in the radial at the feed point, to shift the resonance of the elements to 3.5 MHz. A high-Q coil would work just as well. A small box is mounted on top of each mast. All connections (to the vertical element, radial, and feed lines) are made inside this box. The box also contains a vacuum relay that can switch the stub in and out of the circuit. The stub is supported by stand-off insulators along the metal support mast (see **Fig 11-169**).

The calculated reactance of the stub is 130 Ω. Using 3-mm (#9 AWG) copper wire with a spacing of 20 cm, the length of the stub turned out to be 2.25 meters long to lower the resonant frequency to 3.505 MHz. The same stub, when shortened to 75 cm, resonates the element at 3.65 MHz (center of the band). A nice feature is that the resonant frequency can be changed anywhere between 3.51 and 3.77 MHz by using a movable shorting bar across the stub. This way, you can create different operating windows on 80 meters. A relay can be used to switch the 3.65-MHz shorting bar in and out of the circuit, making the window selection remotely controlled. The direction control box contains a three-position lever switch, which selects the three band segments.

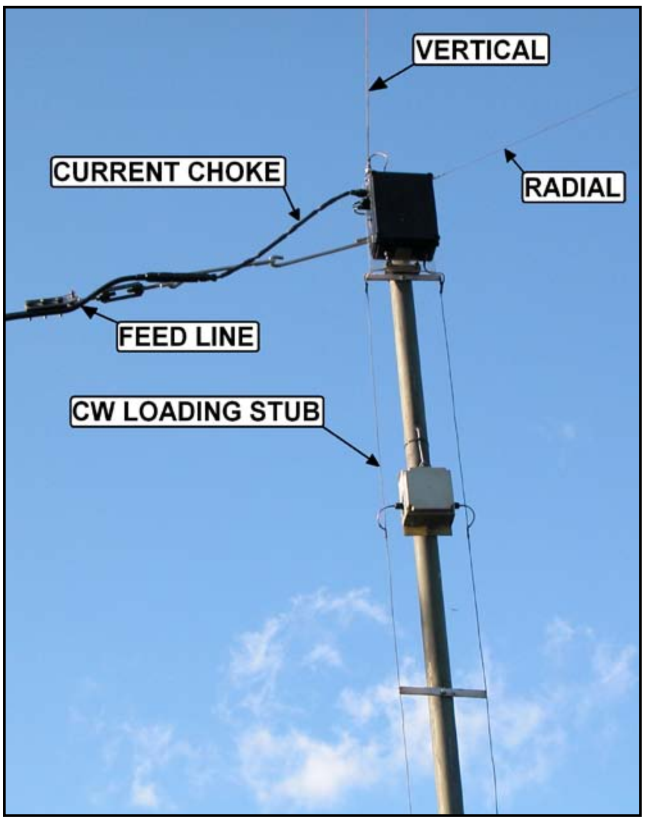


Fig 11-169 — A 15 × 20 × 7 cm plastic box is mounted on the top of the 5-meter support for each of the elevated wire verticals. Inside the box, the vertical wire and the single radial are connected to the feed line, which is equipped with a stack of 100 #73 ferrite beads to remove any common-mode RF from flowing on the outside of the feed line. The stack of 100 cores is outside the box (better cooling!). The box also houses one of the relays to switch the stub to one of the three resonant frequencies (3.51, 3.65 and 3.77 MHz). The stub can be seen running along the steel support mast with another smaller box containing a second shorting relay.

6.3. The 1/4-λ Feed Lines

Each element is fed via an electrical ¼ λ of coaxial feed line with a current balun (100 stacked #73 ferrite beads on a short length of small-diameter Teflon coax) at the feed point. The feed lines were cut to be ¼ λ at 3.75 MHz. If a perfect 90° phase shift is desired at 3.5 MHz, the feed lines can be lengthened by a 1-meter long piece of RG-11A coax (VF = 66%).

6.4. Using the Off-the-Shelf Hybrid Coupler Feed System

6.4.1. Some Measurements

Together with a radial of 18.7 meters, the elements were measured to be resonant on 3.75 MHz where the self impedances measured 40 Ω. The impedance was measured over a frequency range of 2 to 5 MHz using an HP network analyzer with a Smith Chart display. Mutual coupling to other antennas and surrounding structures shows up on the Smith Chart as a kink or a dip in the impedance chart of one or more elements at a specific frequency. It is important that the impedance curves be as nearly alike as possible over the frequency range of interest if the impedance variations when switching antenna directions are to be kept at a minimum. Section 3.5.3.2. deals with the problem of eliminating unwanted mutual coupling.

One word of caution: If the central supporting tower is a base insulated tower, tuned to 160 meters, make sure that the tower is effectively grounded when you use the Four Square. Left floating, this floating central element would act as a half-wave element on 80 meters and interfere heavily with the array. Grounding the central tower can be done in several ways as discussed in Chapter 7 on receiving antennas (Sections 2.11.1 and 3.10). A grounded tower that is resonant at 160 meters and placed in the center of the array does not influence the performance of the Four Square.

6.4.2. Wasted Power

In case of a quadrature fed Four Square with λ/4 side dimensions (over very good ground with lots of buried radials), the use of 50-Ω feed lines results in combined feed-line impedances that load the ports of our 50-Ω hybrid coupler which are quite low and a dumped power that is less than –10 dB. With 75-Ω feed lines, these impedances are 2.25 times higher, resulting in much less power being dissipated in the port 4 load resistor (–15 dB). As already explained before, the main parameter that determines the operational bandwidth of an array fed with a hybrid coupler is the amount of power being dissipated in the load resistor.

Fig 11-170 shows both the dissipated power as well as the array input SWR for the array tuned to the high end of the band (3.7-3.8 MHz), for both the 50 Ω and the 75-Ω feed line impedance case.

6.4.3. Gain and Directivity

Section 3.4.5.1 explains that the feed current in the elements can be assessed by measuring the voltage at the end of the quarter-wave feed lines going to the elements. When I

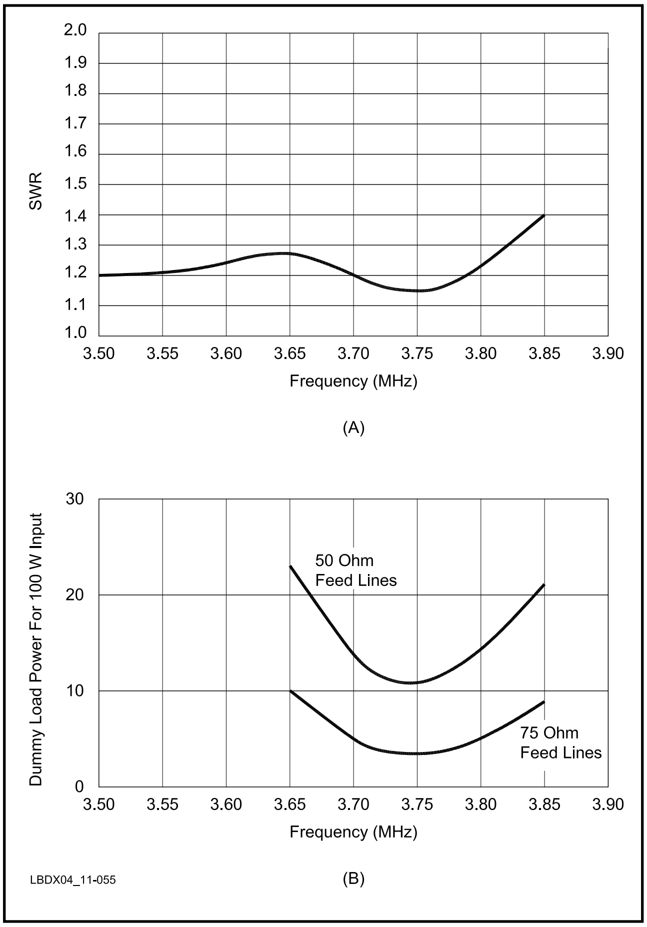


Fig 11-170 — SWR and dissipated-power curves for a Four Square array tuned for operation in the 3.7 to 3.8-MHz portion of the 80 meter band. Note that the dissipated power is much lower with 75-Ω feed line than with the 50-Ω feed line. The SWR curves for both the 50- and the 75-Ω systems are identical. The curve remains very flat anywhere in the band, but it is clear that the power dissipated in the load resistor is what determines a meaningful bandwidth criterion for this antenna.

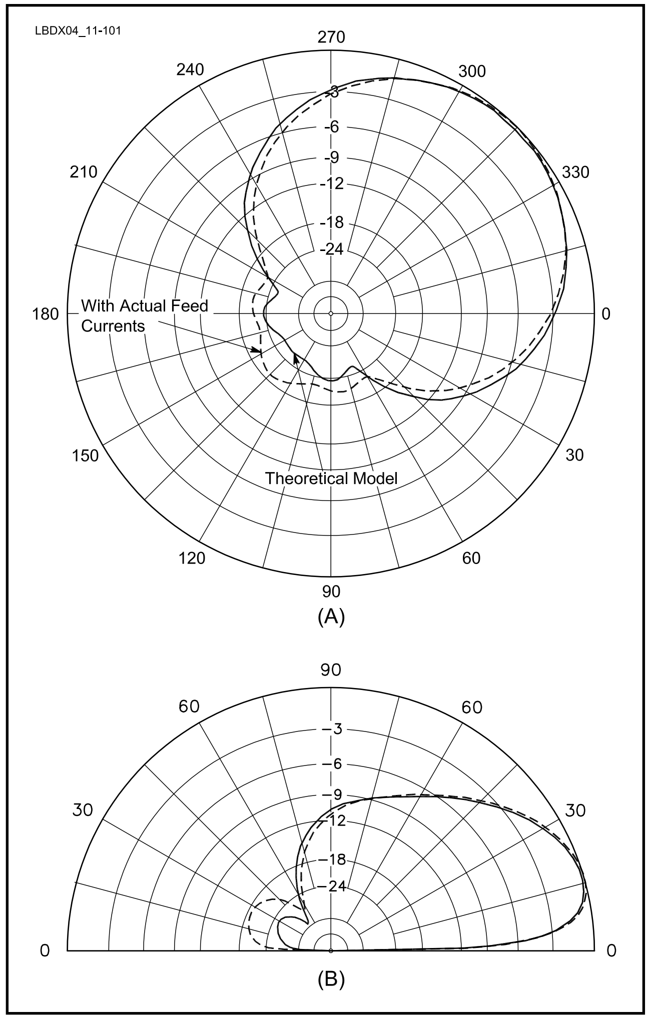


Fig 11-171 — Vertical and horizontal patterns (at a 20° elevation angle) for the theoretical currents and the actual currents at each element of the Four-Square array. Note that although there are some significant current deviations (phase angle and magnitude) from the theoretical values, the array suffers only very slightly from these differences. These patterns were calculated using *EZNEC*.

initially built the array, I used a vector voltmeter to measure the voltages. The results of the measurements using 50-Ω feed lines are listed in **Table 11-27**.

With the current-forcing method employed, the relative element feed-current requirement (in this particular case: equal magnitude, quadrature phase relationship) is reflected in voltages of equal magnitude (where $E = Z_k \times I$ or E = 50 V for a 50-Ω line) at the ends of the ¼-λ feed lines. The table shows the deviation from the theoretical values. From these voltage values the feed currents have been calculated. The resulting gain and F/B performance data as modeled using *EZNEC* are also listed in the table. As expected, the voltage magnitudes and phase angles were not exactly as in the theoretical model (perfect quadrature). The voltage magnitude varied as much as 1.7 dB (41 V versus 50 V), while the phase angle was up to 19° off from the theoretical value for the 50-Ω feed-line case. Table 11-27 also shows the transposed current values at the base of the verticals. In a pleasant surprise, even the relatively important deviations of the 50-Ω impedance case influenced the directivity pattern and gain only very marginally.

Later the 50 Ω ¼-λ feed lines were replaced by 75 Ω lines, which, as expected, resulted in a decrease of wasted power (see **Fig 11-171**). A change from 11% to 4% dumped power represents a relative gain of 0.33 dB, which is respectable.

Using 75 Ω λ/4 (or 3λ/4) feed lines does not make this a 75-Ω system. In this particular case we are still using a hybrid coupler with a 50-Ω nominal design impedance. The 75-Ω cables are used only because they transform the element feed-point impedances to more suitable values, resulting in less power dissipation in the dummy resistor.

Fig 11-171 shows the superimposed vertical radiation patterns of the array with both the theoretical current values as well as the measured values. Note that, for use as a transmit antenna, the actual array comes close to the perfect model,

Table 11-27
Voltages at the Ends of the Quarter-Wavelength Feed Lines of the Four Square Array with One Elevated Radial (50-Ω Feed Lines)

		Element #1	*Element #2*	*Element #3*	*Element #4*
Voltage	Theoretical	50 V, $\angle 0^\circ$	50 V, $\angle -90^\circ$	50 V, $\angle -90^\circ$	50 V, $\angle -180^\circ$
	Measured	41 V, $\angle 0^\circ$	50 V, $\angle -103^\circ$	50 V, $\angle -103^\circ$	44.2 V, $\angle -199^\circ$
Current	Theoretical	1 A, $\angle 0^\circ$	1 A, $\angle -90^\circ$	1 A, $\angle -90^\circ$	1 A, $\angle -180^\circ$
	Calculated from Measurements	1 A, $\angle 0^\circ$	1.22 A, $\angle -103^\circ$	1.22A, $\angle -103^\circ$	1.07 A, $\angle -199^\circ$
Gain	Theoretical	8.13 dBi			
	Calculated from Measurements	8.07 dBi			
F/B	Theoretical	19-25 dBi			
	Calculated from Measurements	17-25 dBi			

although we clearly see a degradation of F/B on the order of 8 dB, and a decrease in RDF and DMF performance.

However, we can make the real antenna perform as well as the model, by optimizing the hybrid coupler as explained in Section 6.6.

6.4.4. Construction

In the original layout, the Comtek Systems 50-Ω hybrid coupler (including a 180° phase shift transformer) and the hybrid-coupler load resistor are located in a cabinet mounted at the base of the 160-meter vertical, which is in the center of the Four Square array. The Comtek unit was removed from its normal housing and the PL-259 hardware was replaced by N connectors.

In order to know at all times how much power is being dissipated in the dummy load, I added a small RF detector to the dummy-load resistor and fed the dc voltage into the shack, where the relative power is displayed on a small moving-coil meter that is mounted on the homemade direction-switching box (**Fig 11-172**). The box also contains the switch to select the subbands. In addition, a level-detector circuit is included, using an LM-339 voltage comparator, which turns on a

Fig 11-172 — Array-direction switch box at ON4UN, with a Four Square direction switch for 40 meters and two switches for 80 meters. A switch selects the Comtek or another feed system (at this time the optimized hybrid feed system is being tested alongside). Relative power and alarm circuit (Fig 11-173) is measuring the power in the dummy load. The lever switch on the left selects one of three band segments on 80 meters.

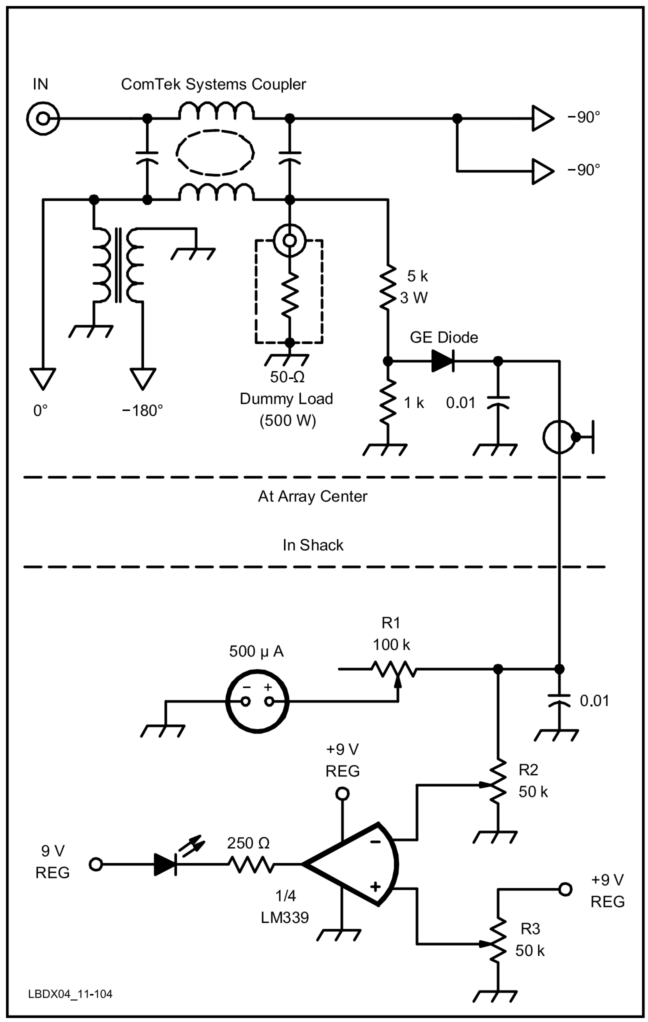


Fig 11-173 — Schematic diagram of the RF detector and voltage comparator used to monitor the RF into the hybrid terminating resistor. The LED will switch on if the voltage coming from the detector is higher than the preset voltage supplied by the potentiometer R3. R1 adjusts the sensitivity of the indicator, and the R2 sets the alarm level.

red LED if the dissipated power goes above a preset value. **Fig 11-173** shows the schematic.

6.4.5. Performance Using the Off-the-Shelf Hybrid Coupler

I have emphasized over and over than you need not look at the input SWR of the hybrid coupler. The SWR says very little about the performance of the system. Assessing the array performance only by measuring its SWR is totally meaningless. With a hybrid coupler, this array shows an SWR of less than 1.5:1 over the entire 80-meter band, wherever the resonance of the elements is.

Based on a sound wasted-power bandwidth criterion (minimum –10 dB), bandwidth of this array is 150 to 200 kHz. Fig 11-170 shows the wasted-power curve for the Four Square array with a single elevated radial, fed with an off-the-shelf Comtek hybrid coupler (see also Section 6.4.1). The steepness of the dissipated-power curve is heavily influenced by the Q factor of the array elements. In the case of this particular Four Square, the elements being made of wire, the Q is high and the bandwidth narrow. Also the fact that I use a single radial instead of a comprehensive (buried) radial system adds to the sharpness of the curve. While changing the frequency away from the design frequency, the single radial (just like the vertical element) will introduce reactance into the feed-point impedance, which would not be the case with a buried radial system.

Practically speaking, this array is by far the best transmit antenna I have ever had on 80 meters. On-the-air tests continuously indicate that the signal strength on DX ranges with the best signals from the continent. As far as directivity is concerned, it is clear that the array has a nice wide forward lobe, and that the relative loss half-way between two adjacent forward lobes is hardly noticeable (typically 2 dB). Long-haul DX very often reports, "You are S9 on the front and I can't copy you off the back." Even on high-angle European signals there is always a good deal of directivity with this array (typically 15 dB).

However, it is *not* a particularly good receiving antenna, as compared to the range of Beverages I have at my QTH. One of the reasons of course is the single elevated radials which cause a big bulge in the vertical radiation pattern at high angles. As explained before, this was done on purpose, so that the antenna would radiate a reasonably strong signal at high angles as well, which is a real asset when working contests.

We have learned that we can greatly improve the directivity by using one of the compensating techniques described in Sections 3.4.6.5, 3.4.6.6 and 3.4.6.7. Those techniques make the 90° hybrid see loads that enable it to deliver voltages at its ports 2 and 3 with θ = 90° and k = 1, on the dot. Such a compensated hybrid coupler feed system is described in Section 6.6.

6.5. Using the Lewallen/Lahlum Phasing-Feed System

An alternative feed system that prevents power from being wasted is the Lewallen/Lahlum L-network feed system, which gives additional flexibility as you can adjust the phase angle and the current magnitude by tuning the elements of the L-networks (see Section 3.4.5).

A drawback of this system is that the SWR and directivity bandwidth of the array are much narrower than with the hybrid coupler system. It is possible, however, to obtain a perfect feed configuration (deviating from k = 1 and θ = 90°) at the design frequency. By its nature, the off-the-shelf hybrid coupler doesn't work well with an array presenting complex loads to the coupler. For covering both the CW and the phone end of 80 meters, I suggest you build two sets of L-networks, one tailored for the CW end and another one for the phone band, and switch them with suitable relays. **Fig 11-174** shows Roger, ON4WU, tuning a Lewallen/Lahlum system.

6.5.1. Using L-Network Feed System in a Non-Quadrature Feed

With this particular array we could only improve the gain of the quadrature-fed configuration by approximately 0.2 dB. The main difference of the optimized configuration is a substantially narrower main forward lobe (84° vs 96° for the quadrature configuration).

A half hour in the company of *EZNEC* resulted in a design, which, using the same geometrical layout of elements, represents only a marginal improvement over the equal-current/quadrature configuration (file: *Ch11-on4un-wire-4sq-single-rad-optim.ez*).

Array data:

Design frequency: 3.775 MHz
Good ground: ρ = 30 mS, ε = 20
Length of verticals: 18.7 meters (2-mm OD wire)
Length of radials: 21.2 meters
Height of feed point/radials: approximately 5 meters
Feed currents:
  I1 = 1 ∠0°
  I2 = I3 = 1.25 ∠–105°
  I4 = 1.5 ∠–220°
Gain: 6.2 dBi (1 dB better than quadrature)
3-dB beamwidth: 88°
F/B: 29 dB
RDF: 9.57 dB
DMF: 14.4 dB

These figures were obtained using real ground with *NEC-4*.

Section 3.4.5.6 covers the detailed calculation of the

Fig 11-174 — Roger, ON4WU, tuning the Lewallen/Lahlum feed system using a five channel scope. The box uses four vacuum variables, two of them in series with a coil, making the combination a variable coil. Note the overlay (generated on the computer on the scope screen, showing the amplitude and phase of the three different signals (see also Fig 11-98).

Lewallen feed system (LC-network) using the *Lahlum-Lnetwork.xls* tool (see also Fig 11-96).

6.5.2. Designing the L-Network Feed System

The Lahlum/Lewallen feed system, as described in Section 3.4.5 enables us to feed the elements with random feed currents and phase angles.

6.6. Using an Optimized Hybrid Coupler Feed System

6.6.1. The Model

In Sections 3.4.6.5 and 3.4.6.6 we described two methods that enable us to let the 90° hybrid coupler do its job under ideal circumstances. Ideal means that the hybrid coupler is loaded on ports 2 and 3 with impedances that ensure a perfect 90° phase shift between ports 2 and 3.

Let us work out such an optimized hybrid coupler system according to W1MK's single-shunt element compensation method described in Section 3.4.6.5. See **Fig 11-175**.

Array design frequency: 3.75 MHz

Element currents and impedances:

I1 = 1 ∠–180°, Z1 = 52.5 + *j* 52 Ω

I2 = I3 = 1 ∠–90°, Z2 = Z3 = 34 + *j* 0 Ω

I4 = 1 ∠0°, Z4 = 7.5 – *j* 2.5 Ω

Note that these impedances are *not* generic, and will differ in each individual case. These impedances should be obtained from a carefully constructed array model, that you have run on *EZNEC* or other modeling software.

At the end of the quarter wave feed lines (75 Ω) we have:

Z1' = 54.1 – *j* 53.6 Ω

Z2' = Z'3 = 165.4 + *j* 2.4 Ω. Two in parallel: 82.7 + *j* 0 Ω

Z4' = 675 + *j* 225 Ω

module 1 SHUNT REACTANCE CALCULATOR for single shunt element hybrid optimization			
	Real	Imag	
ENTER the impedance at port 2 : Z2 (Ω) →	82.70	0.00	Ω
ENTER the impedance at port 3 : Z3 (Ω) →	66.70	-42.50	Ω
ENTER the hybrid design impedance Zo →	75	Ω	
ENTER array design frequency fa (MHz) →	3.750	MHz	
Magnitude voltage at port 2/port 3: k→	1		
OUTPUTS FOR SHUNT ELEMENT ACROSS PORT 3			
Reactance Shunt element	147.18	Ω	
value shunt element =	6.246	µH	
new Z3 =	93.78	0.00	Ω
Fo =	3.973	MHz	
module 2a HYBRID COUPLER DESIGN (by W1MK)			
CASE 1: IMPORTED INPUT DATA FOR SHUNT ACROSS PORT 3			
(imported) fa (antenna frequency) =	3.750	MHz	
(imported) hybrid frequency →	3.973	MHz	
(imported) Zo Design impedance hybrid =	75.00	Ω	
	Real Part	Imag part	
(imported) Impedance load PORT 2 (R2) =	82.70	0.00	Ω
(imported) Impedance load PORT 3 (R3) =	93.78	0.00	Ω
RESULTS			
fo/fa =	1.060		
Hybrid L value (uH) =	3.00	µH	
Hybrid C value (pF) =	267	pF	
Ratio Voltage magnitude Port2/Port3 (k) =	1.000		
Phase angle Voltage port 2 vs. port 3 =	-90.00	°	
Power in Port 4 (vs. Pwr in Port 1) =	-21.9	dB	
Real part input impedance (port 1) =	74.66	Ω	
Imagainare part input impedance (port1) =	3.97	Ω	
Return loss (port 1) =	-31.5	dB	
SWR =	1.05		

Fig 11-175 — Worksheet for the single-shunt element compensated hybrid feed system. Note that we changed Z_0 to 75 Ω to improve the port 4 dump power from –11 to –21 dB.

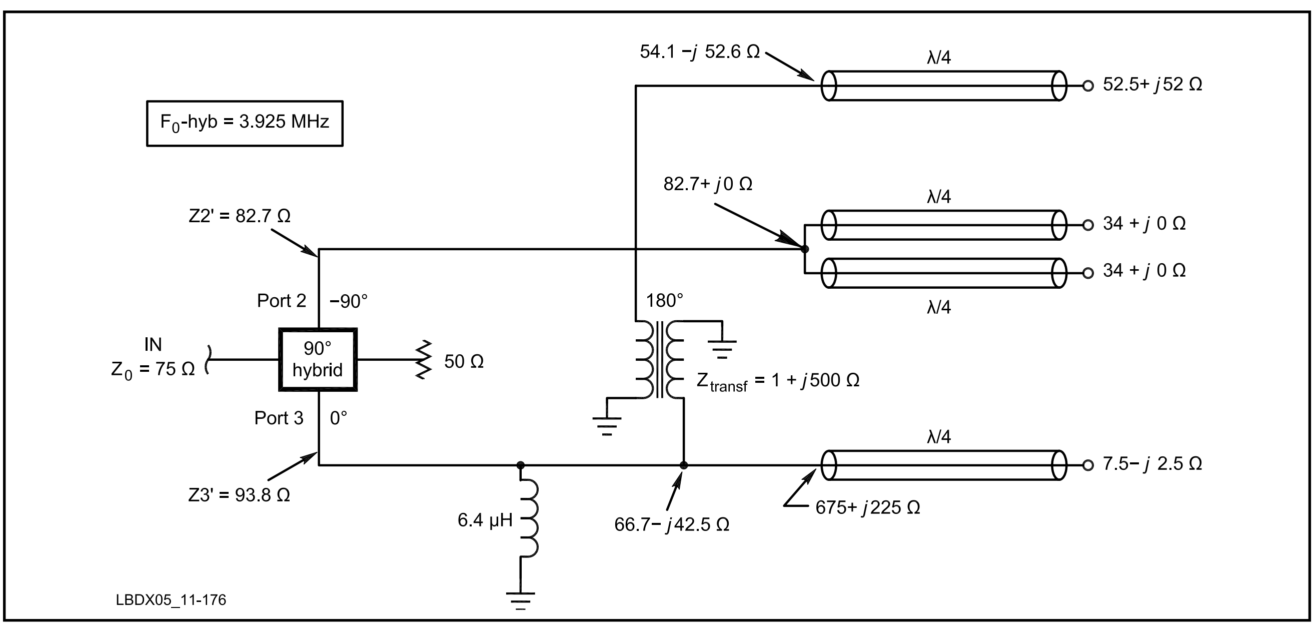


Fig 11-176 — Phase compensated hybrid feed system for the wire Four Square with single radial used at the author's station. The quarter-wave feed lines are made of 75 Ω coax.

Let us assume we use a very good 180° transformer, exhibiting the following load impedance:

$Z_{transformer} = 1 + j\,500\ \Omega$

Z1' plus Z4' + $Z_{transformer}$ in parallel: $Z_{tot} = 66.7 - j\,42.5\ \Omega$

As in this case we have trimmed the element lengths so that the center elements are resonant (in the array configuration, Z2' = Z3' = 82.7 Ω), we will not need any shunt compensation at port 2 of the hybrid. Using the *single-shunt-hybrid-comp.xls* spreadsheet we can calculate the required shunt element (at port 3) to achieve a real impedance load at that port of the hybrid (see Fig 11-143).

Fig 11-176 shows the actual circuit that results. We have adjusted the length of the four elements of the array so that the two center elements show a pure resistive impedance (in this case 34 Ω). Make sure that all four elements exhibit the same self impedance. At the end of the current-forcing feed lines these impedances remain real (after paralleling, 82.7 Ω) which means that no shunt element is required in this leg. Note that the spreadsheet will not accept a 0 value for either the real of the imaginary part of the impedances Z2 and Z3, so we enter 82.7 + *j* 0.01 Ω for Z2.

The program calculated the required shunt element across port 3 as an inductor of 6.4 µH, resulting in a real impedance Z3' = 93.78 Ω.

If we use Z_0 = 50 Ω, the port 4 dump power is down only 11 dB. With Z_0 = 75 Ω we yield –22 dB!

In order to obtain identical voltage magnitudes at port 2 and 3 (looking into different real impedances), the program adjusts the design frequency of the hybrid coupler from 3.650 MHz to 3.925 MHz. You can now enter these data in the *4sq-hyb-w1mk.xls* spreadsheet to verify some of the results.

If we design the hybrid for 75 Ω we can simply use a 75- to 50-Ω L-network or a broadband transformer with a 1:1.2 turns ratio to couple the hybrid to our 50-Ω feed line going to the shack.

We could also have used the "two-shunt elements" or "phase compensation" method, which normally calculates two shunt elements, to make both Z2' and Z3' real. In this particular case, as Z2 is already real (34 Ω), the shunt element at port 2 would not be required, and at port 3 we would find the same value as calculated with the single-shunt reactance compensation method (**Fig 11-177**).

	HYBRID COUPLER OPTIMIZATION TWO SHUNT COMPENSATION DESIGN SYSTEM -w1mk -			
	Enter data in yellow background cells			
	CALCULATING THE SHUNT ELEMENTS			
1	Enter fa (design freq) array →	3.750	MHz	
2	Enter k-value →	1		
		Real part	Imag part	
3	(-90 ° branch) Enter Z2 →	82.67	0.00	Ω
4	Branch 2 shunt element →	-6834328.90	Ω	
5	→	0.01	pF	
6	Z2' (at port 2) →	82.67	Ω	
		Real part	Imag part	
7	(0° branch) Enter Z3 →	66.70	-42.50	Ω
8	Branch 3 shunt element →	147.18	Ω	
9	→	6.25	uH	
10	Z3' (at port 3) →	93.78	Ω	
	HYBRID INPUT DATA			
11	Enter Zo Hybrid →	75.0	Ω	
	OUTPUT DATA			
12	Frequency corr. factor =	1.0597		
13	New Hybrid fo =	3.974	MHz	
14	L-Hybrid =	3.00	uH	
15	C-Hybrd =	267	pF	
16	Dump port (port 4) power ratio =	-21.9	dB	
17	Real part Zin (port 1) =	74.64	Ω	
18	Imag. part Zin (port1) =	3.99	Ω	
19	Port 1 return loss (dB) =	31.5	dB	
20	Port 1 SWR =	1.05		

Fig 11-177 — Calculation sheet using W1MK's two-shunt phase compensation design method (see text for details).

6.6.2. From Virtual to Real World

So far we have played with models and mathematics in a virtual world. There was not yet an antenna you could touch and measure. The model has told us what should be the performance of the planned array, and our spreadsheet program told us that we can make an optimized hybrid feed system making this antenna a top-notch performer over a wide bandwidth (>100 kHz on Top Band).

Now that we have constructed the antenna, time has come to check the symmetry of the array by doing one of the tests described in Section 3.4.6.4.2. Make sure you measure a diagonal isolation of at least 30 dB. You may have to decouple other antenna(s) or tower(s) to achieve enough isolation. This test requires the use of a VNA. Try to get the isolation numbers as high as possible, with the array in all four directions. It you don't have a VNA or don't have access to a VNA (it's nice to have good friends), skip this step, but I would not do it. Maybe it's time to get a VNA yourself?

In the virtual world we calculated Z2 and Z3 with *EZNEC*. Now that we built the antenna, we can actually measure it, be it indirectly. (You can actually measure it directly if the diagonal isolation is infinite, but I would not count on it, unless you measured your diagonal isolation and found 40 dB or so.) We will now measure Z22, Z33, Z2,3, Z3,2 and the impedance of Z2 and Z3 in parallel following the procedure outlined in Sections 3.4.6.4.4 and 3.4.6.4.5. Once you have calculated Z2 and Z3 using the *two-port-coupling.xls* spreadsheet (see Fig 11-45), you can use these in the *two-shunt-hybrid-comp.xls* spreadsheet, which will calculate the shunt elements and f_0.

It is interesting to compare the values of your measurements with the values you obtained in the virtual world from your model. The numbers should be in the same ballpark. If these are way off from one another, it is better to look for the reason. Good luck.

Next you make the shunt element as calculated and install it across port 3. Next the hybrid coupler should be made based on the calculated f_0 (see Section 3.4.6.1).

The ultimate proof of the array is in the measuring of the element drive currents, which can be done using one of the measurement procedures as described in Section 3.6.

7. ARRAYS OF SLOPING VERTICALS

In the chapter on dipoles, I describe the vertical half-wave dipole as well as the sloping half-wave dipole and its evolution into a quarter-wave vertical with one radial. Sloping verticals are well suited for making a Four Square array from using a single, tall tower as a support. In all these arrays the elements should be arranged in such a way that the feed points are located on a square measuring ¼ λ on the side.

7.1. Four Square with Bent Sloping Dipoles

In this configuration we use four sloping dipoles from a central tower. As under such circumstances these dipoles, due to the weight of the coaxial feed line, can never be perfectly straight, we have included this "sag" in our model. **Fig 11-178** shows a practical layout that I analyzed. Quadrature feeding does not result in a very good pattern. Through modeling we came to the following design:

Feed currents:

I1 = 1 ∠0° (back)

I2 = I3 = 1 ∠–139° (center)

I4 = 1 A, ∠–263° (front)

Gain: 4.56 dBi over good ground (approximately 4 dB over a single sloping vertical)

3-dB beamwidth: 87° (vs 108° if quadrature fed!)

RDF: 8.06 dB

DMF: 11.9 dB

The feed-point impedances calculated including a grounded central support tower are:

Z1 = 91 + j 117 Ω (front)

Z2 = Z3 = 40.7 Ω

Z4 = 31 + j 37 Ω (back)

These figures were obtained using high accuracy real ground with *NEC-4*.

As this array is not quadrature-fed, a Lahlum/Lewallen feed must be used (see Section 3.4.5.8).

7.2. The K8UR Sloping-Dipoles Four Square Array

D.C. Mitchell, K8UR, described his four-element sloping array (Ref 975). He uses half-wave sloping dipoles (sometimes called slopers) where the bottom half is sloped back toward the tower. This eliminates most of the high-angle radiation, as the much of the horizontal component is now canceled due to the folding of the elements.

The four feed points form a square measuring λ/4 on its side. The ends of the dipoles are folded back to the tower as shown in **Fig 11-179**, which also shows the radiation patterns.

Mitchell uses the hybrid-coupler feed system. In his design

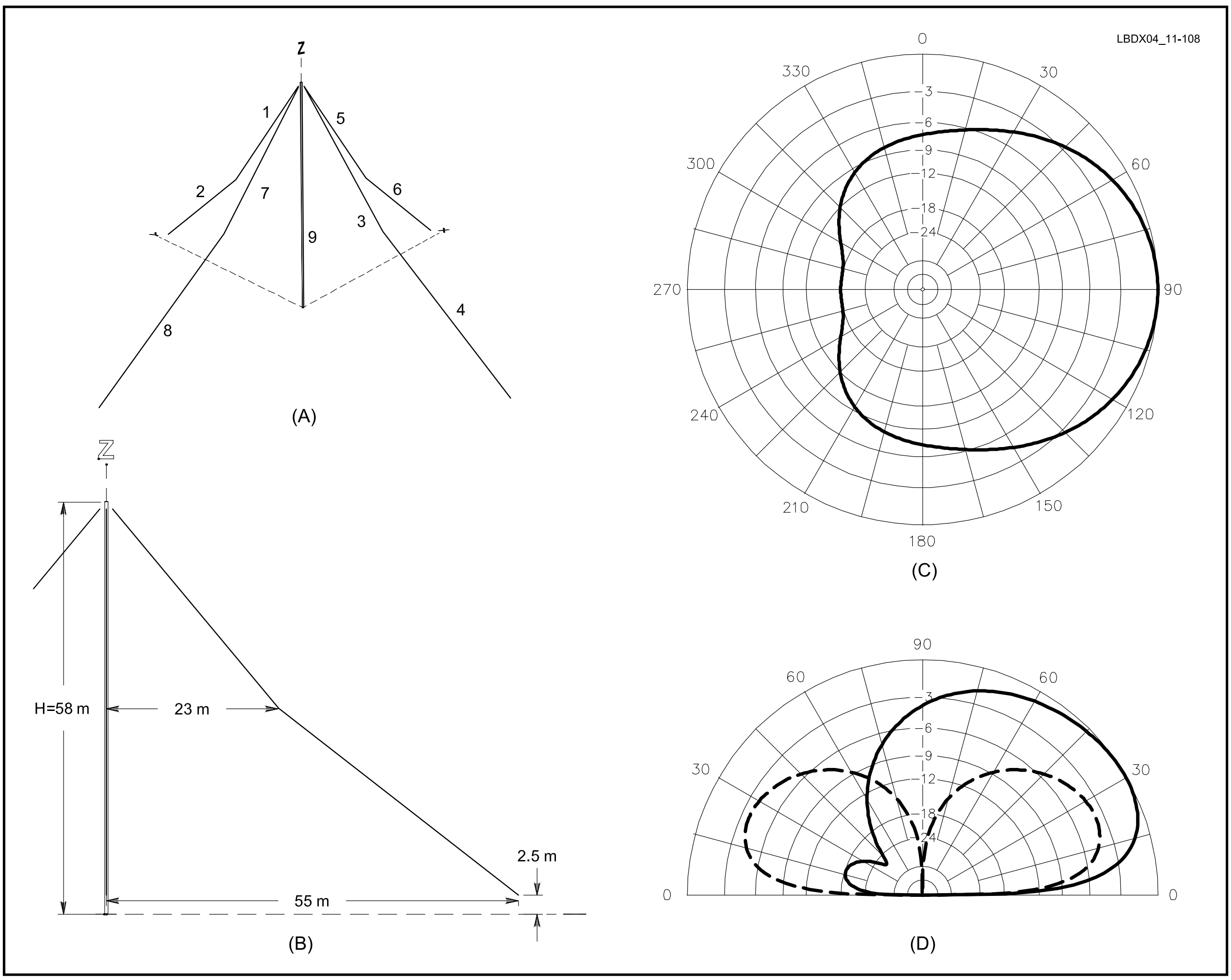


Fig 11-178 — Layout of 160-meter Four Square using half-wave dipoles suspended from a single tall tower. At B, layout of one of the sloping dipoles, showing the droop due to the weight of the feed line. At C and D, horizontal and vertical radiation patterns for this array. The vertical pattern is compared with that for a single vertical.

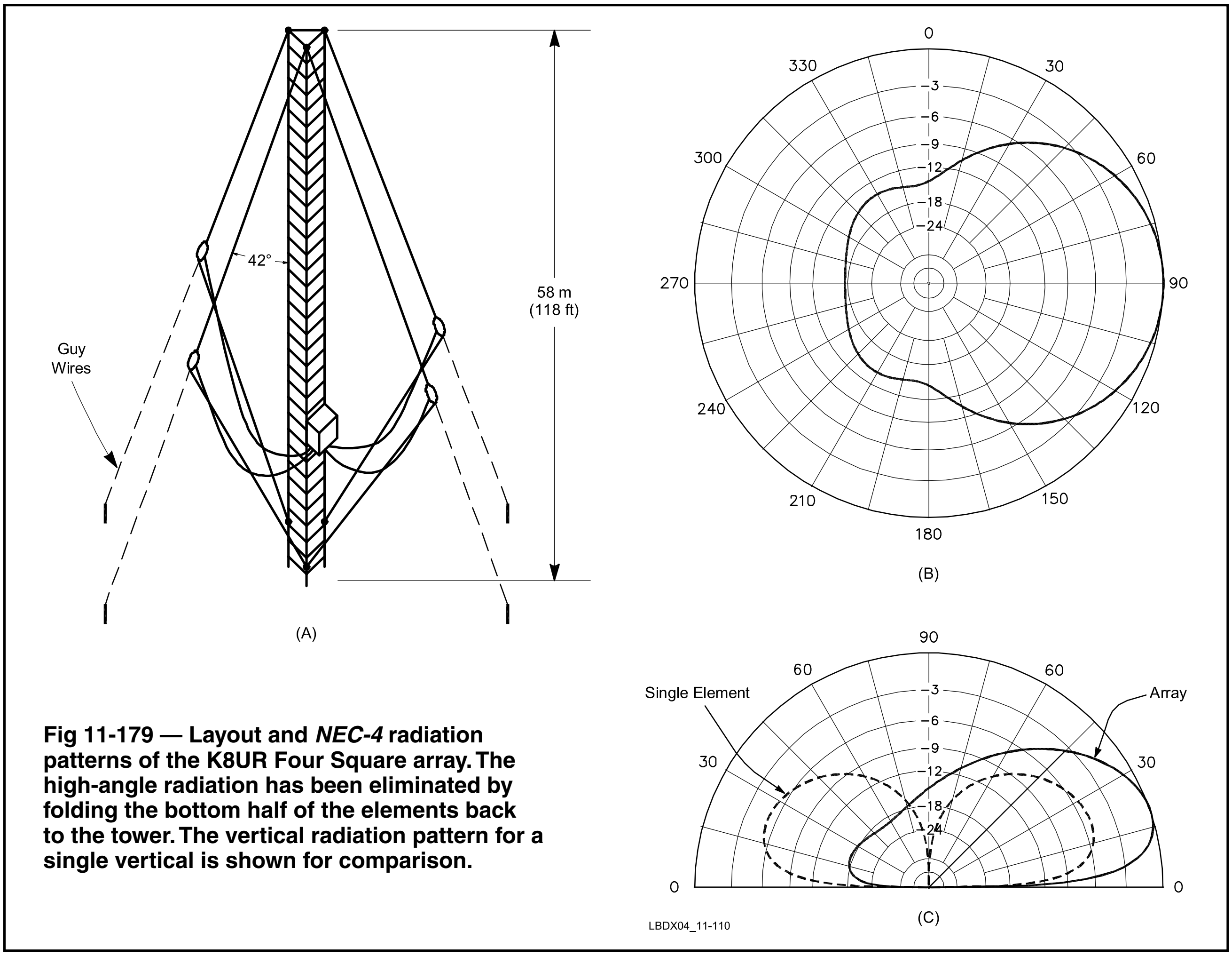


Fig 11-179 — Layout and *NEC-4* radiation patterns of the K8UR Four Square array. The high-angle radiation has been eliminated by folding the bottom half of the elements back to the tower. The vertical radiation pattern for a single vertical is shown for comparison.

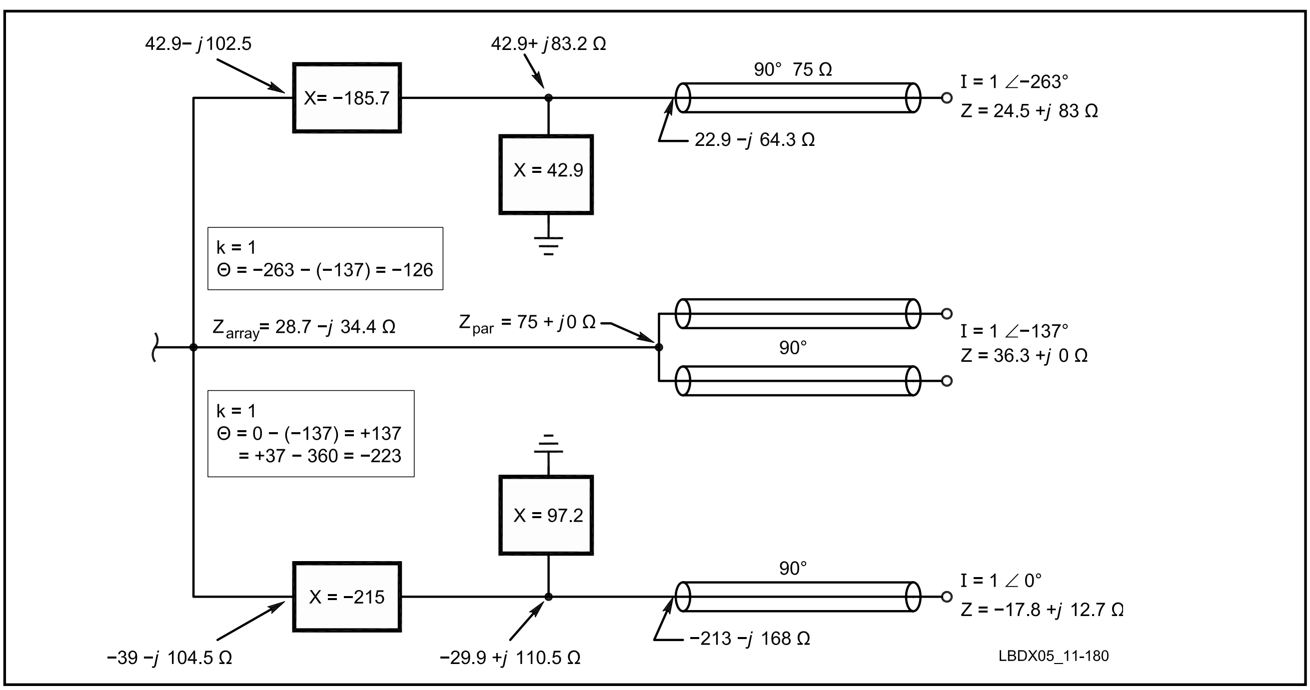


Fig 11-180 — One of the possible Lahlum/Lewallen feed systems (calculation including cable loss). In this case we feed the center elements directly. If we choose to feed the back element directly the array impedance turns out to be significantly lower (17 + *j* 28 Ω).

of the network, he has replaced the 180° phasing line with a 180° phase reversal transformer, taking care of the required 180° phase shift. These hybrid couplers were later commercialized by Comtek Systems.

7.2.1. Data, Four Square K8UR Fashion

Whereas the original design was fed in quadrature, we now know that much better directivity can be obtained with phasing angles that are much larger than 90°. Some modeling turned out a design with excellent properties:

Design data:

The applied feed currents were:

I1 = 1 ∠–263° (front)

I2 = I3 = 1 ∠–137° (center)

I1 = 1 ∠0° (back)

The feed-point impedances calculated with a grounded 58 meter support tower are:

Z1 = 24.5 + *j* 83 Ω (front)

Z2 = Z3 = 6.3 Ω (center)

Z4 = –17.8 + *j* 12 Ω (back)

The following data were obtained by modeling the array with *NEC-4* (high accuracy real ground):

Gain: 6.5 dBi (over average ground)

3-dB beamwidth: 88.4° (120° if quadrature fed!)

RDF = 11.07 dB

DMF = 22.7 dB

If you have a tall tower, a Four Square sloping array à la K8UR is the way to go! Clean pattern, excellent gain, good RDF and DMF because there is no high angle radiation — the horizontal radiation component coming from the sloping wires is canceled out by the folding back of the elements.

7.2.2. Feed System, Four Square K8UR Fashion

Fig 11-180 shows a Lahlum/Lewallen feed system for this array. This array lends itself to a feed design according to the crossfire principle (see Section 3.4.4 and **Fig 11-181**), as we end up with exactly 75 Ω with the two feed lines to the center elements paralleled.

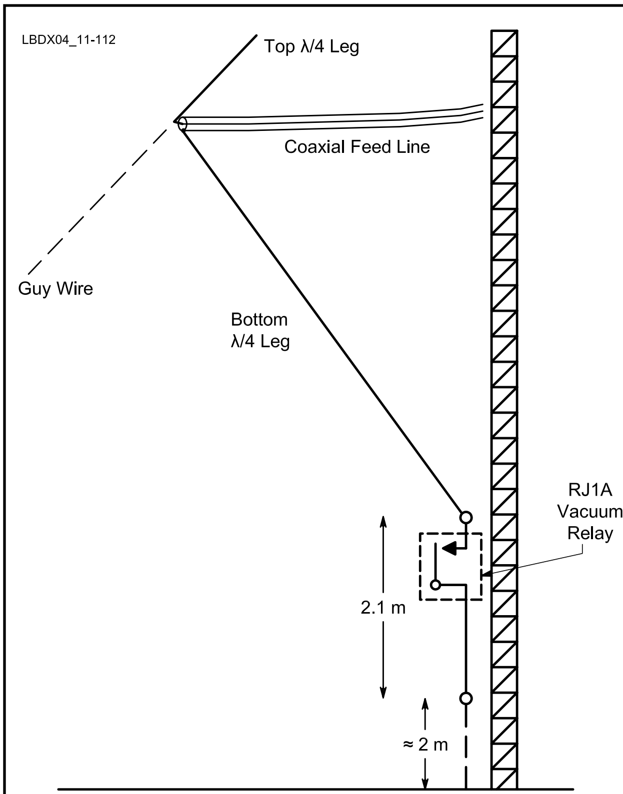


Fig 11-182 — Method for switching a K8UR-style dipole from 3.8 to 3.5 MHz. The bottom end of the dipole is lengthened with a piece of wire (which can be called capacitive loading) to lower its resonant frequency to 3.5 MHz. K4PI uses a 2.1-meter long wire, spaced about 0.5 meter from the tower. The relay must be able to withstand high voltage. Mike uses RJ1A vacuum relays.

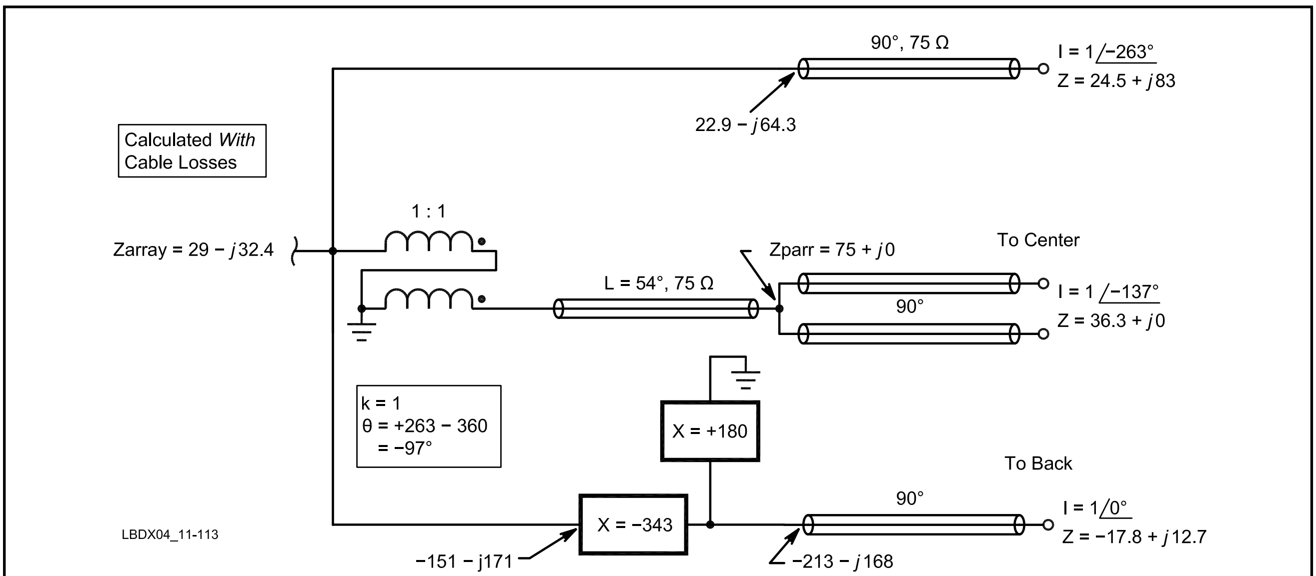


Fig 11-181 — Crossfire principle applied to the K8UR array with optimized phasing.

I inserted a 1:1 (180° phase shift) transformer in the center branch. The required phase shift vs the front element (the element being fed directly) is –137 – (–263) = +126°. The coax has to take care of –180 + 126 = –54°.

The back element is to be fed with a phase angle of +263° vs the front element, and that is the same as +263 – 360 = –97°

7.2.3. Switching From 3.8 to 3.5 MHz

Mike Greenway, K4PI, developed an innovative way for switching the dipoles of his K8UR-type array from the phone end of the band to the CW end of the band. See **Fig 11-182**.

7.3. The Four Square Array with Sloping Quarter-Wave Verticals

If you cannot run long catenary cables (as shown in Fig 11-167) to support your wire vertical elements, you can stick some insulating booms (fiberglass or aluminum broken up by insulators) in the top of your tower as shown in **Fig 11-183**. Make sure the central support does not upset the radiation patern. If it is grounded, it may be too close to resonance and you may have to decouple it (see Chapter 7, Section 3,10).

If you want to operate this array on both the CW and the phone ends of the band, it is best first to model the antenna at approximately 3.51 MHz. In that case the sloping vertical wires should be approximately 20.9 meters (assuming a 2 mm OD copper conductor). On 3.775 MHz the elements are too long and need to be shortened by a series capacitor of approximately 600 pF. A relay will be needed to switch the capacitor in and out of the circuit. At the same time the current-forcing feed line can be lengthened on 3.5 MHz to maintain its proper (λ/4) length. See also Fig 11-165.

It is clear that the same configuration can be used on 160 meters, where everything will be approximately twice the size.

7.3.1. Data, Four Square with Sloping λ/4 Verticals

Design data:
F = 3.75 MHz
Length sloping wires: 19.86 meters
Side of square: 20 meters
Feed currents:

- I1 = 1 ∠0°
- I2 = I3 = 1 ∠–120°
- I4 = 0.9 ∠–235°

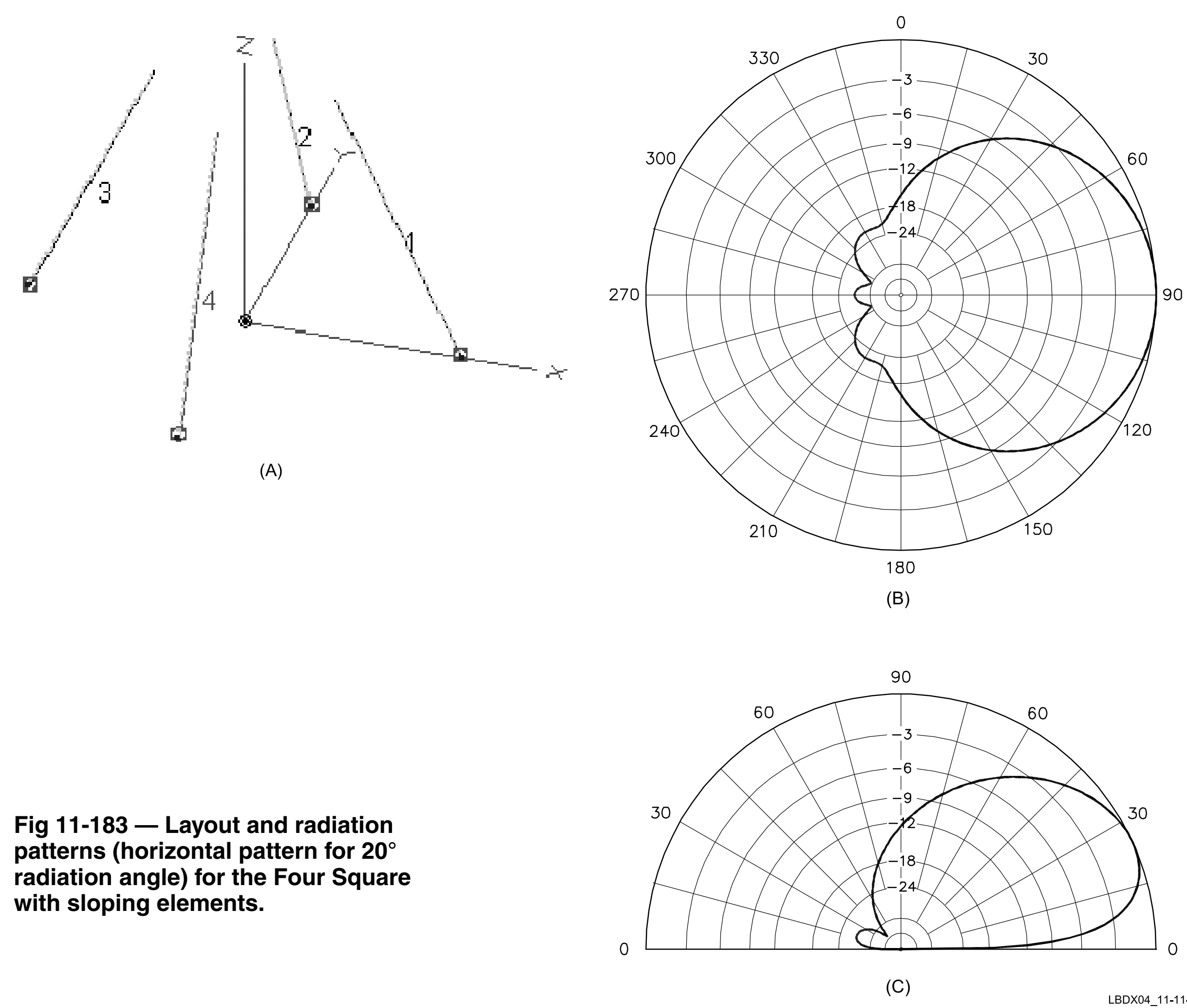


Fig 11-183 — Layout and radiation patterns (horizontal pattern for 20° radiation angle) for the Four Square with sloping elements.

The following data were obtained by modeling the array with *NEC-4* (high accuracy real ground):

Feed-point impedances:

Z1 = 24.5 + *j* 60 Ω (front)

Z2 = Z4 = 25.6 + *j* 5 Ω (center)

Z3 = –7 + *j* 5 Ω (back)

Gain: 6.04 dBi

3-dB beamwidth: 87°

RDF: 10.55 dB

DMF: 22.4 dB

7.3.2. Feed System, Four Square with Sloping λ/4 Verticals

Fig 11-184 shows a possible Lahlum/Lewallen feed system with optimized current magnitudes and phase angles, similar to the WA3FET design.

7.4. An Attractive 160-Meter Four Square with 18.5-Meter Tall Verticals

One does *not* need to use full-size quarter-wave long vertical radiators to build a good-performing array. Using short verticals has a few consequences though:

- Lower feed point impedances.
- Lower impedances mean that an excellent radial system is even more important.
- A low loss top-loading configuration is essential.

Top loading means wire loading. While inverted-L elements can be used, they will radiate a lot at high angle and largely destroy the RDF and DMF directivity figures of the array. Top loading wires need to be arranged in such a way that far-field cancellation occurs of all radiation off the loading wires.

The design shown in **Fig 11-185** is a good example.

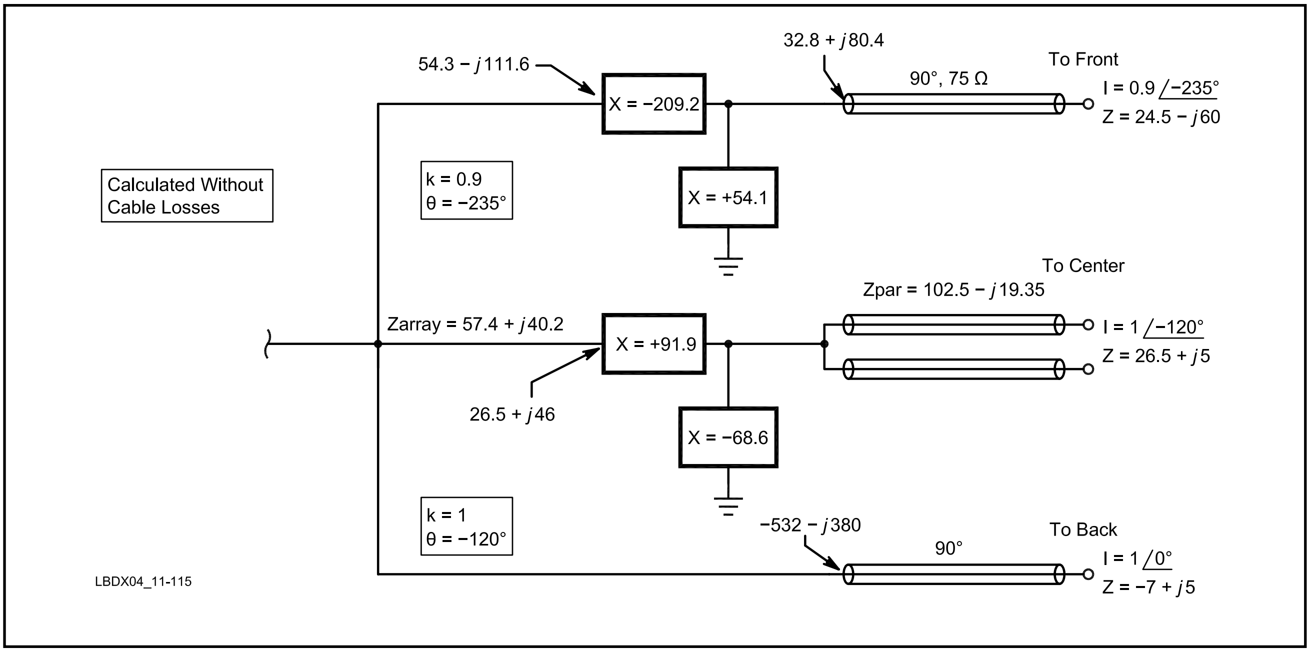


Fig 11-184 — One of the possible Lahlum/Lewallen feed methods for the Four Square with sloping elements, shown in Fig 1-183.

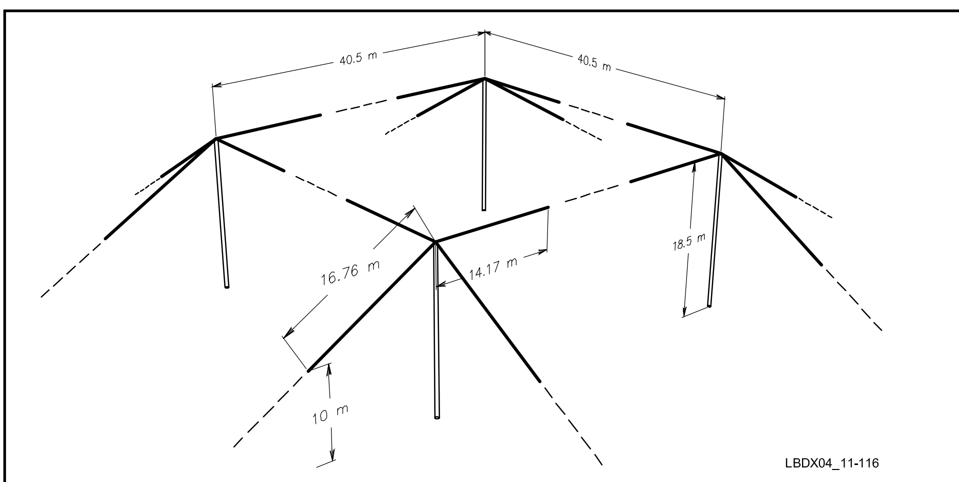


Fig 11-185 — This reduced height Four Square measures λ/4 along the side of the square. The lengths of the loading wires were calculated to eliminate any high angle radiation from these wires. Some fine tuning of the resonant frequency of the elements can be done by changing the height of the ends of the sloping loading wires (nominally 10 meters high).

Compared to a Four Square with full-size (39 meter long) elements, the quadrature fed version of this array with 18.5 meter long elements (46% of full-size) trades in only 0.5 dB in gain, assuming an identical radials system is used (with 2 Ω equivalent loss resistance).

7.4.1. Data, Reduced-Height Four Square, Quadrature Fed

Design data:
Side square: $\lambda/4$
Elements: 18.5 meter tall with top loading wires (see Fig 11-185)
Modeling file: *Ch11-4sq-160-short-el.ez* and *Ch11-4sq-160-short-el-incl-feedl.ez*
Feed currents (quadrature feeding):

I1 = 1 ∠–180° (front element)
I2 = I4 = 1 ∠–90° (center elements)
I3 = 1 ∠0° (back element)

Modeling (file *Ch11-4sq-160-short-el-incl-feedl.ez*) over average ground (ρ = 0.5 mS and ε = 13), including the real feed line losses (using RG-213) gave:
Gain: 5.51 dBi
3-dB beamwidth: 100°
RDF = 10.20 dB
DMF = 19.1 dB
The black box Z2 and Z3 (see Fig 11-41) are:

$Z2 = 55.9 + j\,15.2\ \Omega$
$Z3 = 55 - j\,44.5\ \Omega$

To include the parallel inductance of the phase reversal transformer we connect Z3 in parallel with $Z = 0 + j\,500$, which results in $Z3 = 65.4 - j\,41.3\ \Omega$

7.4.2. Optimized Hybrid Feed System, Reduced-Height Four Square, Quadrature Fed

As this array is quadrature fed you can feed it with an off-the-shelf hybrid coupler. Much better though is to apply one of the hybrid coupler optimization systems as described in Sections 3.4.6.5 and 3.4.6.6 and build your own hybrid coupler. Let us work out such an optimized hybrid coupler system according to W1MK's two-shunt phase compensation method described in Section 3.4.6.6.

In the exercise that follows we use the impedance data calculated by the modeling program (see above). If we want to

HYBRID COUPLER OPTIMIZATION TWO SHUNT COMPENSATION DESIGN SYSTEM -w1mk -				
	Enter data in yellow background cells			
	CALCULATING THE SHUNT ELEMENTS			
1	Enter fa (design freq) array →	1.830	MHz	
2	Enter k-value →	1		
		Real part	Imag part	
3	(-90 ° branch) Enter Z2 →	55.90	15.20	Ω
4	Branch 2 shunt element →	-220.78	Ω	
5	→	393.93	pF	
6	Z2' (at port 2) →	60.03	Ω	
		Real part	Imag part	
7	(0° branch) Enter Z3 →	65.40	-41.30	Ω
8	Branch 3 shunt element →	144.86	Ω	
9	→	12.60	uH	
10	Z3' (at port 3) →	91.48	Ω	
	HYBRID INPUT DATA			
11	Enter Zo Hybrid →	75.0	Ω	
	OUTPUT DATA			
12	Frequency corr. factor =	1.2360		
13	New Hybrid fo =	2.262	MHz	
14	L-Hybrid =	5.28	uH	
15	C-Hybrd =	469	pF	
16	Dump port (port 4) power ratio =	-44.5	dB	
17	Real part Zin (port 1) =	70.80	Ω	
18	Imag. part Zin (port1) =	15.21	Ω	
19	Port 1 return loss (dB) =	19.3	dB	
20	Port 1 SWR =	1.24		

Fig 11-186 — Calculation sheet for the phase compensated hybrid feed system. See text for details. By changing Z_0 from 50 to 75 Ω we improved the port 4 dump power situation by 30 dB!

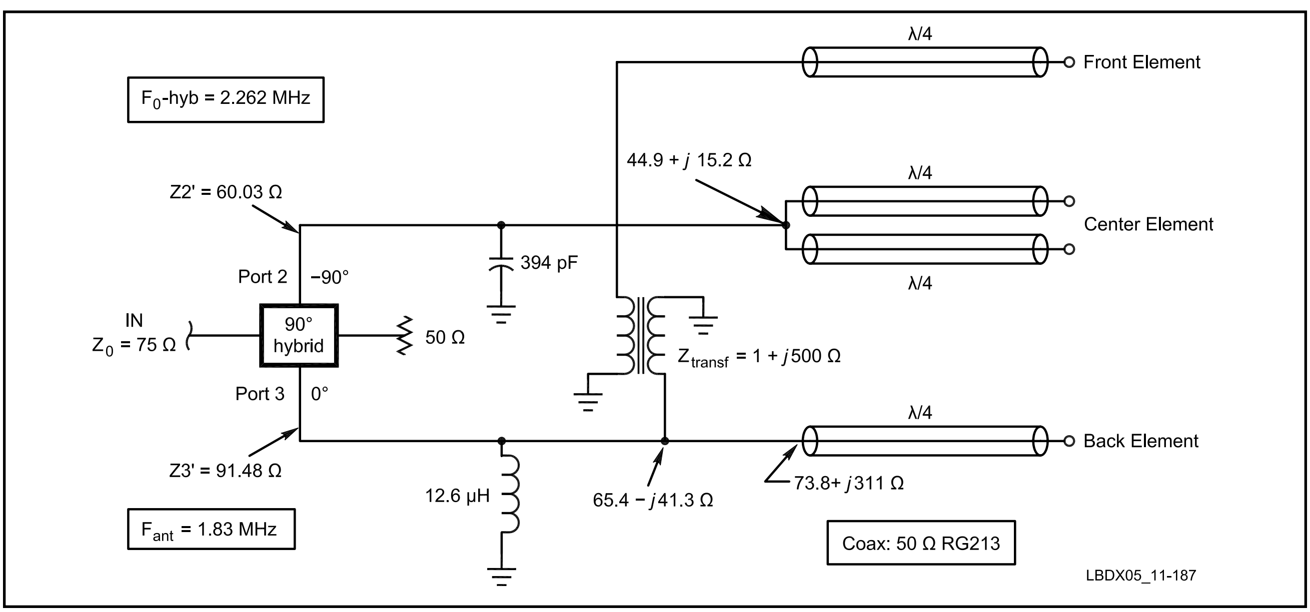


Fig 11-187 — Final circuit for the optimized hybrid coupler feed system, using the phase compensation method developed by W1MK.

HYBRID COUPLER DESIGN (by W1MK)			
INPUTS			
ENTER fa (operating frequency) →	1.83	MHz	
ENTER fo (hybrid frequency) →	2.262	MHz	
ENTER Zo Design impedance hybrid →	75	Ω	
	Real Part	Imag Part	
ENTER Impedance load PORT 2 (R2) →	60.06	0	Ω
ENTER Impedance load PORT 3 (R3) →	91.49	0	Ω
RESULTS			
fo/fa =	1.236		
Hybrid L value (uH) =	5.28	µH	
Hybrid C value (pF) =	469	pF	
Ratio Voltage magnitude Port2/Port3 (k) =	1.000		
Phase angle Voltage port 2 vs. port 3 =	-90.00	°	
Power in Port 4 (vs. Pwr in Port 1) =	-44.7	dB	
Real part input impedance (port 1) =	70.12	Ω	
Imagainare part input impedance (port1) =	15.05	Ω	
Return loss (port 1) =	-19.3	dB	
SWR =	1.24		

(X4) SHUNT REACTANCE CALCULATOR		
ENTER Shunt cap. →	200.00	pF
Shunt X =	-351.80	Ω
ENTER Shunt Coil→	3.00	uH
Shunt X =	34.49	Ω
Above is a simple calculator to change component (L and c) values into reactance values. Below: the results with different Z4 and possible X4		
PORT 4 LOAD		
ENTER port 4 load (Z4) →	50.00	Ω
ENTER port 4 shunt react (X4) →	999999.00	Ω
NEW RESULTS		
k =	1.000	
θ =	-89.86	°
port4 loss=	-44.8	dB
Port 1 Rin =	70.12	Ω
Port 1 Xin =	15.05	Ω
Return loss (port 1) =	-19.3	dB
SWR =	1.24	

Fig 11-188 — The *4sq-hyb-w1mk.xls* spreadsheet shows the final design data using Z_0 = 75 Ω. As we can see on the right the fact that we use a 50 Ω dummy load only changes θ by about 1°, which is quite acceptable.

design the feed system for an array that was built, we should measure Z2 and Z3, and apply the two-port coupling software as shown in Fig 11-45 to calculate the real Z2 and Z3.

The spreadsheet *two-shunt-hybrid-comp.xls* calculates the required shunt elements to achieve real (nonreactive) impedance loads at port 2 and 3 of the hybrid (see **Fig 11-186**). Note that the lowest port 4 dump power (–44.5 dB!) is achieved with Z_0 = 75 Ω (–14.3 dB with Z_0 = 50 Ω). This will, however, require a series capacitor (X_C = –15.45 Ω) plus a 1.5:1 impedance ratio transformer to match the hybrid input impedance to 50 Ω. To achieve the proper voltage magnitude ratio (1:1), f_o = 2.262 MHz. **Fig 11-187** shows the circuit of the feed system.

All of this can be calculated using the *4sq-hyb-w1mk.xls* spreadsheet (**Fig 11-188**).

7.4.3. Reducing the Current in the Center Elements

We learned in Section 3.4.6.5.8 that we can modify the classical quadrature configuration and reduce the magnitude of the feed current to the center elements in order to improve the directivity. If we reduce the center-element feed currents to (in this case) as little as 75%, we can get rid of the big high angle back lobe and turn it into two smaller back lobes. That significantly improves the array directivity (see Fig 11-53).

The feed voltages at Z2 and Z3 are:

V2 = 0.75 V ∠–90°

V3 = 1 V ∠0°

We have learned in Section 3.4.6.4.1 that in a perfectly symmetrical Four Square, the branch impedances Z2 and Z3 are independent of the source voltages (applied to Z2 and Z3), which means that for the optimized version we can use the same Z2 and Z3 values as above (see also modeling file *Ch11-4sq-160-short-el-optim-incl-feedl.ez*).

If we plug these values in the *two-shunt-hybrid-comp.xls* spreadsheet, and specify k = 0.75, we end up with exactly the same values as calculated in Section 7.4.2, the only difference being a change in f_o (New f_o = old f_o × k_{new}/k_{old} or 2.262 × 0.75 = 1.696 MHz.) This is exactly what the software program (see **Fig 11-189**) calculates. Note that with Z_0 = 75 Ω we reach –44 dB dump power, but we will require a 75 to 50 Ω transformer of some kind at the input. With Z_0 = 50 Ω the dump power rejection is still –14.3 dB, which is quite acceptable.

	HYBRID COUPLER OPTIMIZATION TWO SHUNT COMPENSATION DESIGN SYSTEM -w1mk -			
	Enter data in yellow background cells			
	CALCULATING THE SHUNT ELEMENTS			
1	Enter fa (design freq) array →	1.830	MHz	
2	Enter k-value →	0.75		
		Real part	Imag part	
3	(-90 ° branch) Enter Z2 →	55.90	15.20	Ω
4	Branch 2 shunt element →	-220.78	Ω	
5	→	393.93	pF	
6	Z2' (at port 2) →	60.03	Ω	
		Real part	Imag part	
7	(0° branch) Enter Z3 →	65.40	-41.30	Ω
8	Branch 3 shunt element →	144.86	Ω	
9	→	12.60	uH	
10	Z3' (at port 3) →	91.48	Ω	
	HYBRID INPUT DATA			
11	Enter Zo Hybrid →	50.0	Ω	
	OUTPUT DATA			
12	Frequency corr. factor =	0.8888		
13	New Hybrid fo =	1.627	MHz	
14	L-Hybrid =	4.89	uH	
15	C-Hybrd =	978	pF	
16	Dump port (port 4) power ratio =	-14.3	dB	
17	Real part Zin (port 1) =	49.79	Ω	
18	Imag. part Zin (port1) =	12.49	Ω	
19	Port 1 return loss (dB) =	18.1	dB	
20	Port 1 SWR =	1.28		

Fig 11-189 — Calculation sheet for the phase compensated hybrid feed system using k = 0.75, which results in better directivity. Note that only the f_0-hybrid has changed and that the hybrid input impedance has only changed slightly.

7.4.4. From Virtual to Real World

Refer to Section 6.6.2. and follow all the steps described in this section. The only difference with the example from Section 6.6.2 is that we now calculate *two* shunt elements.

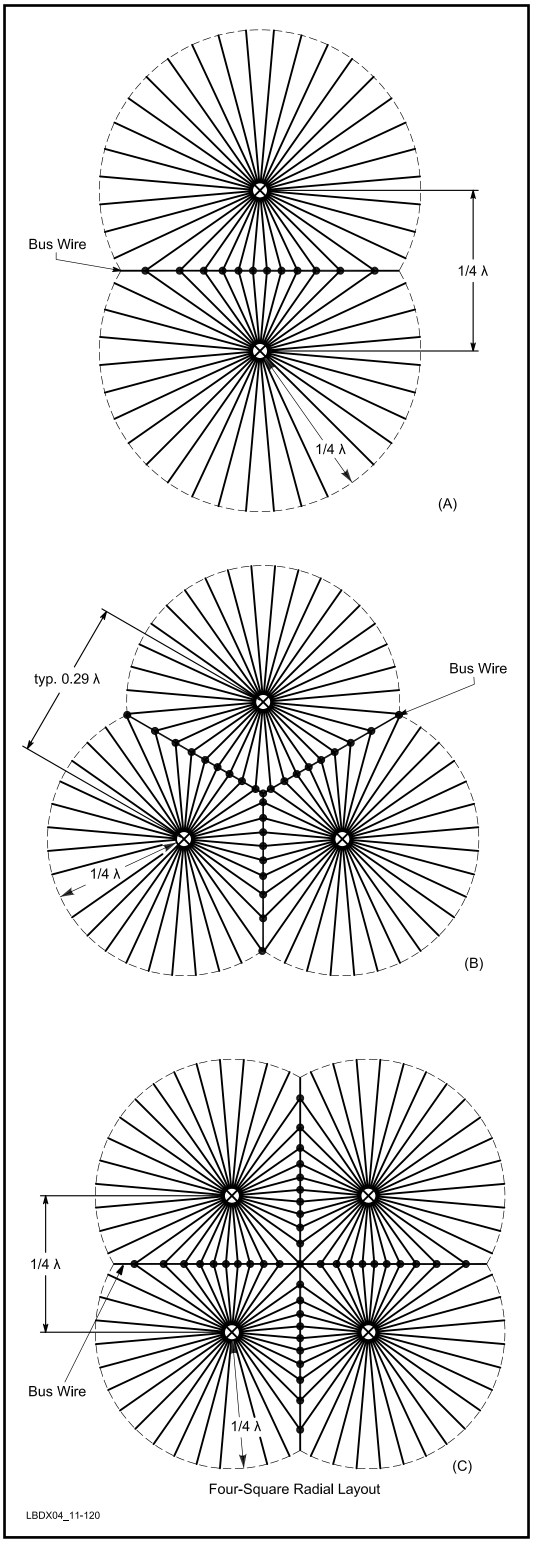


Fig 11-190 — Recommended radial layouts for multi-element arrays.

7.5. Low Cost, Reduced-Size Array for 80/160 Meters

Most of us cannot even dream of putting up a gain array on 160 meters. At least, we think we can't. Robye, W1MK, made his dream come true and has a 2-element end-fed array on 160 (switchable in four directions), that you can hardly see in the back of his yard. In addition, the same array is used on 80 meters as a Four Square.

In between the trees, he found just enough room to make an 80-meter Four Square, using top loaded elements that are only 13.3 meters long. He uses four umbrella-shaped top loading wires, each about 20 feet long, sloping at an angle of approximately 30°. This brings the elements in resonance at approximately 3.4 MHz. A single vertical of this configuration will do almost as well as a full-size quarter-wave vertical, provided one can keep the ground losses low enough. One of the major problems with such a short radiator is the operational

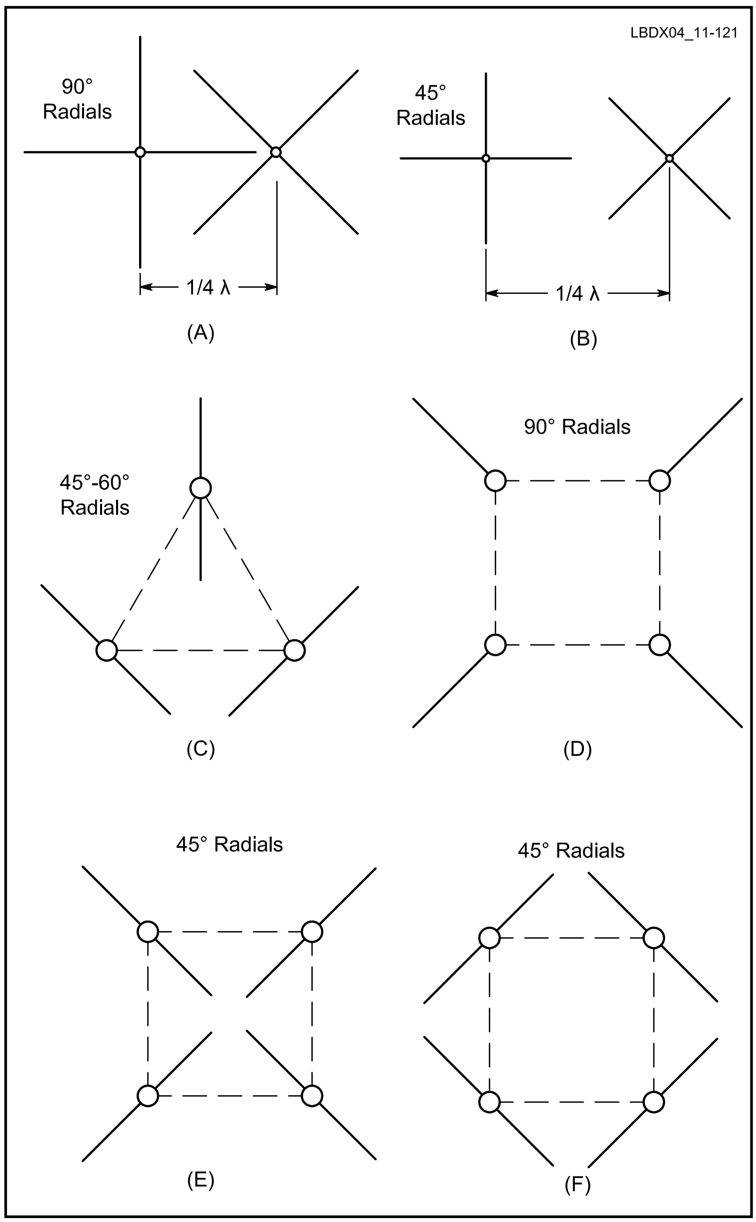


Fig 11-191 — Recommended elevated radial layout schemes.

bandwidth, but as W1MK hardly ever works phone, a coverage of 3.5 to 3.56 MHz was good enough for him.

This 80-meter array was developed in the late 1980s long before Robye developed the optimized hybrid circuit feed systems described in this book (Section 3.4.6.6). The logical choice, given the required bandwidth was to use a Lewallen-type L-network feed system. In a detailed article (available on the CD which comes with this book — *2 band vert array W1MK.pdf*) Robye describes in detail the WA3FET configuration (see Section 4.7.2) he used in order to obtain best gain and directivity, as well as how he uses his hybrid coupler measurement setup (see Section 3.6) to tune his array for optimum performance on 80 meters.

Building a good performing and economical Four Square array with elements that are merely 13 meters tall is quite a performance, but using these same shorties to make an array for 160 meters is even more of a challenge. Robye did it, successfully. Robye uses only two of the diagonally positioned elements and leaves the two other elements floating (resonant on 80 meters, so no coupling at all). The spacing is about 30 meters (0.19 λ), which is just fine.

The main issue is the low radiation resistance. The 80-meter elements modeled $5 - j\,315\ \Omega$, and measured approximately $10 - j\,315\ \Omega$, whereby the resistive part would vary somewhat between summer and winter. This low impedance and the relatively high loss make it impossible to get more gain from a Four Square (on a square measuring only 21 meters on its side) than from a 2-element end-fire array with 30-meter spacing. The 2-element end-fire configuration actually results in a gain that is greater than if the antenna were configured as a ⅛ λ spaced Four Square (with the same short elements). So Robye decided to make a 2-element end-fire array that could be aimed in four directions.

As a receiving array, this 2-element however suffers somewhat relative to the four-element array because of the pattern shape. That made W1MK decide to use the Four Square with resistive loading (up to 75 Ω) as a receive antenna on 160 meters (see Chapter 7, Section 1.21 and following). Robye confirms that the gain is low as expected, but the pattern is good.

All of this involves a lot of high speed vacuum relays, phase matching boxes, and a lot of coax. But when you are limited in the space available you can compensate by using more hardware. That is the concept of what W1MK does successfully on 80/160 meters. The details of all of that are available on the CD (file: *2 band vert array W1MK.pdf*).

Of course, you cannot expect an array with size restriction as described above to be a winner when it comes to operational bandwidth. It is clear that any sort of loading that is designed to make an antenna shorter will also compromise bandwidth.

So, if bandwidth is a primary concern, go for an array with:

1) Full size, "fat" elements.
2) Wide spaced elements (¼ λ spacings or wider).
3) Fewest components in the phasing system.

8. RADIAL SYSTEMS FOR ARRAYS

8.1. Buried Radials

Radials of the elements of an array cross each other. It is standard procedure to install a bus (#6 or #8 AWG copper strip) halfway between the elements and to connect the radials to this bus. The radials can be any wire size, if many are used. The size will be dictated more by mechanical strength than current carrying capability. **Fig 11-190** shows various typical radial layouts for a 2-element cardioid array, a 3-element triangular array and for the classic Four Square array.

8.2. Elevated Radials

When many elevated radials are used on each of the elements of an array, these radials become nonresonant, and they can be connected to a bus system in exactly the same way as shown in **Fig 11-191**.

When only a few radials are used (typically one to four radials), the situation is very different. In this case the radials can couple heavily with adjacent (especially parallel) radials from other elements and can upset the directivity of the array, and create uncontrolled and unwanted high-angle radiation from the elevated radials.

Fig 11-191 shows a few possible layouts that try to minimize the coupling.

9. CONCLUSION

Many years ago, most antenna builders knew only one approach to making their own antenna perform: cut and try, measure (the few things he could measure, like SWR), and evaluate on the air (no reference to compare it against). Most of us had only very limited technical means at our disposal.

Today serious antenna builders use the "three legged stool" principle to be successful. The three legs of this stool that support performance are: design and analysis, modeling, and measuring.

LBDX05_11-192

Design and Analysis

To successfully design an antenna you must know what makes an antenna work. Over the years our understanding of how and why antennas work has improved greatly. Part of that is based on better understanding and part on modern analysis techniques. We've always known these darn things (sometimes) worked (often, though, we liked to think they worked), but we did not always know (mostly not) exactly why and how. Understanding down to the nitty gritty detail the why and how may not be for the average amateur, but using the results of knowledge and analysis and translating this knowledge into practical design tools that even the non-professional engineer can use is important. This is, to a large extent, what this book is all about. No fairy tales, no guesswork but knowledge and common sense.

In every edition of this book I have been able to describe evolutions in the field of antenna array concept and design, as well as in array feed systems. I am very much indebted to each and every one of you who, over the years have been willing to share your innovations with the Amateur Radio community through this book. This Chapter 11, as well as others, is full of novelties!

Modeling

Until now I had never seen a modeling tool, either in amateur literature or in professional literature, that allowed me to model a hybrid coupler of the type we use with our antenna arrays. These hybrids were mysterious black boxes. Thanks to Robye Lahlum, W1MK, we can for the first time introduce such a tool in this book. It is a comprehensive modeling tool that does two things: it makes us understand how a hybrid coupler works, and helps us design a hybrid coupler system that works as it should.

Now that we all know what makes the hybrid coupler work, the next step is to be able to model the hybrid as part of an antenna system, in order to be able to easily generate bandwidth performance data. For years *EZNEC* has made it possible for an antenna model to include a number of components such as feed lines, transformers, resistors, inductors and capacitors in an antenna system. This makes it possible to do an operational antenna bandwidth assessment of the antenna system.

Especially for this book, Roy Lewallen, W7EL, designed a hybrid that uses standard components (L, C and a transformer) and that can be included in the *EZNEC* antenna model of quadrature fed arrays using a lumped coupled 90° hybrid. This allows us to do operational bandwidth assessment for these types of arrays, and is, together with the unveiling of all the mysteries surrounding the hybrid coupler, a giant step forward in better understanding and more thorough knowledge of amateur arrays fed by a hybrid coupler.

Measuring

Measuring is knowing. A long time ago we had a grid dipper, and when coaxial cable came about, we had SWR meters. These were the tools the hams had to test their antennas. Today technically inclined hams have access to affordable measuring and test equipment that can compete with the most expensive professional equipment. It is literally impossible to adjust a Four Square for optimum performance without some quality test equipment. A top range antenna analyzer can take you a long way, but a real VNA with a multiplexer box and vector scope software will allow you to measure and know every detail of your antenna array. Never in the past have hams had access to such high-performance tools. Why do we need these tools?

We need these tools to evaluate the results of our design and construction efforts. For antenna arrays the ultimate test is in measuring operational parameters and characteristics. And when we see that our test results closely match our initial modeling results, it proves we had a good model, that we performed accurate and dependable measurements and finally that we have an array that works like we designed it. These matching results cannot be coincidence. It proves you've done a good job.

We have come to the point where even the average ham can create an array that in design, modeling and measurement (and therefore, performance), exceeds the best professional work from even 30 years ago.

But tools, models and formulas aren't everything. Sometimes the world may even have too many of those, but way too little common sense and wisdom.

Contents

CHAPTER 12

Other Arrays

Chapter 11 on phased arrays only covered arrays made of vertical (omnidirectional) radiators. You can, of course, design phased arrays using elements that, by themselves, already exhibit some horizontal directivity, horizontal dipoles for example.

Even at relatively low heights (0.3 λ), arrays made of horizontal elements (dipoles) can be quite attractive. Their intrinsic radiation angle is certainly higher than for an array made of vertical elements, but unless the electrical quality of the ground is good to excellent, the horizontal array may actually outperform the vertical array even at low angles.

The vertical radiation angle (wave angle) of arrays made with vertical elements (typical λ/4 long elements) depends only on the quality of the ground in the Fresnel zone. Radiation angles range typically from 15° to 25°. The same is true for arrays made with horizontally polarized elements, but we have learned that reflection efficiency is better over bad ground with horizontal polarization than it is with vertical polarization (see Chapter 9, Section 1.1.2 and Chapter 8, Section 1.2.1.1).

The elevation angle for antennas with horizontally polarized elements basically depends on the height of the antenna above ground. For low antennas (with resulting high elevation angles), the quality of the ground right under the antenna (in the near field) will also play a role in determining the elevation angle (see Chapter 8, Section 1.2.1.4). But as DXers, we are not interested in antennas producing wave angles that radiate almost at the zenith.

Over good ground, a dipole at λ/4 height radiates its maximum energy at the zenith. Over average ground, the wave angle is 72°. The only way to drastically lower the radiation angle with an antenna at such low height is to add another element.

If we install a second dipole at close spacing (eg, λ/8), and at the same height (λ/4), and feed this second dipole 180° out-of-phase with respect to the first dipole, we achieve two things:

- Approximately 2.5 dB of gain in a bidirectional pattern.
- A lowering of the elevation angle from 72° to 37°!

At the zenith angle the radiation is a perfect null, whatever the quality of the ground is. This is because at the zenith the reflected wave from element number 1 (reflected from the ground right under the antenna) will cancel the direct wave from element number 2. The same applies to the reflected wave from element number 1 and the direct 90° wave from element number 2. All the power that is subtracted from the high angles is now concentrated at lower angles. Of course there also is a narrowing of the horizontal forward lobe. Example: A λ/2 80-meter dipole at a height of 25 meters has a –3 dB forward-lobe beamwidth of 124° at an elevation angle of 45°. The 2-element version, described above, has a –3-dB angle of 95° at the same 45° elevation angle. The impedance of the two dipoles has dropped very significantly to approximately 8 Ω.

Fig 12-1 shows the elevation angles for three types of antennas over average ground: a horizontal dipole, two half waves fed 180° out-of-phase (spaced λ/8), and a 2-element Yagi. From this graph you can see that the only way to achieve a reasonably low radiation angle from a horizontally polarized antenna at low height of λ/3 or less is to add a second element. The 180° out-of-phase element lowers the radiation angle at lower antenna heights (below 0.35 λ) significantly more than a Yagi or a 2-element all-fed array. It also has the distinct advantage of suppressing all the high-angle radiation, which is not the case with the Yagis or all-fed arrays.

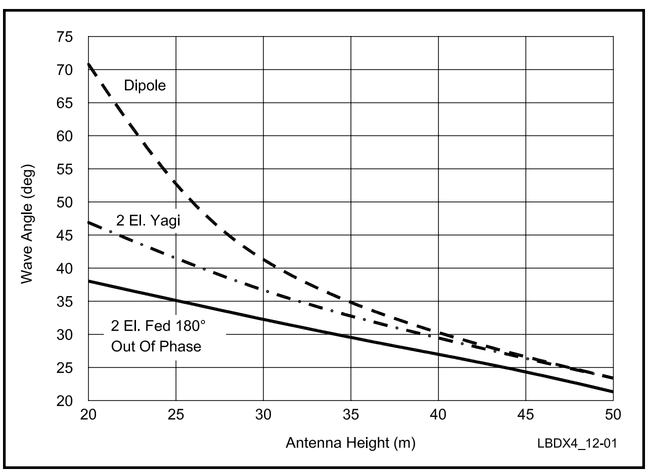
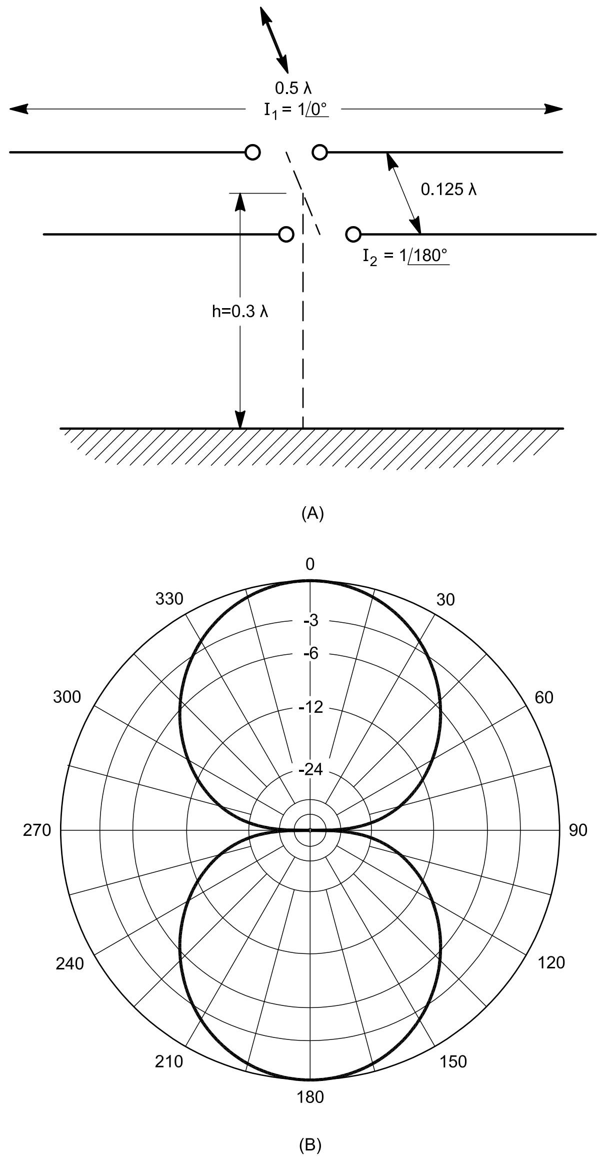


Fig 12-1 — Vertical elevation angle (wave angle) for three types of antennas over average ground: a half-wave dipole, a 2-element parasitic Yagi array and two close-spaced half-wave dipoles fed 180° out-of-phase. Note the remarkable superiority of the last antenna at low heights. The graph is applicable for 80 meters.



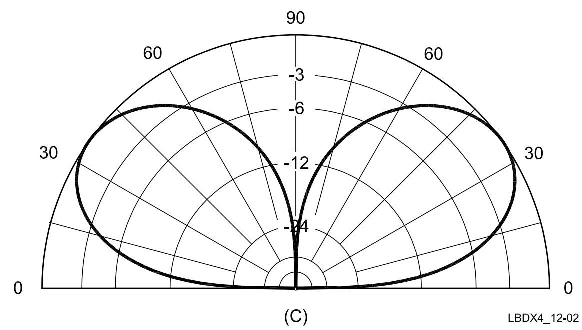


Fig 12-2 — Configuration and radiation patterns of two close-spaced half-wave dipoles fed 180° out-of-phase at a height of 0.3 λ above average ground. The azimuth pattern at B is taken for an elevation angle of 36°. Note in the elevation pattern at C that all radiation at the zenith angle is effectively canceled out (see text for details).

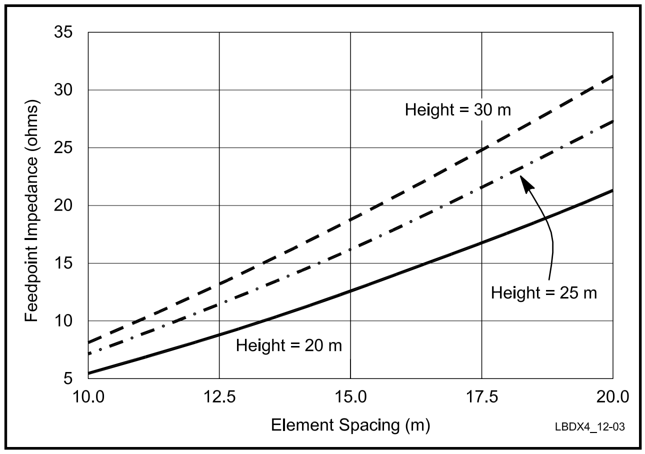
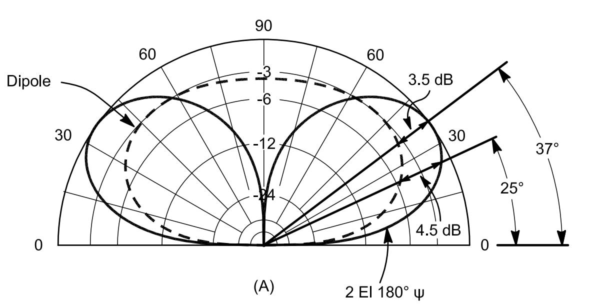


Fig 12-3 — Feed-point impedance of the 2-element close-spaced array with elements fed 180° out-of-phase, as a function of spacing between the elements and heights above ground. The design frequency is 3.75 MHz.



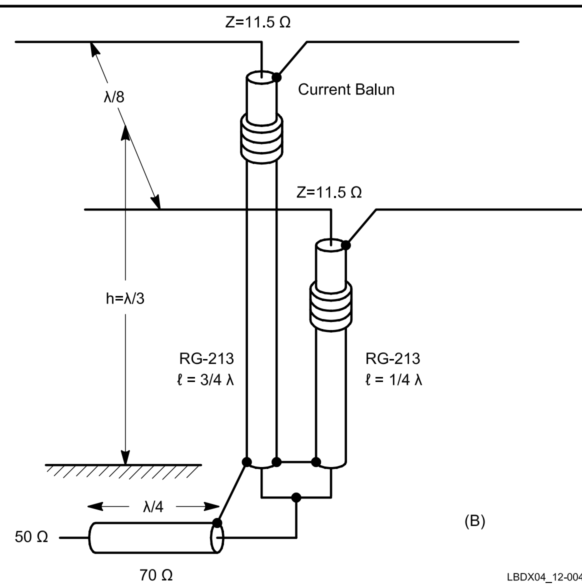


Fig 12-4 — At A, vertical radiation pattern of the 2-element close-spaced array compared to a single dipole at the same height of 0.3 λ (25 meters for 3.8 MHz). The feed method for a spacing of λ/8 is shown at B. The feed-point impedance is about 100 Ω at the junction of the λ/4 and the 3λ/4 50 Ω feed lines. A λ/4 long 70-Ω feed line can be used to provide a perfect match to a 50 Ω feed line.

1. TWO-ELEMENT ARRAY SPACED λ/8, FED 180° OUT-OF-PHASE

The vertical and the horizontal radiation patterns of the 2-element array are shown in **Fig 12-2**. As the antenna elements are fed with a 180° phase difference, feeding is simple. The impedances at both elements are identical. **Fig 12-3** gives the feed-point impedance of the elements as a function of the spacing between the elements and the height. Within the limits shown, spacing has no influence on the gain or the directivity pattern. Very close spacings give very low impedances, which makes feeding more complicated and increases losses in the system. A minimum spacing of 0.15 λ is recommended.

Compared to a single dipole at the same height, this antenna has a gain of 3.5 dB at its main elevation angle of

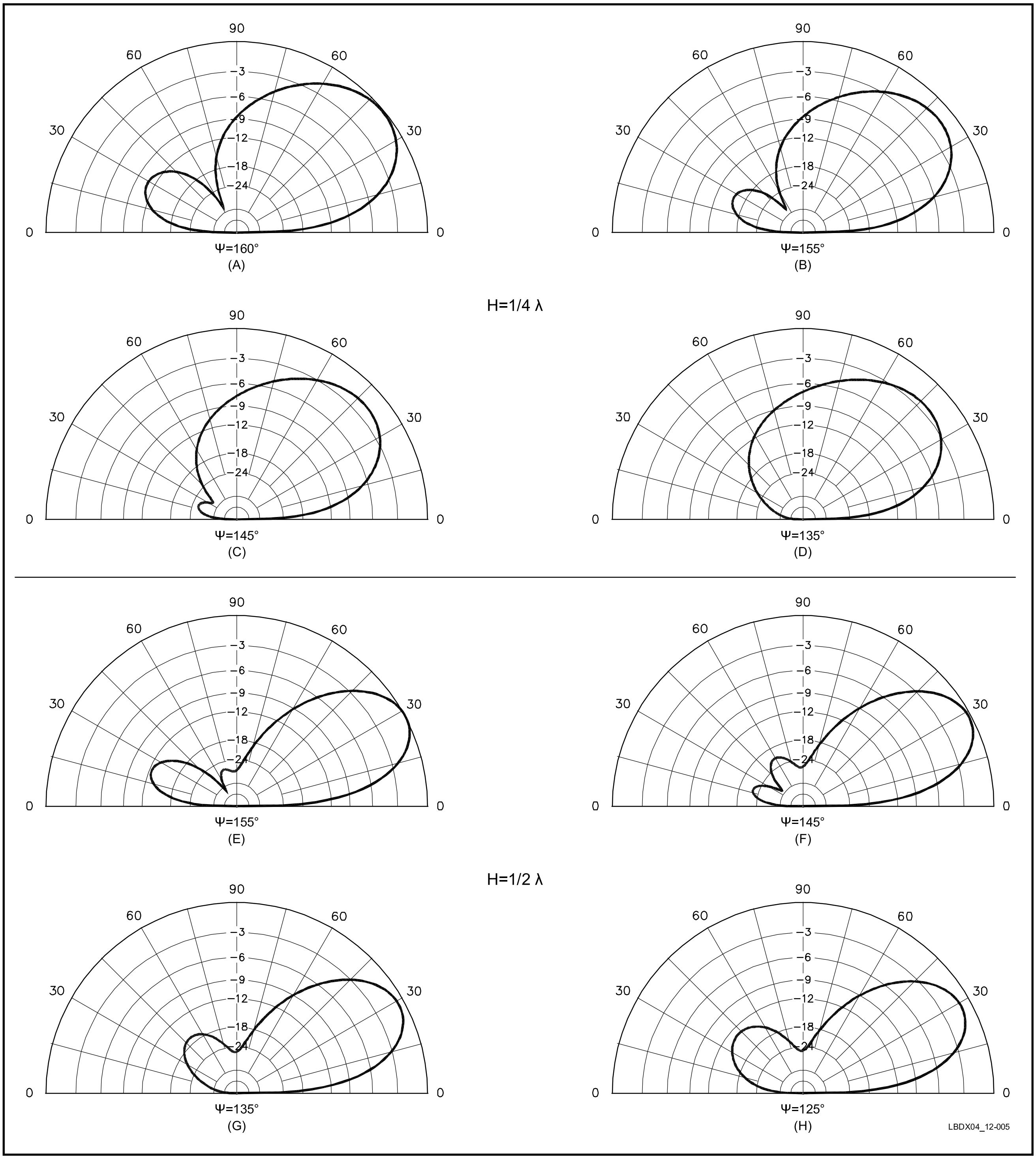


Fig 12-5 — Vertical radiation patterns of the 2-element all-fed array for different phasing angles. The current magnitude is the same for both elements. All patterns are plotted to the same scale. Patterns are shown for antenna heights of λ/4 (at A through D) and λ/2 (at E through H).

37°, and of 4.5 dB at an elevation angle of 25° (see **Fig 12-4**).

Feeding the array is done by running a λ/4 feed line to one element, and a 3λ/4 feed line to the other element. The feed point at the junction of the two feed lines is approximately 100 Ω for an element spacing of 0.125 λ. A λ/4 long 75-Ω cable will provide a perfect match to a 50-Ω feed line.

You will have a 5:1 SWR on the two feed lines, so be careful when running high power! Another feed solution that may be more appropriate for high power is to run two parallel 50-Ω feed lines to each element, giving a feed line impedance of 25 Ω. In this case the SWR will be a more acceptable 2.2:1 on the line. At the end of the feed lines (λ/4 and 3λ/4) the impedances will be 54 Ω. The parallel combination will be 27 Ω, which can be matched to a 50-Ω line through a quarter-wave transformer of 37.5 Ω (two parallel 75-Ω cables) or via a suitable L-network.

2. UNIDIRECTIONAL TWO-ELEMENT HORIZONTAL ARRAY

Starting from the above array, we can now alter the phase of the feed current to change the bidirectional horizontal pattern into a unidirectional pattern. The required phase to obtain beneficial gain and especially front-to-back ratio varies with height above ground. At λ/2 and higher, a phase difference of 135° produces a good result. At lower heights, a larger phase difference (eg, 155°) helps to lower the main wave angle. This is logical, as the closer we go to the 180° phase difference, the more the effect of the phase radiation cancellation at high angles comes into effect.

Fig 12-5 shows the vertical radiation patterns obtained with different phase angles for a 2-element array at λ/4 and λ/2. Note that as we increase the phase angle, the high-angle radiation decreases, but the low-angle F/B worsens. The higher phase angle also yields a little better gain. For antenna heights between λ/4 and λ/2, a phase angle of 145° seems to be a good compromise.

Feeding these arrays is not so simple, since the feed-current phase angles are not in quadrature (phase angle differences in steps of 90°). For a discussion of feed methods see Chapter 11 on vertical arrays. *Current forcing* using a modified Lewallen feed system seems to be the best choice.

The question that comes to mind is, "Can we obtain similar gain and directivity with a parasitic array?" Let's see.

3. TWO-ELEMENT PARASITIC ARRAY (DIRECTOR TYPE)

Our modeling tools teach us that we can indeed obtain exactly the same results with a parasitic array. A 2-element director-type array produces the same gain and a front-to-back ratio that is even slightly superior.

As a practical 2-element parasitic-type wire array, I have developed a Yagi with two inverted-V-dipole elements. **Fig 12-6** shows the configuration as well as the radiation patterns obtained at a height of 25 meters (0.3 λ on 80 meters). To make the array easily switchable, both wire elements are made equally long (39.94 meters for a design frequency of 3.8 MHz). The inverted-V-dipole apex angle is 90°. A 25 meter high mast or tower is required. At that height we need to install a 10 meter long horizontal support boom, from the end of which we can hang the inverted-V dipoles. The gain is 3.9 dB versus an inverted-V dipole at the same height, measured at the main elevation angle of 45°.

A loading capacitor with a reactance of $-j\,65\ \Omega$ produces the right current phase in the director. The radiation resistance of the array is 24 Ω. To make the array easily switchable, we run two feed lines of equal length to the elements. From here

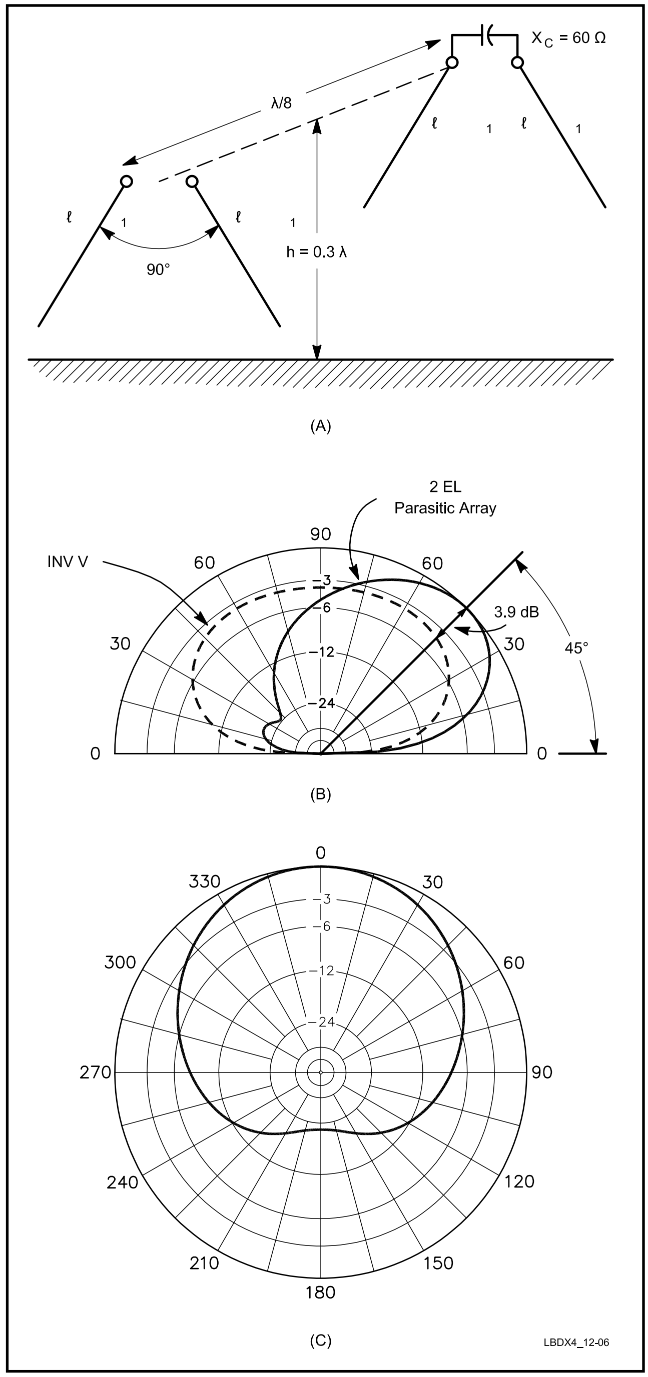


Fig 12-6 — Configuration and calculated radiation patterns for the 2-element parasitic array using inverted-V dipole elements. The array is installed with an apex angle of 90°, at a height of 0.3 λ (25 meters for 3.8 MHz). Element spacing is λ/8. The vertical pattern of a single inverted-V dipole is included at B for comparison. At C, the azimuth pattern is shown for an elevation angle of 45°. The gain at the 45° peak elevation angle is 3.9 dB over the single inverted-V dipole.

on there are two possibilities:

- We use a length of coax feed line to provide the required reactance of $-j\,65\ \Omega$ at the element.
- We use a variable capacitor at the end of a λ/2 feed line. The theoretical value of the capacitor is:

$$\frac{10^6}{2\pi \times 3.8 \times 65} = 644 \text{ pF}$$

Now we calculate the length of the open feed line that exhibits a capacitance of 644 pF on 3.8 MHz. The reactance at the end of an open feed line is given by:

$$X = Z_C \times \tan(90 - \ell) \qquad \text{(Eq 12-1)}$$

where:

Z_C = characteristic impedance of the line
ℓ = length of the line in degrees

This can be rewritten as

$$\ell = 90 - \arctan\frac{X}{Z_C} \qquad \text{(Eq 12-2)}$$

In our case we need X = – 60 Ω. Thus,

$$\ell = 90 - \arctan\frac{60}{50} = 39.8°$$

The physical length of this line is given by:

$$L_{meters} = \frac{833 \times Vf \times \ell}{1000 \times F_q} \qquad \text{(Eq 12-3)}$$

where

Vf = velocity factor (0.66 for RG-213)z
F_q = design frequency in MHz
ℓ = length in degrees

$$L_{meters} = \frac{833 \times 0.66 \times 39.8}{1000 \times 3.8} = 5.76 \text{ meters}$$

Fig 12-7 shows the feed and switching arrangements according to the two above-mentioned systems.

4. TWO-ELEMENT DELTA-LOOP ARRAY (REFLECTOR TYPE)

We can also design a 2-element delta-loop configuration using a somewhat shorter boom to separate the elements at the apex. If the ground conductivity is excellent, and if we can install radials (a ground screen), the 2-element delta-loop array should provide a lower angle of radiation and comparable gain to the 2-element inverted-V-dipole array described in Section 3.

4.1. Two-Element Delta Loop with Sloping Elements

Since the low-impedance feed point of the vertically polarized delta loop is quite a distance from the apex, and as most of the radiation comes from the high-current areas of the antenna, we can consider using delta-loop elements that are sloping away from the tower. We could not do this with the inverted-V, 2-element array, since the high-current points are right at the apex.

In this example I have provided a boom 6 meters in length at the top of the support at 25 meters. From the tips of the boom we slope the two triangles so that the base lines are now 8 meters away from the support and approximately 2.5 meters above the ground.

Fig 12-8 shows the radiation pattern obtained with the array when the loops are fed with equal current magnitude and with a phase difference of 120°. Note the tremendous F/B at low angles (more than 45 dB!). Gain over a single-element loop is 3.5 dB. The wave angle is 18° over a very good ground. One

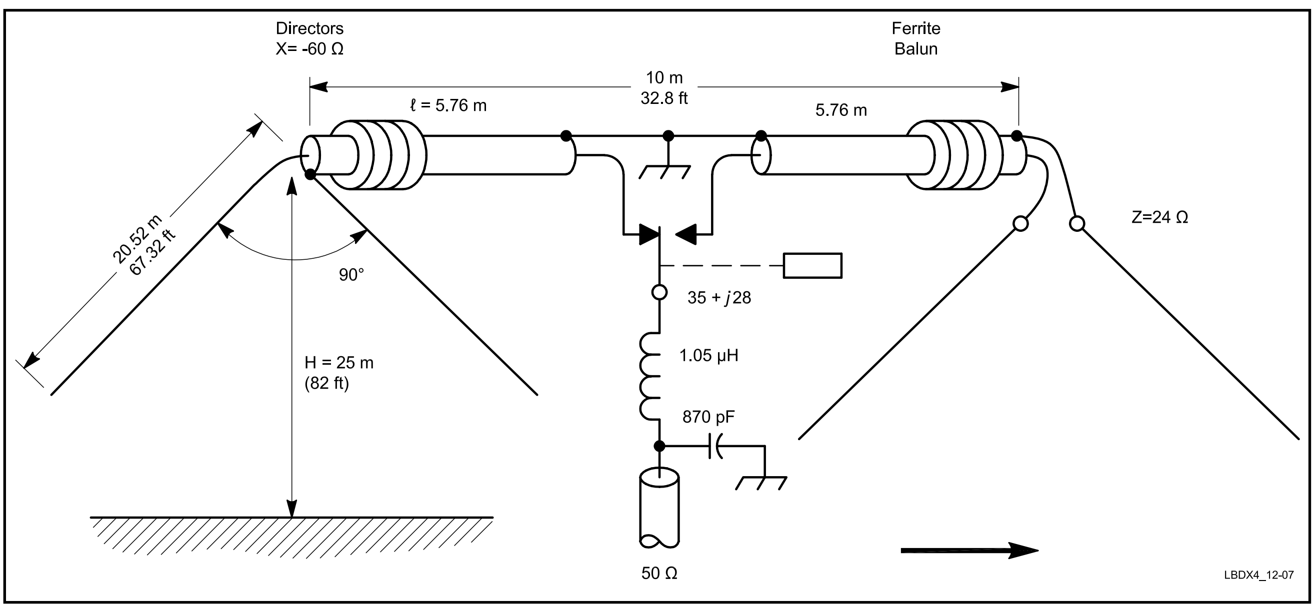


Fig 12-7 — Feeding arrangement for the 2-element parasitic array shown inFig 12-6. Two lengths of RG-213 run to a switch box in the center of the array. The coax feeding the director is left open at the end, producing a reactance of – *j* 65 Ω (equivalent to 644 pF at 3.8 MHz) at the element feed point. The radiation resistance of the 2-element array is 29 Ω. An L network can be provided to obtain a perfect match to the 50 Ω feed line. A current type of balun (eg, stack of ferrite beads) must be provided at both element feed points.

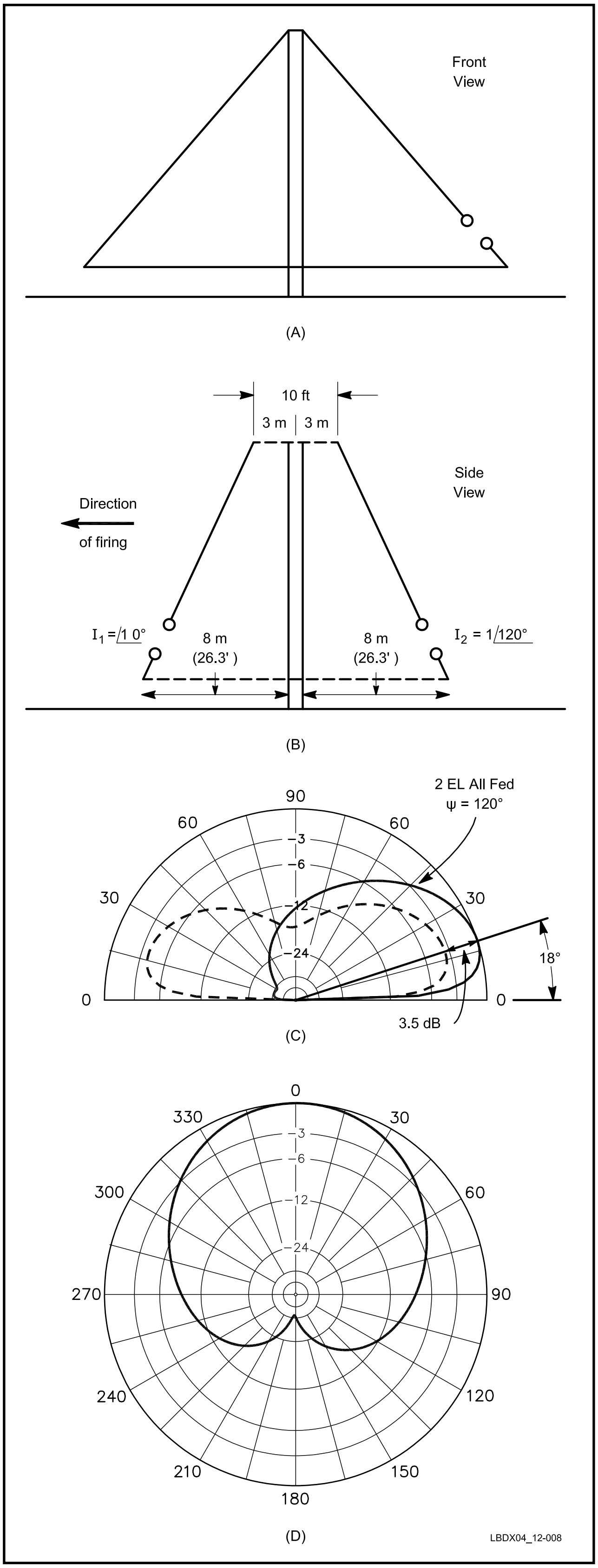


Fig 12-8 — Configuration and radiation patterns of a 2-element delta-loop array, using sloping elements. The elements are fed with equal-magnitude currents and with a phase difference of 120°. The horizontal pattern at D is for an elevation angle of 18°. The apex height for the loops is 25 meters.

of the problems is, of course, the feed system for an array that is not fed in quadrature.

Fig 12-9 shows the radiation patterns for the 2-element array with a parasitic reflector. The gain is the same as for the all-fed array and 3.4 dB over a single delta-loop element. The parasitic array shows a little less F/B at low angles, as compared to the all-fed array (see Fig 12-8), but the difference is slight.

As with the 2-element dipole array, my personal preference goes to the parasitic array, since the all-fed array is not fed in quadrature, which means that the feed arrangement is all but simple (it requires a modified Lewallen feed system). The obvious feed method for the 2-element parasitic array uses two equal-length feed lines to a common point mid-way between the two loops. A small support can house the switching and matching hardware.

As with the 2-element inverted-V array, we use two loops of identical length, and use a length of shorted feed line to provide the required inductive loading with the reflector element. The length of the feed line required to achieve the required 140 Ω inductive reactance is calculated as follows:

$$X_L = Z_C \times \tan \ell \qquad \text{(Eq 12-4)}$$

where

X_L = required inductance
Z_C = cable impedance
ℓ = cable length in degrees
This can be rewritten as

$$\ell = \arctan \frac{X_L}{X_C} \qquad \text{(Eq 12-5)}$$

or

$$\ell = \arctan \frac{140}{75} = 61.8°$$

The physical length is given by

$$L_m = \frac{833 \times VF \times \ell}{1000 \times F_q} \qquad \text{(Eq 12-6)}$$

where

L_m = length, meters
ℓ = length in degrees
VF = velocity factor of the cable
F_q = design frequency, MHz

We use foam-type RG-11 (Vf = 0.81), because solid PE-type coax (VF = 0.66) will be too short to reach the switch box.

$$L_{meters} = \frac{833 \times 0.81 \times 61.8}{1000 \times 3.8} = 10.97 \text{ meters}$$

Fig 12-10 shows the feed line and the switching arrangement for the array. Note that the cable going to the reflector must be short-circuited. The two coaxial feed lines must be equipped with current-type baluns (a stack of ferrite beads).

The impedance of the array varies between 75 Ω and 150 Ω, depending on the ground quality. If necessary, the impedance can easily be matched to the 50-Ω feed line using a small L-network. This array can be made switchable from the SSB end of the band to the CW end by applying the capacitive loading technique as described in Chapter 10.

Since this array was published in the Second Edition of this book, I have received numerous comments from people who have successfully constructed it.

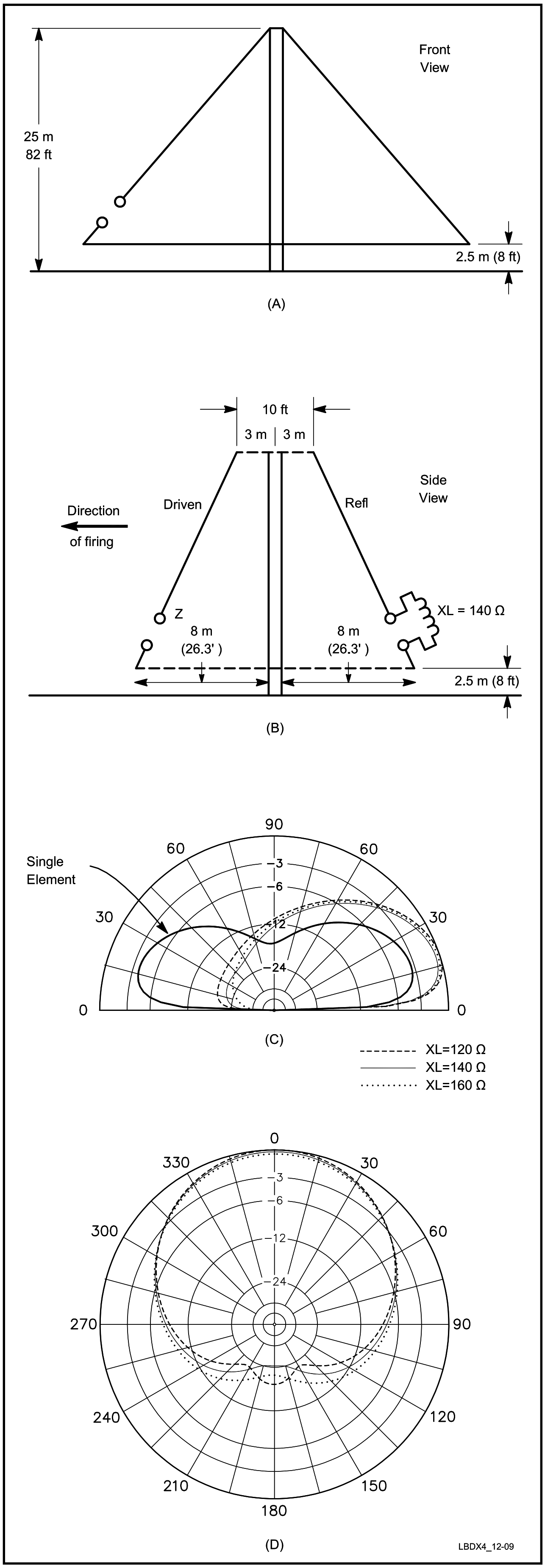


Fig 12-9 — Radiation patterns for the 2-element delta-loop array having the same physical dimensions as the all-fed array of Fig 12-8, but with one element tuned as a reflector. In practice both triangles are made equal size, and the required loading inductance is inserted to achieve the phase angle. Patterns shown are for different values of loading coils (X_L = 120, 140 and 160 Ω). The feed-point impedance of the array will vary between 80 Ω and 150 Ω, depending on the ground quality.

5. THREE-ELEMENT DIPOLE ARRAY WITH ALL-FED ELEMENTS

A 3-element phased array made of λ/2 dipoles can be dimensioned to achieve a very good gain together with an outstanding F/B ratio. Three elements on a λ/4 boom (giving λ/8 spacing between elements) can yield nearly 6 dB of gain at the major radiation angle of 38° over a single dipole at the same height (over average ground).

A. Christman, KB8I, described a 3-element dipole array with outstanding directional and gain properties. (Ref 963.) I have modeled a 3-element inverted-V-dipole array using the same phase angles. The inverted-V elements have an apex angle of 90º, and the apex at 25 meters above ground. The radiation patterns are shown in **Fig 12-11**.

The elements are fed with the following currents:

I1 = 1 $\angle -149°$ A
I2 = 1 $\angle 0°$ A
I3 = 1 $\angle 146°$ A

With the antenna at 25 meters above ground and elements that are 39.72 meters long (design frequency = 3.8 MHz), the element feed-point impedances are:

Z1 = $-36 + j\,24.5$ Ω
Z2 = $12.3 + j\,25$ Ω
Z3 = $7.6 - j\,12.2$ Ω

If you are confused by the minus sign in front of the real part of Z1, it just means that in this array, element number 1 is actually *delivering* power into the feed system, rather than taking power from it. This is a very common situation with driven arrays, especially where close spacing is used.

A possible feed method consists of running three λ/4 lines to a common point. Current forcing is employed: We use 50-Ω feed lines to the outer elements, and two parallel 50-Ω lines to the central element. The method is described in detail in Chapter 11 on vertical arrays.

It is much easier to model such a wonderful array and to calculate a matching network than to build and align the matching system. Slight deviations from the calculated impedance values mean that the network component values will be different as well. There is no method of measuring the driven impedances of the elements. All you can do in the way of measuring is use an HF vector voltmeter and measure the voltages at the end of the three feed lines. The voltage magnitudes should be identical, and the phase as indicated above (E1, E2 and E3). If they are not, the values of the networks can be tweaked to obtain the required phase angles. Good luck!

We have seen that we can just about match the performance of a 2-element all-fed array with a parasitic array. We will see that the same can be done with a 3-element array.

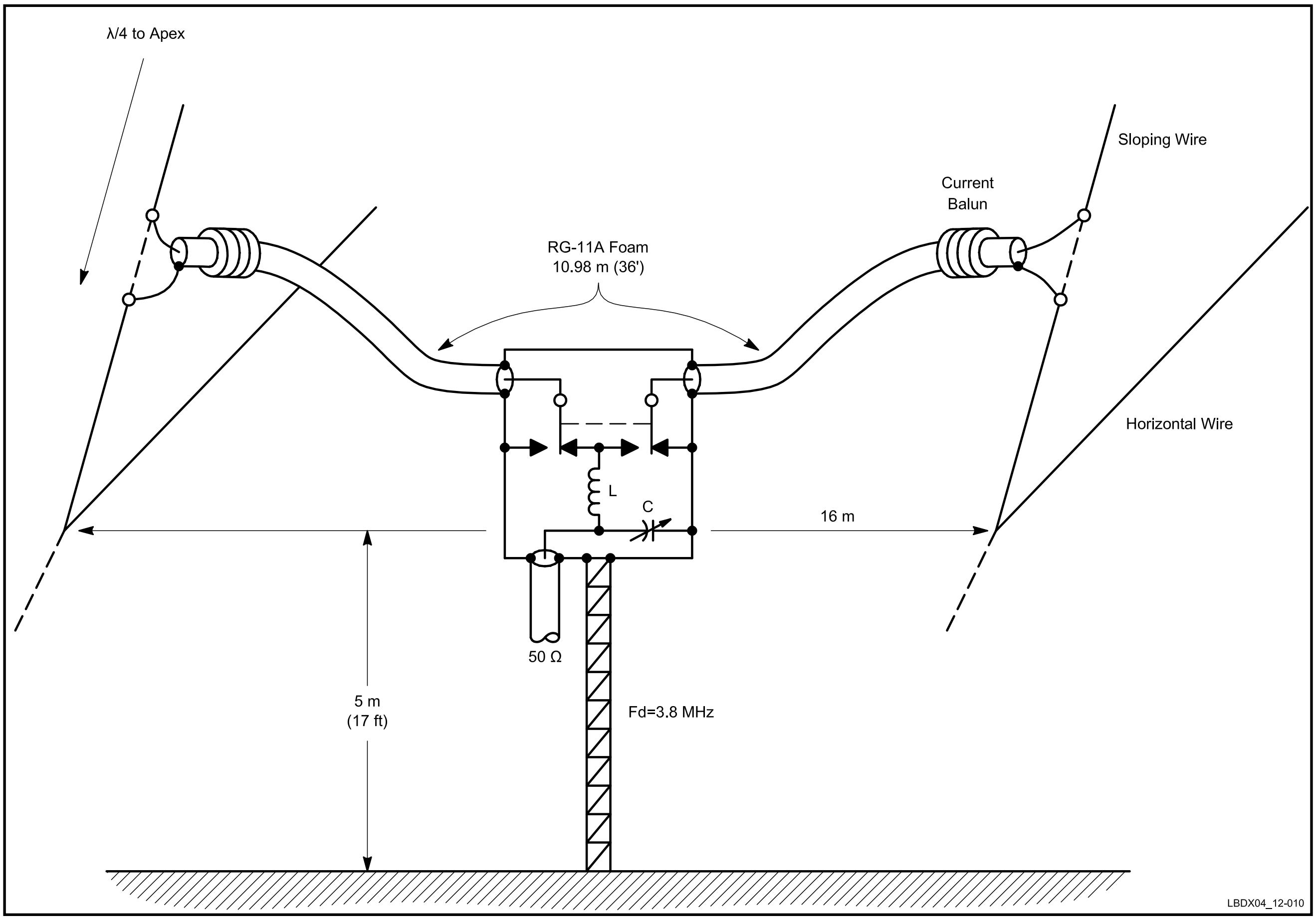


Fig 12-10 — Feeding and direction-switching arrangement for the 2-element parasitic delta-loop array as shown in Fig 12-9. The length of the 75 Ω feed lines going from the feed points to the switch box is 61.8°. For 3.8 MHz, and using foam-type coax (Vf = 0.81), this equals 10.98 meters. The spacing between the elements at the height of the feed points is about 15 meters. Note that the feed line to the reflector needs to be short-circuited. A simple L-network provides a perfect match for a 50 Ω feed line.

6. THREE-ELEMENT PARASITIC DIPOLE ARRAY

The model that was developed has a gain of 4.5 dB over a single inverted V-element (at the same height) for its main elevation angle of 43°. The F/B ratio is just over 20 dB, as compared to just over 30 dB with the all-driven array. At the same antenna height (0.3 λ), the radiation angle of the 3-element parasitic was also slightly higher (43 Ω) than for the 3-element all-fed array (38 Ω), modeled over the same average ground.

Fig 12-11 shows the superimposed patterns for the all-driven and the parasitic 3-element array (for 80 meters at 25 meters height). Note that the 3-element all-fed has a better rejection at high angles. This is because the currents in the outer elements have a greater phase shift (versus the driven element) than in the parasitic array. These phase shifts are:

Reflector:
All-driven array: –149°
Parasitic array: –147°

Director:
All-driven array: +147°
Parasitic array: +105°

This demonstrates again that, with an all-driven array, we have more control over all the parameters that determine the radiation pattern of the array. Like the 2-element array described in Section 3, the 3-element array is also made using three elements identical in length. The required element reactances for the director and reflector are obtained by inserting the required inductance or capacitance in the center of the element. In practice we bring a feed line to the outer elements as well. The feed lines are used as stubs, which represent the required loading to turn the elements into a reflector or director.

The question is, which is the most appropriate type of feed line for the job, and what should be its impedance? **Table 12-1** shows the stub lengths obtained with various types of feed lines. The length of the open-ended stub serving to produce a negative reactance (for use as a director stub) is given by:

$$\ell^\circ = 90 - \arctan\frac{X_C}{Z_C} \qquad \text{(Eq 12-7)}$$

For the short-circuited stub serving to produce a positive reactance (for the reflector), the formula is:

$$\ell^\circ = \arctan\frac{X_L}{Z_L}$$

- From Table 12-1 we learn the 450-Ω stub requires a very long length to produce the required negative reactance for the director (17.28 meters).
- When made from 50-Ω or 75-Ω coax, we obtain attractive short lengths. The disadvantage is that you need to put a current balun at the end of the stubs to keep any current from flowing on the outside of the coax shield.
- A third solution is to use a 100-Ω shielded balanced line, made of two 50-Ω coax cables. The lengths are still very attractive, and you no longer require the current balun.
- A final solution is to use the 450-Ω transmission line for the reflector (1.71 meters long) and to load the line with an extra capacitor to turn it into a capacitor. I assumed a velocity factor of 0.95 for the transmission line. You must check this in all cases (see Chapter 11 on vertical arrays). The capacitive reactance produced by an open-circuited line of 1.71 meters length at 3.8 MHz is:

$$X_L = 450 \tan(90° - 8.22°) = +j\,3115\,\Omega$$

This represents a capacitance value of only:

$$\frac{10^6}{2\pi \times 3.8 \times 3115} = 13.4 \text{ pF}$$

The required capacitive reactance was $-j$ 55 Ω, which represents a capacitance value of

$$\frac{10^6}{2\pi \times 3.8 \times 55} = 762 \text{ pF}$$

Table 12-1
Required Line Length for the Loading Stubs of the Parasitic Version of the 3-Element Array of Fig 12-11

Z_C (Ω)	*VF*	*Length (Degrees)*	*Length (Meters)*	*Length (Feet)*
Director				
50	0.66	42.3	6.12	20.08
75	0.66	53.75	7.77	25.49
100	0.95	83.03	8.85	29.04
450	0.95	83.03	17.28	56.70
Reflector				
50	0.66	52.53	7.58	13.39
75	0.66	40.91	5.91	13.39
100	0.66	33.02	4.77	15.65
450	0.95	8.22	1.71	5.61

Other data:
Design frequency = 3.8 MHz, wavelength = 78.89 meters
Director X_C = –55 Ω
Reflector X_L = +65 Ω

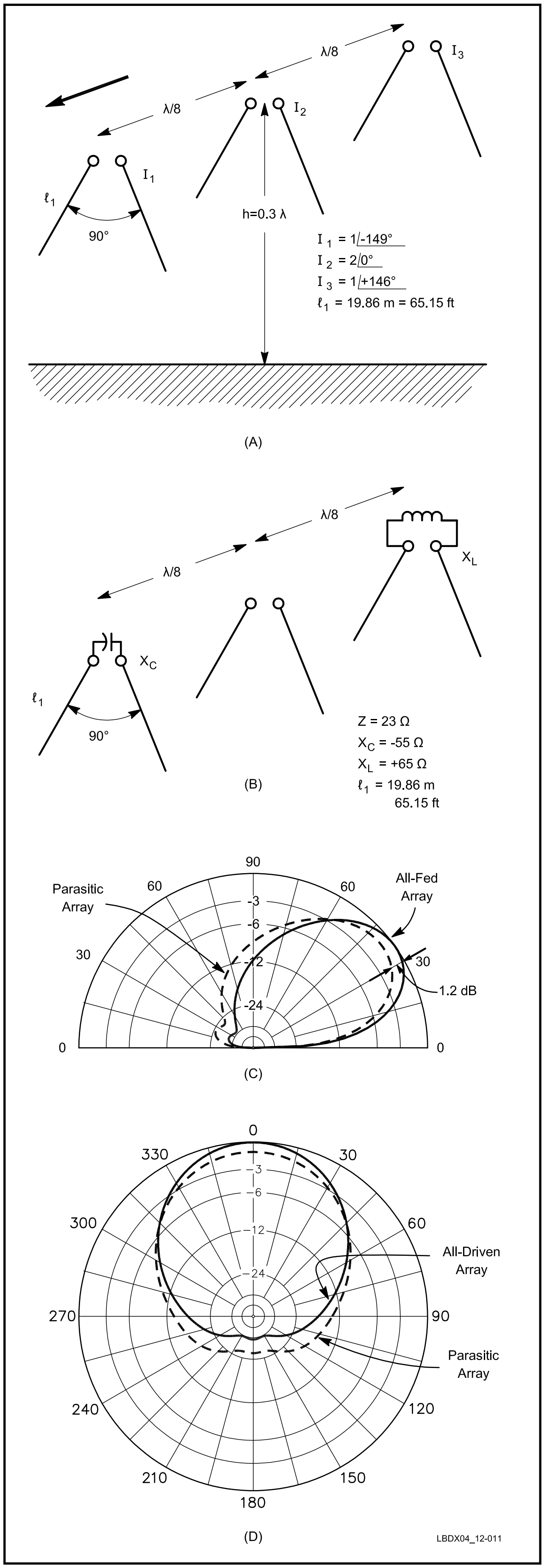


Fig 12-11 — Configuration and radiation patterns for two types of 3-element inverted-V-dipole arrays with apexes at 0.3 λ. At both C and D, one pattern is for the all-fed array and the other for an array with a parasitic reflector and director. The all-fed array outperforms the Yagi-type array by approximately 1 dB in gain, as well as 10 dB in F/B.

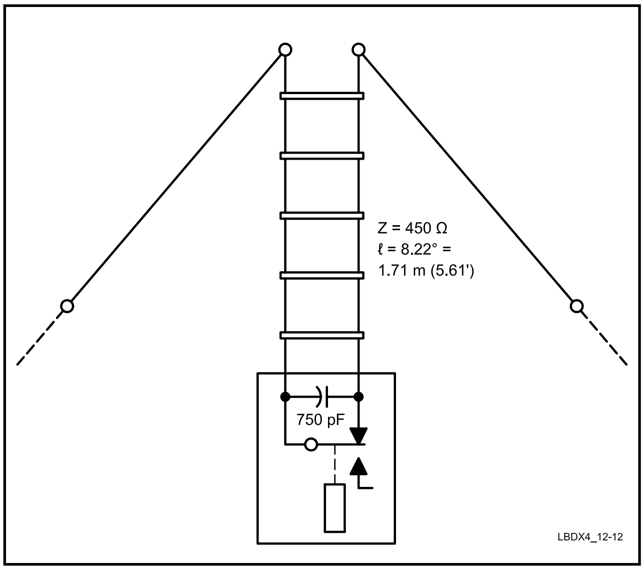


Fig 12-12 — The 3-element parasitic type inverted-V dipole array is made with elements that have exactly the same length. The required element loading is obtained by inserting the required capacitance or inductance in the center of these elements. This is obtained by using stubs, as shown here. With a 450-Ω transmission line we require only a short 1.71 meter long piece of short-circuited line to make a stub for the reflector. For the director we connect a 750-pF capacitor across the end of the open-circuit line. This can be switched with a single-pole relay, as explained in the text.

This means we need to connect a capacitor with a value of 762 – 13.4 = 750 pF across the end of the open stub. This last solution seems to be the most flexible one. A parallel connection of two transmitting-type ceramic capacitors, 500 pF and 250 pF, will do the job perfectly. If you want even more flexibility you can use a 500-pF motor-driven variable in parallel with a 500-pF fixed capacitor. This will allow you to tune the array for best F/B.

The practical arrangement is shown in **Fig 12-12**. From each outer element we run a 1.71-meter long piece of 450-Ω line to a small box mounted on the boom. The box can also be mounted right at the center of the inverted-V element, whereby the 1.71 meter transmission line is shaped in a large 1-turn loop. The box houses a small relay, which either shorts the stub (reflector) or opens, leaving the 750-pF capacitor across the line.

Is the relative "inferiority" of the parasitic array due to the low height? In order to find out I modeled the same antennas at λ/2 height. **Fig 12-13** shows the vertical and the horizontal radiation patterns for the all-driven and parasitic-array versions of the 3-element inverted-V array at this height. Note that the all-driven array still has 0.9 dB better gain than the parasitic array. The F/B is still a little better as well, although the difference is less pronounced than at lower height. The optimum pattern was obtained when loading the director with a –50-Ω impedance and the reflector with a +30-Ω impedance. The gain of the all-fed array is 5.7 dB versus a dipole at the same height (at 28° elevation angle). For the 3-element parasitic array, the gain is 4.8 dB versus the dipole at its main elevation angle of 29°.

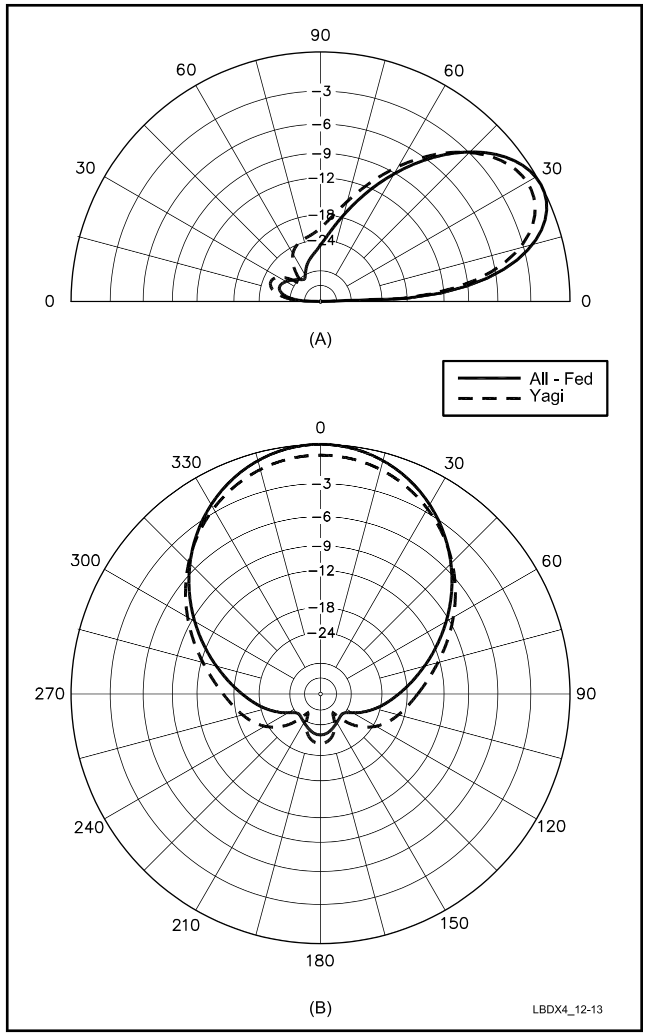


Fig 12-13 — Radiation pattern of the 3-element inverted-V type array at a height of λ/2. Note that the all-fed array still outperforms the Yagi-type array, but with a smaller margin than at a height of 0.3 λ (Fig 12-11). To produce an optimum radiation pattern, the values of the loading impedances were different than those for a height of 0.3 λ. See text for details.

In looking at the vertical radiation pattern it is remarkable again that the all-driven array excels in F/B performance at high angles. Notice the "bulge" that is responsible for 5 to 10 dB less F/B in the 35° to 50° wave-angle region.

It must be said that I did not try to further optimize the parasitic array by shifting the relative position of the elements. By doing this, further improvement could no doubt be made. This, of course, would make it impossible to switch directions, since the array would no longer be symmetrical.

6.1. Conclusion

All-fed arrays made of horizontal dipoles or inverted-V dipoles always outperform the parasitic-type equivalents in gain as well as F/B performance. As they are not fed in quadrature, it is elaborate or even "difficult" to feed them correctly.

The parasitic-type arrays lend themselves very well for remote tuning of the parasitic elements. Short stubs (open-ended to make a capacitor, or short-circuited to make an inductor) make good tuning systems for the parasitic elements. Switching from director to reflector can easily be done with a single-pole relay and a capacitor at the end of a short open-wire stub.

The same 3-element array made of fully horizontal (flat top) dipoles exhibits 1.0 dB more gain than the inverted-V version at the same apex height.

7. DELTA LOOPS IN PHASE (COLLINEAR)

Two delta loops can be erected in the same plane and fed with in-phase currents to provide gain and directivity. In order to obtain maximum gain, the loops must be separated about λ/8, as shown in **Fig 12-14**. In this case the two loops, fed in phase exhibit a gain of almost 3.5 dB over a single loop! The array has a front-to-side directivity of at least 15 dB, not negligible. The impedance on a single loop is between 125 and 160 Ω. Each element can be fed via a 75 Ω λ/2 feed line. At the point where they join the impedance will be 60 to 80 Ω. The radiation patterns and the configuration are shown in Fig 12-14.

This may be an interesting array if you have two towers with the right separation and pointing in the right direction. As with all vertically polarized delta loops, the ground quality is very important as to the efficiency and the low-angle radiation of the array (see Chapter 10 on large loops).

Putting the loops closer together results in a spectacular drop in gain. Loops with touching tips only exhibit approximately 1-dB gain over a single element—they're not worth the effort!

In one of his articles on elevated radials, John Belrose, VE2CV, mentioned the half-diamond loop, which has a significant resemblance to the delta loop (Ref 7824). I modeled this array and compared it to the 2-element delta loop shown in Fig 12-14. **Fig 12-15** shows both the horizontal and the vertical radiation pattern of both antennas in overlay. The 2-element delta has almost 0.7 dB more gain and has excellent high-angle rejection, while the half-diamond loop has some very strong high-angle response, which is of course due to the way the radials are laid out, resulting in zero high-angle cancellation. The extra gain that was thought to be achieved by laying radials in one direction, is apparently more than wasted in high angle radiation. It seems that the two in-phase delta loops are still, by far, the best choice.

8. ZL SPECIAL

The ZL Special, sometimes called the HB9CV, is a 2-element dipole array with the elements fed 135° out-of-phase. This configuration is described in Section 2. It is the equivalent of the vertical arrays described in Chapter 11.

These well-known configurations make use of a specific feeding method. The feed points of the two elements are connected via an open-wire feed line that is crossed. The crossing introduces a 180° phase shift. The length of the line, with a spacing of λ/8 between the elements, introduces an additional phase shift of approximately 45°. The net result is 180° + 45° = 225° phase shift, lagging. This is equivalent to 360° – 225° = 135° leading.

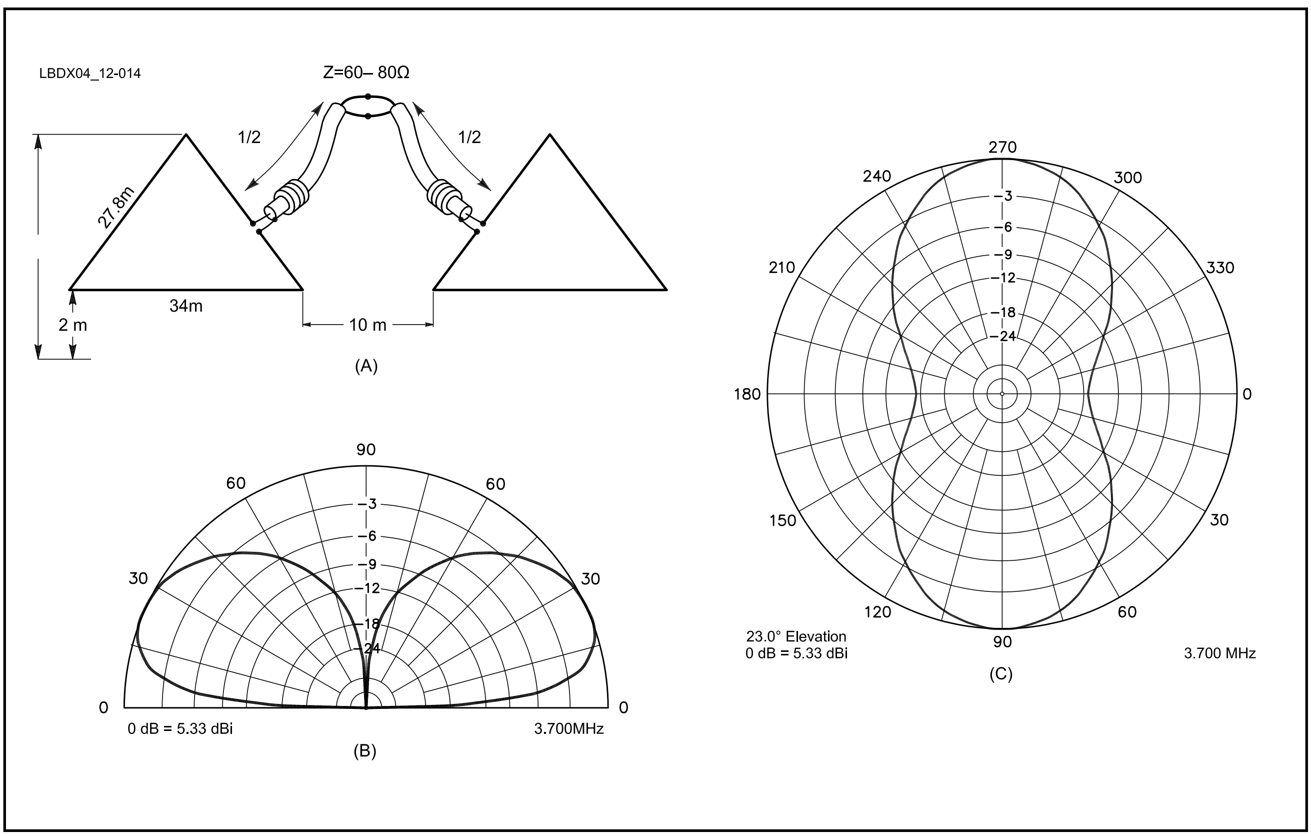
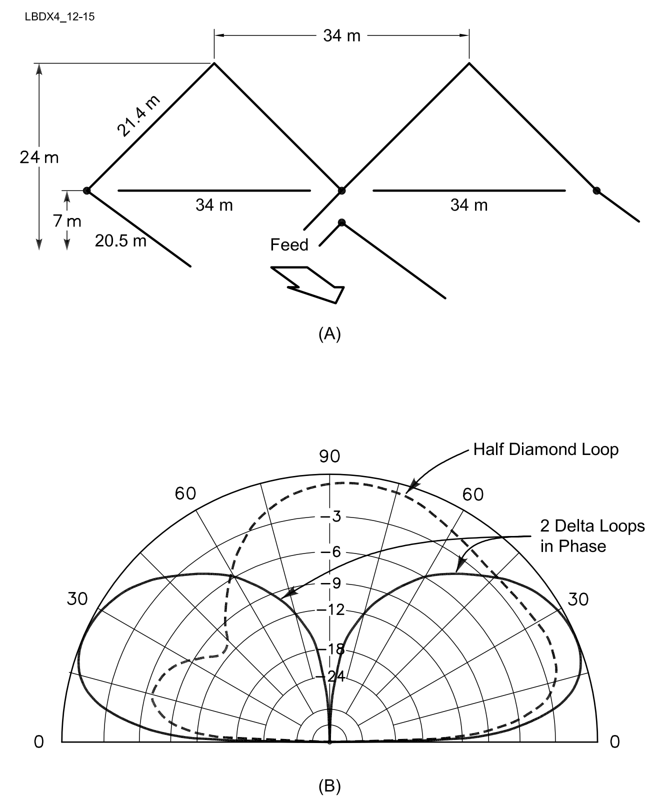


Fig 12-14 — Configuration of the 2-element collinear delta-loop array with 10 meter spacing between the tips of the deltas. This array has a gain of 3.0 dB over a single delta loop. The loops are fed λ/4 from the apex on the sloping wire in the center of the array (see text for details). The pattern at C is for an elevation angle of 18°.



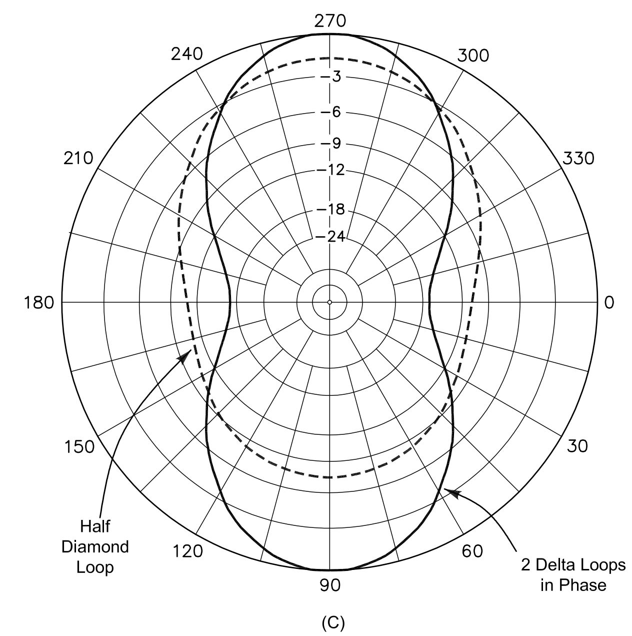
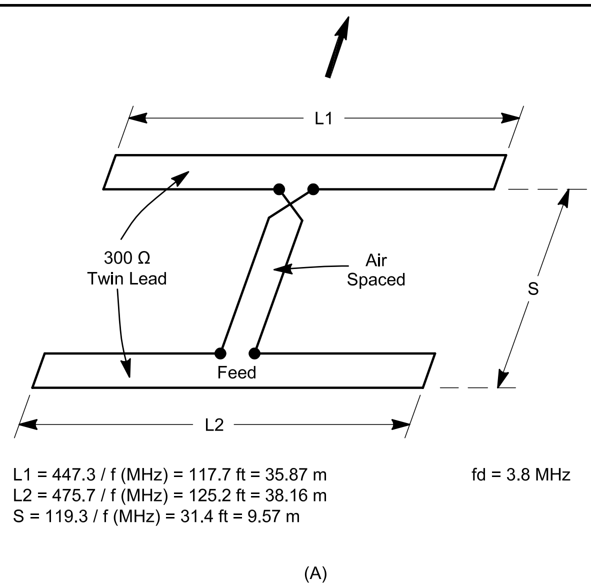
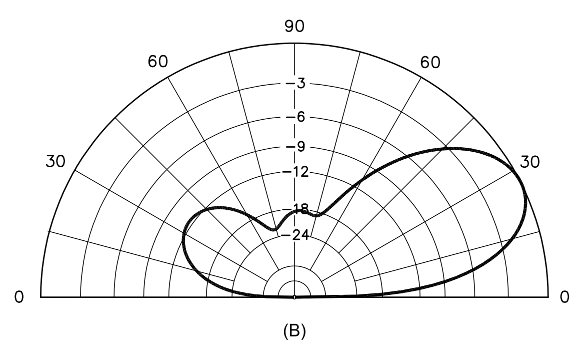


Fig 12-15 — There is some similarity between the half-diamond loop, described by VE2CV and shown at A, and two delta loops in phase. Overlays of the vertical (B) and horizontal (C) patterns show, however, that the 2-element delta loop has better high-angle discrimination, in addition to almost 1 dB more gain.





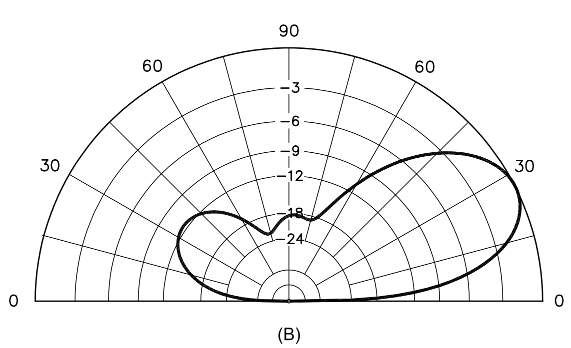


Fig 12-16 — The ZL Special (or HB9CV) antenna is a popular design that gives good gain and F/B for close spacing. Radiation patterns were calculated with *ELNEC* for the dimensions shown at A, for a height of λ/2 above average ground. The horizontal pattern at C is for an elevation angle of 27°.

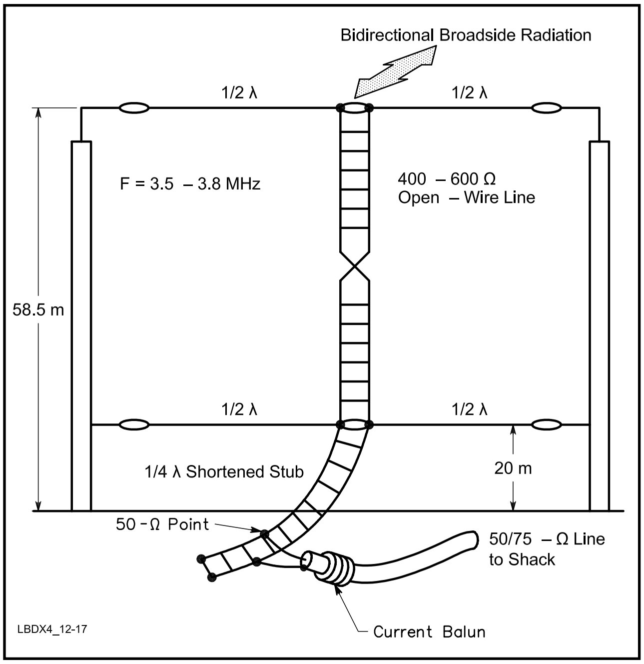


Fig 12-17 — Typical Lazy-H configuration for 80 meters. The same array can obviously be made for 40 meters with all dimensions halved.

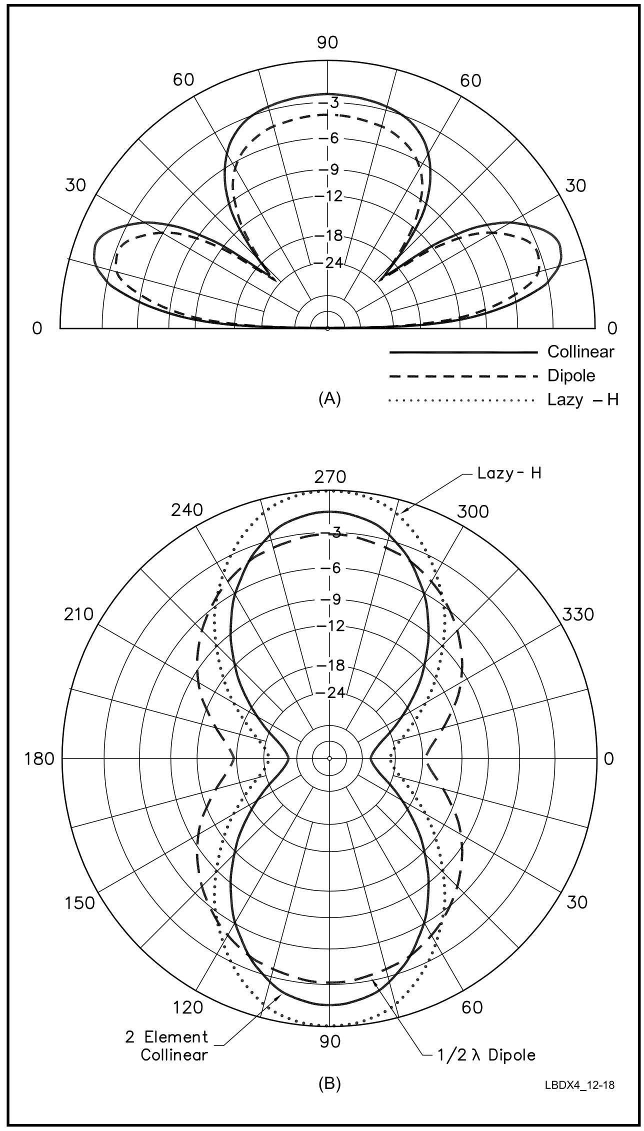


Fig 12-18 — Vertical and horizontal radiation patterns of the 80-meter Lazy H shown in Fig 12-17, compared to the patterns of a flat top dipole and a 2 × λ/2 collinear at the same height (over average ground).

Different dimensions for this array have been printed in various publications. Correct dimensions for optimum performance will depend on the material used for the elements and the phasing lines. Jordan, WA6TKT, who designed the ZL Special entirely with 300-Ω twinlead (Ref 908), recommends that the director (driven element) be 447.3/f_{MHz} and the reflector be 475.7/f_{MHz}, with an element spacing of approximately 0.12 λ.

Using air-spaced phasing line with a velocity factor of 0.97, the phasing-line length is 119.3°. This configuration of the ZL Special with practical dimensions for a design frequency of 3.8 MHz is given in **Fig 12-16**, along with radiation patterns. As it is rather unlikely that this antenna will be made rotatable on the low bands, I recommend the use of open-wire feeders to an antenna tuner. Alternatively, a coaxial feed line can be used via a balun.

9. LAZY H

The Lazy-H antenna is an array that is often used by low-banders that have a bunch of tall towers, where they can support Lazy-Hs between them. **Fig 12-17** shows a typical Lazy-H layout for use on 80 meters. Such a 4-element Lazy-H has a very respectable gain of about 11 dBi over average ground, as shown in **Fig 12-18**. Its gain at a 20° elevation angle is nearly 4 dB above a flat-top dipole at the same height, and 1.7 dB over a collinear (two λ/2 waves) at the same height. The outstanding feature of the Lazy-H is however, that the 90° (zenith) radiation, which is very dominant with the dipole and the collinear, is almost totally suppressed. This makes it a good DX listening antenna as well!

The easiest way to feed the array is shown in Fig 12-18. A λ/4 open-wire line, shorted at its end, is probed to find the low-impedance point (50 or 75 Ω). Fine adjustment of the length of the line and the position of the tap make it possible to find a perfect resistive 50 or 75-Ω point. One of the popular antenna analyzers is a valuable tool to find the exact match. The same antenna can be used for both ends of the 80 meter band, all that is required is a different set of values for the length of the λ/4 stub and the position of the tap. This can be achieved with some rather simple relay switching.

10. BOBTAIL CURTAIN

The Bobtail Curtain consists of three phased λ/4 verticals, spaced λ/2 apart, where the center element is fed at the base, while the outer elements are fed via a horizontal wire section between the tips of the verticals. Through this feeding arrangement, the current magnitude in the outer verticals is half of the current in the center vertical. The current distribution in the top wire is such that all radiation from this horizontal section is effectively canceled. The configuration as well as the radiation patterns are shown in **Fig 12-19**.

The gain of this array over a single vertical is 4.4 dB.

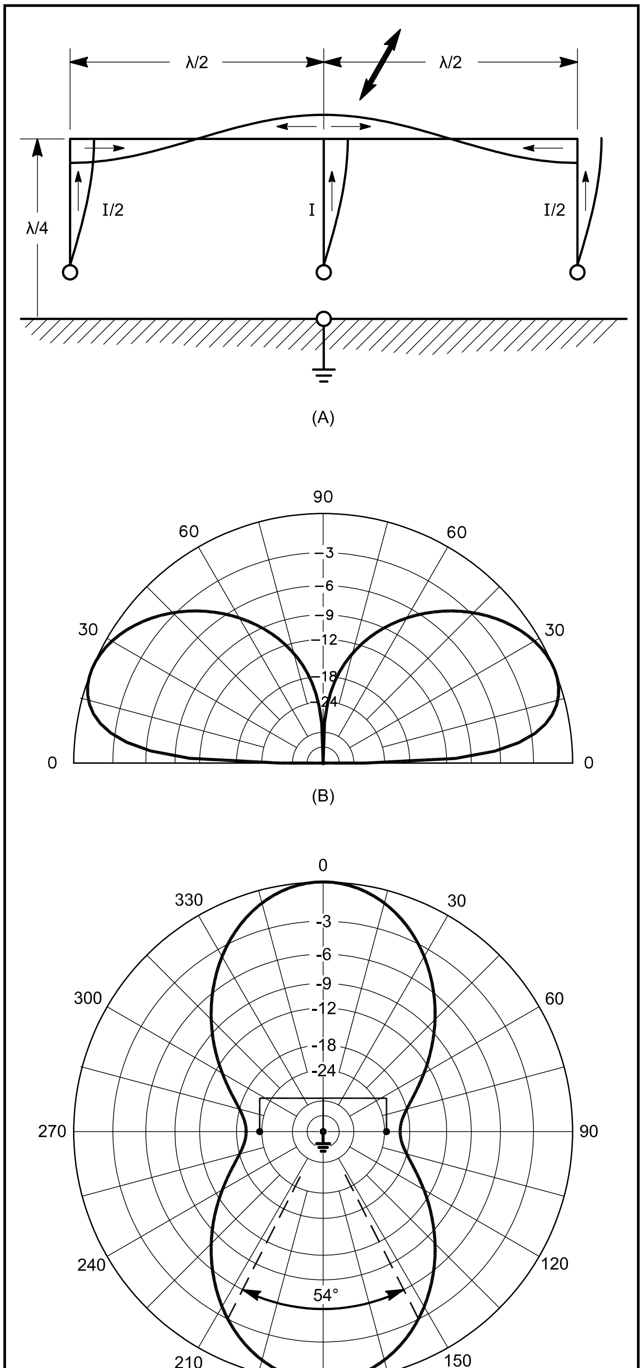


Fig 12-19 — Configuration and radiation patterns for the Bobtail Curtain. This antenna exhibits a gain of 4.4 dB over a single vertical element. The current distribution, shown at A, reveals how the three vertical elements contribute to the low-angle broadside bidirectional radiation of the array. The horizontal section acts as a phasing and feed line and has no influence on the broadside radiation of the array. The horizontal pattern at C is for an elevation angle of 22°.

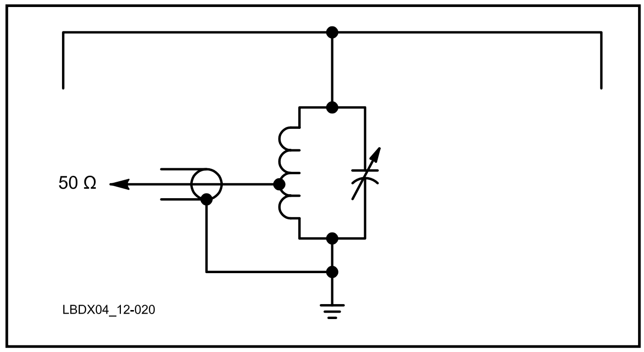


Fig 12-20 — The Bobtail Curtain is fed at a high-impedance point with a parallel-tuned circuit, where the coax is tapped a few turns from the cold end of the coil. The array can be made to operate over a very large bandwidth by simply retuning the tuned circuit.

The –3-dB forward-lobe beamwidth is only 54°, which is quite narrow. This is because the radiation is bidirectional. K. Svensson, SM4CAN, who published an interesting little booklet on the Bobtail Array, recommends the following formulas for calculating the lengths of the elements of the array.

Vertical radiators: $\ell = 68.63/F_{MHz}$

Horizontal wire: $\ell = 143.82/F_{MHz}$

where
F_{MHz} = design frequency, MHz
ℓ = length in meters

The antenna feed-point impedance is high (several thousand ohms). The array can be fed as shown in **Fig 12-20**. This is the same feed arrangement as for the voltage-fed T antenna, described in Chapter 9 on vertical antennas. In order to make the Bobtail antenna cover both the CW as well as the phone end of the band, it is sufficient to retune the parallel resonant circuit. This can be done by switching a little extra capacitor in parallel with the tuned circuit of the lower frequency, using a high-voltage relay.

The bottom ends of the three verticals are very hot with RF. You must take special precautions so that people and animals cannot touch the vertical conductors.

Do not be misled into thinking that the Bobtail Array does not require a good ground system just because it is a voltage-fed antenna. As with all vertically polarized antennas, it is the electrical quality of the reflecting ground that determines the efficiency and the low-angle radiation of the array.

11. HALF-SQUARE ANTENNA

As its name implies, the Half-Square is half of a Bi-Square antenna (on its side), with the ground making up the other half of the antenna (see Chapter 10 on large loop antennas). It can also be seen as a Bobtail with part of the antenna missing.

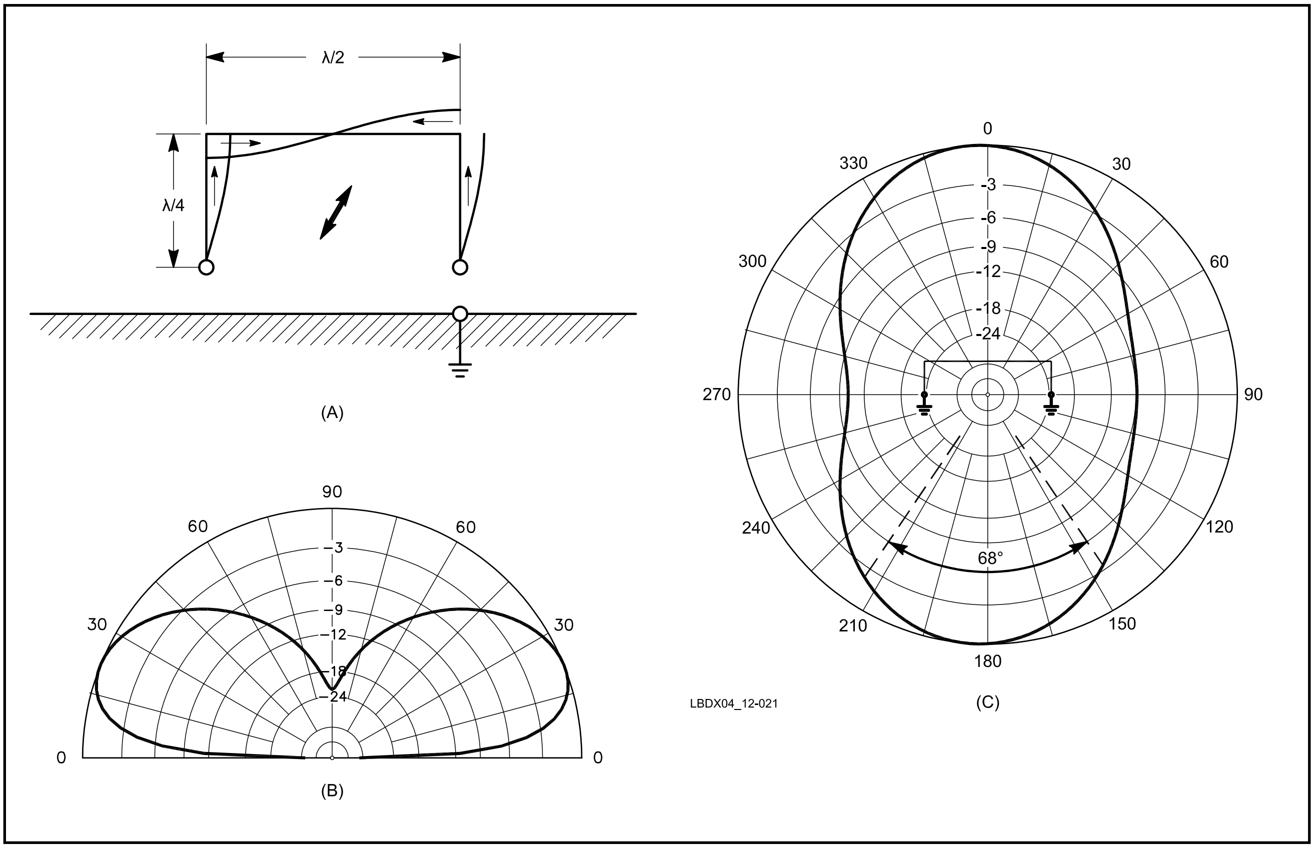


Fig 12-21 — Configuration and radiation patterns of the Half-Square array, with a gain of 3.4 dB over a single vertical. The antenna pattern is somewhat asymmetrical because the currents in the vertical conductors are not identical. The azimuth pattern at C is for an elevation angle of 22°.

Fig 12-21 shows the antenna configuration and the radiation patterns. The feed-point impedance is very high (several thousand ohms), and the antenna is fed like the Bobtail. The gain is somewhat less than 3.4 dB over a single λ/4 vertical. The forward-lobe beamwidth is 68°, and the pattern is essentially bidirectional. There is some asymmetry in the pattern, which is caused by the asymmetry of the design: The current flowing in the two verticals is not identical. As far as the required ground system is concerned, the same remarks apply as for the Bobtail antenna.

Contents

CHAPTER 13

Yagis and Quads

Tim Duffy, K3LR, needs no introduction to readers of this book. The way Tim runs his Dayton Antenna Forum and his own super contest station tells a lot about the man. He is thorough, well organized, punctual, and a super host on top of it all! No wonder there's a long line of operators who want to operate from K3LR in the big contests!

Tim is a telecommunications executive and has been employed in the broadcast and wireless engineering discipline for over 30 years. Tim is a graduate of the Pennsylvania State University and has been a licensed Amateur Radio operator since 1972. He currently maintains his large 9-tower, 12-operating position multi-multi station in Western Pennsylvania and experiments with large high-gain contest antennas.

Tim took the time to review this chapter on Yagis and Quads, for which I am very thankful.

Tim Duffy, K3LR, a well-known contester and superstation builder from Western Pennsylvania.

On the higher HF bands, almost all dedicated DXers use some type of rotatable directional antenna. Directional antennas produce gain to be better heard. They also show directivity, which is a help when listening. Yagi and cubical-quad antennas are certainly the most popular antennas on those bands.

On the low bands, rotatable directive antennas are large. Forty-meter Yagis and quads — even full-size — exist in greater numbers these days. On 80 meters there are only a few full-size Yagis and quads, while reduced-size Yagis and quads are a little more common. They seem to come and go, and are rather difficult to keep in the air. In the previous editions of this book I wrote, "On 160 meters, rotatable Yagis still belong to dreamland." This is no longer true (see Section 2.10).

I have had the chance to operate a 3-element full-size quad, as well as a 3-element full-size Yagi, on 80 meters, and I must admit that it is only when you have played with such monsters that you appreciate what you are missing without them. The same is even more true on 40 meters, where full-size Yagis and quads appear in ever-growing numbers. Until the day I had my own full-size 40-meter Yagi, I always considered 40 as my worst band. Now that I have the full-size Yagi, I think it has become my "best" band.

Much of the work presented in this chapter is the result of a number of major antenna projects that were realized with the help of Roger Vermet, ON6WU, who has been a most enthusiastic supporter and advocate in all my antenna work.

1. ARRAYS WITH PARASITIC ELEMENTS

In Chapter 11 on vertical arrays I discuss arrays of antennas, where each antenna element is fed with a dedicated feed line. During the analysis of these arrays I noticed that elements sometimes exhibit a negative impedance, which means that these elements do not draw power from the feed line but actually deliver power into the feed system.

In such a case mutual coupling has already supplied enough (or too much) current into the element. Negative feed-point impedances are typical with close-spaced arrays, where the coupling is heavy.

Parasitic arrays are arrays where (most often) only one element is fed, and where the other elements obtain their feed current only by mutual coupling with the various elements of the array. To obtain the desired radiation pattern and gain, feed-current magnitudes and phases need to be carefully adjusted.

This is done by changing the relative positions of the elements and by changing the lengths of the elements. The exact length of the *driven element* will not influence the pattern nor the gain of the array; it will only influence the feed-point impedance.

Unlike with driven arrays, you cannot obtain just any specific feed-current magnitude and angle. In driven arrays you "force" the antenna currents, which means you add (or subtract) feed current to the element current already obtained through mutual coupling. You could, for example, make a driven array with three elements in-line where all elements have an identical feed current. You cannot make a parasitic array where the three elements have the same current phase and magnitude. Arrays with parasitic elements are limited in terms of the current distribution in the elements.

In a Yagi or quad or other parasitic array the parasitic elements are adjusted (in length and position) to provide the required current with the lagging phase angle for the director, and with the required leading phase angle for the reflector.

2. QUADS VERSUS YAGIS

It is not my intention to get into the debate of quads versus Yagis. But before I tackle both in more depth, let me clarify a few points and kill a few myths:

- For a given height above ground, the quad does *not* produce a markedly lower radiation angle than the Yagi. The vertical radiation angle of a horizontally polarized antenna depends on the height of the antenna above ground.
- There is a very slight difference (perhaps a few degrees, depending on actual height) in favor of the quad, as there is some more squeezing of the vertical plane due to the effect of the stacked two horizontal elements that make a horizontally polarized cubical quad (Ref 980).
- For a given boom length, a quad will produce slightly more gain than a Yagi. This is logical since the aperture (capture area) is larger. The principle is simple: Everything being optimized, the antenna with the largest capture area has the highest gain, or can show the highest directivity.
- Yagis as a rule are easier to build and maintain. A Yagi is two-dimensional, and the problems involved with low-band antennas are simplified by an order of magnitude. Problems of wire breaking are nonexistent with Yagis. Large Yagis are also easier to handle and to install on a tower than large quads.
- There are other factors that will determine the eventual choice between a Yagi or a quad, such as material availability, maximum turning radius (the quad takes less rotating space) and, of course, personal preference.

3. YAGIS

There have been a number of good publications on Yagi antennas. Until about 20 years ago, before we all knew about the effect of tapered elements, the W6SAI/W2LX *Beam Antenna Handbook* was in many circles considered the Yagi "bible." Over 40 years ago I built my first Yagi based on information from this work.

Dr Jim Lawson, W2PV (SK), wrote a very good series on Yagis back in the early 1980s. Later the ARRL published his work in the excellent book, *Yagi Antenna Design* (Ref 957). Lawson explained how he scientifically designed a winning contest station, based on high-level engineering work.

Lawson was the first in amateur circles to methodically study the effect of tapered elements. He came up with a tapering algorithm that is still widely called the *W2PV algorithm*. It calculates the correct electrical length of an element as a function of the length and diameter of individual tapered sections.

3.1. Modeling Yagi Antennas

We now have very sophisticated modeling software available for Yagis, most of them based on the method of moments. See Chapter 4 to see what's available. Here are some things you should keep in mind:

- Make sure you know exactly what you want before you start: maximum boom length, maximum gain, maximum directivity, large SWR bandwidth, etc.
- Always model the antenna first in free space.
- Always model the antenna on a range of frequencies (eg, 7.0 to 7.3 MHz), so you can assess the SWR, gain and F/B of the design over the whole band.
- Make sure the feed-point impedance is reasonable (it can be anything between 18 Ω and 30 Ω, in some special cases 50 Ω).
- When the array is optimized and meets your requirements in free space, you should repeat the exercise over real ground at the actual antenna height, usually using a *NEC-2*-derived program such as *EZNEC*.
- If the antenna is stacked with other antennas, include the other antennas in the model as well. This is especially so when considering stacking Yagis for the same band. F/B may be totally ruined due to stacking. Stacks need to be optimized as stacks!
- If you consider making a Yagi with loaded elements, first model the full-size equivalent. When applying the loading devices, don't forget to include the resistance losses and possible parasitic capacitances or inductances.
- If you are about to model your own Yagi using loading devices, such as linear-loading stubs or capacity-loading wires, you should be very careful. The best approach is to first model the antenna using all wires of the same diameter. This should prove the feasibility of the concept. Next, you should determine the resonant frequencies of the individual elements, by removing other elements from the model. These resonant frequencies are excellent guides for the actual tune-up of the antenna.

3.2. Mechanical Design

Making a perfect electrical design of a low-band Yagi is a piece of cake nowadays with all the magnificent modeling software available. The real challenge comes when you have to turn your model into a mechanical design! When building a mechanically sound 40-meter Yagi, there is no room for guesswork. Don't ever take anything for granted when you are building a very large antenna. If you want your beam to survive the winds and ice loading you expect, you *must* go through a fair amount of calculations. The same holds true for an 80-meter Yagi, of course, but with the magnitude squared!

Physical Design of Yagi Antennas, by D. Leeson, W6QHS (now W6NL), (Ref 964) covers all aspects of mechanical Yagi design. Leeson uses the "variable area" principle to assess the influence of wind on the Yagi. The book unfortunately does not give any design examples of practical full-size 40 or for 80-meter Yagis. The only low-band antenna covered is the Cushcraft 40-2CD, a shortened 2-element 40-meter Yagi. Leeson's modification to strengthen the Cushcraft 40-2CD has become a classic, and is a must for everyone who has this antenna and who does not want to see it ripped to pieces in a storm.

Over the years standards dealing with mechanical issues for towers and antennas have evolved. The well- known EIA RS-222 standard has evolved from 222-C through suffix D and eventually to the RS-222-G standard. The E-version (and also ASCE 74) treats wind statistics and force on cylindrical elements more realistically than C and D, and the difference shows up in the question of forces on cylinders at an angle to the wind. This affects boom strength and rotating torque. The article by K5IU (Ref 958) uses the E approach, as well as the *ON4UN Low Band Software* modules dealing with boom strength and torque balancing.

Kurt Andress, K7NV, wrote an interesting software package that addresses all of the mechanical issues concerning antenna strength. *YS* (Yagi Stress) is easy to use, has lots of data about materials and tubing in easy-to-access form. For information see **k7nv.com/yagistress/**. A free trial download can be obtained from WXØB's Web site at **www.arraysolutions.com**.

All of these tools deal with static wind-load models. The question, of course, is how reliable all these models are in a complex aerodynamic situation. As Leeson puts it, "...but we're not dealing with mathematical models when the wind is roaring through here at 134 mi/h. Either model (C or E) results in booms that break upward in the wind if you ignore vertical gusting..." In particular locations, such as hilltop QTHs, there may be vertical updraft winds that can break a boom unless three-way boom guys are used. But these are rather extreme conditions, not the run-of-the mill situations.

The real proof of the pudding is in the building of big antennas, and even more so keeping them up year after year. The mathematics involved in calculating all the structural aspects of a low-band Yagi element are rather complex. It is a subject that is ideally suited for computer assistance. Together with my friend Roger Vermet, ON6WU, I have written a comprehensive computer program, *Yagi Design*, which was released in early 1988 and updated a few times since.

In addition to the traditional electrical aspects, *Yagi Design* tackles the mechanical-design aspects. This is especially of interest to the prospective builder of 40- and 80-meter Yagi antennas. While Yagis for the higher HF bands can be built "by feel," 40- and 80-meter Yagis require much closer attention if you want these antennas to stay up.

The different modules of the *Yagi Design* software are reviewed in Chapter 4 on low-band software. This book is not a textbook on mechanical engineering, but a few definitions are needed in order to better understand some of the formulas I use in this chapter.

3.2.1. Terms and Definitions

Stress: Stress is the force applied to a material per unit of cross-sectional area. Bending stress is the stress applied to a structure by a bending moment (also called momentum). Shearing stress is the stress applied to a structure by a shearing moment. The stress is expressed in units of force divided by units of area (usually expressed in kg/mm^2 or $lb/inch^2$).

Breaking Stress: The breaking stress is the stress at which the material breaks.

Yield Stress: Yield stress is the stress where a material suddenly becomes plastic (non-reversible deformation). The yield stress to breaking stress ratio differs from material to material. For aluminum the yield stress is usually close to the breaking stress. For most steel materials the yield stress is approximately 70% of the breaking stress. Never confuse breaking stress with yield stress, unless you want something to happen that you will never forget.

Elastic Deformation: Elastic deformation of a material is deformation that will revert to the original shape after removal of the external force causing the deformation.

Compression or Elongation Strain: Compression strain is the percentage change of dimension under the influence of a force applied to it. Being a ratio, strain is an abstract figure.

Shear Strain: Shear strain is the deformation of a material divided by the couple arm. It is a ratio and thus an abstract figure.

Shear Angle: This is the material displacement divided by the couple arm. As the angles involved are small, the ratio is a direct expression of the shear angle expressed in radians. To obtain degrees, multiply by $180/\pi$.

Elasticity Modulus: Elasticity modulus is the ratio stress/strain as applied to compression or elongation strain. This is a constant for every material. It determines how much a material will deform under a certain load. The elasticity modulus is the material constant that plays a role in determining the sag of a Yagi element. The elasticity modulus is expressed in units of force divided by the square of units of dimension (unit of area).

Rigidity Modulus: Rigidity modulus is the ratio shear-stress/strain as applied to shear strain. The rigidity modulus is the material constant that will determine how much a shaft (or tube) will twist under the influence of a torque moment (eg, the drive shaft between the antenna mast and the rotator). The rigidity modulus is expressed in units of force divided by units of area.

Bending Section Modulus: Each material structure (tube, shaft, plate T-profile, I-profile, etc) will resist a bending moment differently. The section modulus is determined by the shape as well as the cross-section of the structure. The section modulus determines how well a particular shape will resist a bending moment. The section modulus is proper to a shape and not to a material. The bending section modulus for a tube is given by:

$$S = \pi \times \frac{OD^4 - ID^4}{32 \times OD} \qquad \text{(Eq 13-1)}$$

where

OD = outer diameter of tube
ID = inner diameter of tube

The bending section modulus is expressed in units of length to the third power.

Shear Section Modulus: Different shapes will also respond differently to shear stresses. The shear stress modulus determines how well a given shape will stand stress deformation. For a hollow tube the shear section modulus is given by:

$$S = \pi \times \frac{OD^4 - ID^4}{16 \times OD} \qquad \text{(Eq 13-2)}$$

where

OD = outer diameter of tube

ID = inner diameter of tube

The bending section modulus is expressed in units of length to the third power.

3.2.2. Conversions

The metric system is used throughout this book. In this chapter I still use *kilogram-force* as a measure of *force*, while for some time the *Newton* really is the official unit of force. A possible reason for doing so: If you calculate with weights expressed in kg (kilogram), it is easier to express force in kg (kilogram-force). If you want to convert a force expressed in kg (kg-force) to N or even to pound-force, or a momentum expressed in kgm (kilogram-meter) to Nm (Newton-meter), inch-pounds of foot-pounds, here is a handy listing of a few conversions:

Force

1 kg (kilogram-force) = 9.807 Newton

1 Newton = 0.102 kg (kilogram-force)

1 kg (kilogram-force) = 2.2 pound-force

Weight

1 kg (kilogram) = 2.2 lb (pounds)

Moment (momentum)

1 kgm (kilogram-force meter) = 9.807 Nm (Newton meter)

1 kgm (kilogram-force meter) = 86.97 Inch-pounds = 7.23 foot-pounds

Length

1 km (kilometer) = 0.621 mile

1 m (meter) = 39.37 (inch) = 3.28 ft

1 cm = 0.3937 inch

1 mm = 0.03937 inch

Area

1 m^2 = 10.76 sq ft

1 a (are) = 100 m^2 = 0.0247 acre

1 ha (hectare) = 10,000 m^2 = 2.47 acres

Volume

1 l (liter) = 0.22 gallon = 0.264 US gallon

3.3. Computer-Designed 3-Element 40-Meter Yagi at ON4UN

Let us go through the design of a very strong and lasting 3-element full-size 40-meter Yagi. When I write "lasting" I mean "long lasting." The design I will describe has been up 20 years now, and it still looks like new, and has gone through many bad winds (150 km/h) and even a couple of quite unusual ice loading sessions.

This is not meant to be a step-by-step description of a building project, but I will try to cover all the critical aspects of designing a sound and lasting 40-meter Yagi. The Yagi described also happens to be the Yagi I have been using successfully over the past several years on 40 meters (it has brought several new European records in major contests on 40 meters). The design criteria for the Yagi are:

- Low Q, good bandwidth (>200 kHz), F/B optimized.
- Survival at wind speeds up to 140 km/h with the elements broadside to the wind.
- Maximum ice load 10 mm at 60 km/h wind.

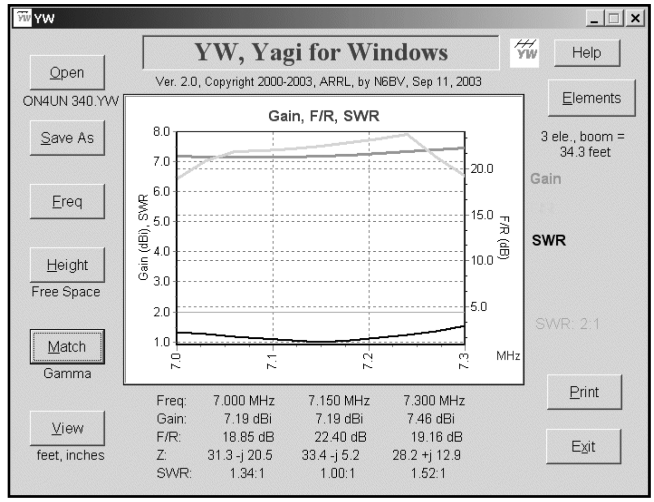


Fig 13-2 — Free-space performance data for full-sized 3-element Yagi design number 10 from the *Yagi Design* software suite. This was created by the *YW* (*Yagi for Windows*) program from the *ARRL Antenna Book* CD.

- Lifetime greater than 20 years
- Boom length 10.7 meters maximum (only because I happened to have this boom)

3.3.1. Selecting an Electrical Design

Design number 10 from the database of the *Yagi Design* software program meets all the above specifications. **Fig 13-2** shows a copy of the *YW* main screen for my 40-meter Yagi. While I could have selected another design with up to 0.5 dB more gain, I selected this design because of its excellent F/B pattern and wide bandwidth for SWR, gain and F/B.

I mounted this Yagi 5 meters above my 20-meter Yagi (design number 68 from the database), 30 meters above ground. The combination of both antennas was modeled once more over real ground at the final height using a *MININEC*-based modeling program, to see if there would be an important change in pattern and gain due to the presence of the second antenna. The performance figures (gain, F/B) and directivity pattern of the 40-meter Yagi changed very little at the 5-meter stacking distance.

3.3.2. Principles of Mechanical Load and Strength Calculations for Yagi Antennas

R. Weber, K5IU, brought to our attention (Ref 958) that the variable-area method, commonly employed by most Yagi manufacturers, and used by many authors in their publications as well as software, has *no* basis in science, nor is there any experimental evidence for the method.

The variable-area method assumes that the direction of the force created by the wind on an element is always in line with the wind direction, and that the magnitude is proportional to the area of the element as projected onto a plane perpendicular to the wind direction (proportional to the sine of the wind angle).

The scientifically correct method of analyzing the wind-force behavior, called the "cross-flow" principle, says that the direction of the force due to the wind is *always* perpendicular to the plane in which the element is situated and that its magnitude is proportional to the square of the sine of the wind angle.

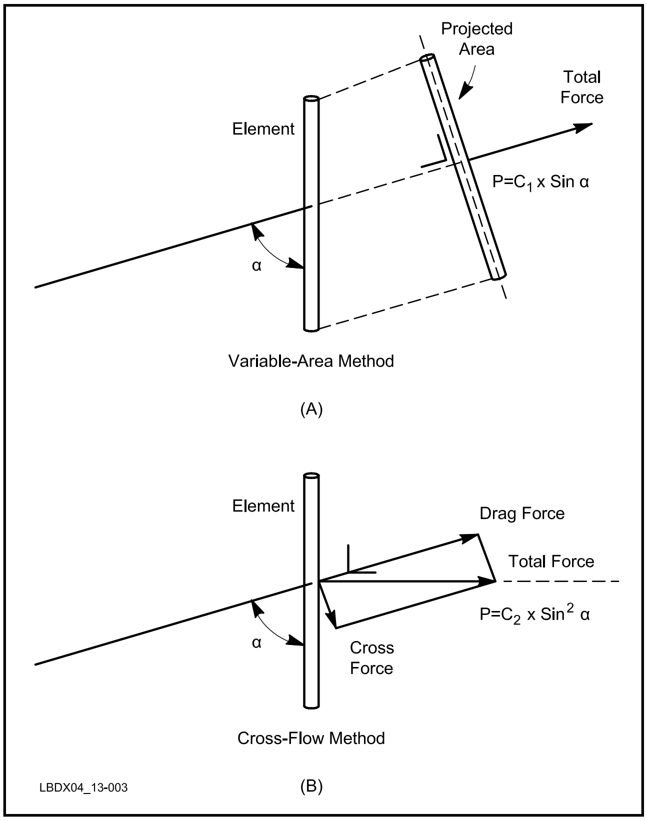


Fig 13-3 — In many cases, the amateur literature uses the "variable area" method shown at A for calculating the effect of wind on an element. The principle says that the direction of the force created by the wind on an element is *always* in-line with the direction of the wind, which is clearly incorrect. If this were correct, no plane would ever fly! The "cross-flow" principle, illustrated at B, states that the direction of the force is *always* perpendicular to the element, and is the resultant of two components, the drag force and the cross force (which is the lifting force in the case of an airplane wing). See text for details.

Fig 13-3 shows both principles. It is easy to see that the cross-flow principle is the correct one. The experiment described by K5IU can be carried out by anyone, and should convince anyone who has doubts: "Take a 1-meter long piece of aluminum tubing (approximately 25 mm in diameter) for a car ride. One person drives, while another sits in the passenger seat. The passenger holds the tube in his hand and puts his arm out the window positioning the tube vertically. The tube is now perpendicular to the wind stream (wind angle = zero). It is easy to observe a force (drag force) that is *in-line* with the wind (and at the same time perpendicular to the axis of the tube). The passenger now rotates the tube approximately 45°, top end forward. The person holding the tube will now clearly feel a force that pushes the tube *backward* (drag force), but at the same time tries to *lift* (cross force) the tube. The resulting force of these two components (the drag and cross force) is a force that is *always* perpendicular to the direction of the tube. If the tube is inclined with the bottom end forward, the force will try to push the tube downward."

This means that the direction of the force developed by the wind on an object exposed to the wind is not necessarily the same as the wind direction. There are some specific conditions where the two directions are the same, such as the case where a flat object is broadside to the wind direction. If you put a plate (1 meter^2) on top of a tower, and have the wind hit the plate at a 45° angle, it will be clear that the push developed by the wind hitting the plate will not be developed in the direction of the wind, but in the direction perpendicular to the plane of the flat plate. If you have any feeling for mechanics and physics, this should be fairly evident.

To remove any doubt from your mind, D. Weber states that Alexandre Eiffel, builder of the Paris Eiffel tower, used the cross-flow principle for calculating his tower. And it still stands there after more than 100 years.

Now comes a surprise: Take a Yagi, with the wind hitting the elements at a given wind angle (forget about the boom at this time). The direction of the force caused by the wind hitting the element at whatever wind angle will always be perpendicular to the element. This means that the force will be in-line with the boom. The force will not create any bending moment in the boom; it will merely be a compression or elongation force in the boom. All of this, of course, provided the element is fully symmetrical with respect to the boom.

This force in the boom should not be of any concern, as the boom will certainly be strong enough to cope with the bending moments caused by wind broadside to the boom. These bending moments in the boom at the mast attachment plate are caused only by the force created by the wind on the boom only (by the same "cross-flow" principle) or any other components that have an exposed wind area in-line with the boom.

If the mast-to-boom plate is located in the center of the boom, the wind areas on both sides of the mast are identical, and the bending moments in the boom on both sides of the mast (at the boom-to-mast plate) will be identical. This means there is no *mast torque*. If the areas are unequal, mast torque will result. This mast torque puts extra strain on the rotator, and should be avoided. Torque balancing can be done by adding a *boom dummy*, which is a small plate placed near the end of the shorter boom half, and which serves to reestablish the balance in bending moments between the left and the right side of the boom.

This may seem strange since intuitively you may have difficulty accepting that the extreme case of a Yagi having one element sitting on one end of a boom would not create any rotating torque in the mast, whatever the wind direction is. Surprisingly enough, this is the case. You cannot compare this situation with a weathervane, where the boom area at both sides of the rotating mast is vastly different. It is the vast difference in boom area that makes the weathervane turn into the wind.

Fig 13-4 shows the situation in theory, and what's likely to happen in the real world. At A and B the wind only sees the element (the boom is not visible), and if the element is fully symmetrical with respect to the boom, there will be no torque moment at the element-to-boom interface. Hence this is a fully stable situation. At C the situation where the boom is facing the wind is shown. The element is invisible now and as the boom is supposed to be wind-load balanced, the boom by itself creates no torque at the boom-mast interface. At D we see that the cross-flow principle only creates a force in-line with the boom. This means that this example still guarantees a well-balanced situation, and the structure will not rotate in the wind.

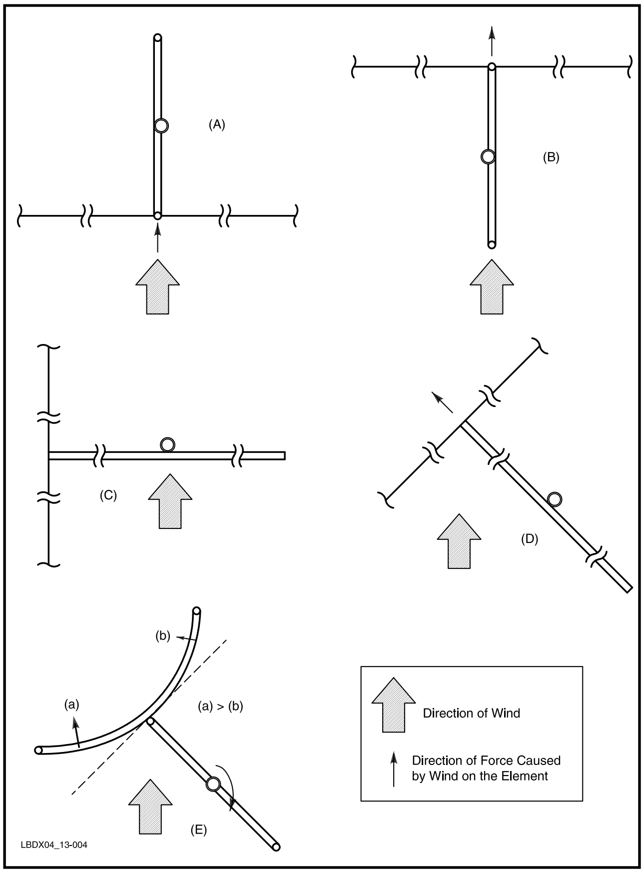


Fig 13-4 — Analysis of the influence of the wind on the mast torque for a single element sitting on the end of a boom. In all cases, A through D, no mast torque is induced. Only in case E, where the element is deformed by the wind, will mast torque be induced. See text for details.

But let's be practical. The wind blowing on the long flexible elements of a Yagi will make the elements bend slightly, as shown in Fig 13-4E. In this case now it is clear that the pressure induced by the wind on side (a) of the element will be much greater than on side (b) as side (a) now faces the wind much more than side (b). In this case the antennas will tend to rotate in the sense indicated by the arrow.

Taking all of this into account it seems to be a good idea not only to try to achieve full boom (area) symmetry but full element (area) symmetry as well. Leeson came to the conclusion that he prefers to balance in the element plane by offsetting the element ensemble to eliminate the need for a torque balancing element, then using a vane (boom torque compensating plate) on the now unbalanced boom. If offsetting the element ensemble creates an important weight imbalance, this can always be compensated for by inserting some form of weight in the boom near one tip.

Not adding extra dummy elements seems to be a good idea, as in dynamic situations (wind turbulence) these may actually deteriorate the situation rather than improve it. Since in principle the Yagi elements do *not* contribute to the boom moments, and therefore not to the mast torque, it makes no sense to create dummy elements to try to achieve a torque-balanced Yagi.

The Mechanical Yagi Balance module of the *Yagi Design* software (available on the CD that comes with this book) addresses all the issues as explained above and uses the cross-flow principle. It uses latest data from the latest EIA/TIA-222-E specification, which is somewhat different from the older EIA standard RS-222-C.

3.3.3. Element Strength Calculation

While it is standard procedure to correct boom sag using truss cables, element sag must be controlled to a maximum degree by using the properly designed tapered sections for making the element. Guyed elements are normally only used with 80-meter Yagis. Unguyed 40-meter full-size tubular elements (24 meters long) can be built to withstand very high wind speeds, as well as a substantial degree of ice loading.

The mathematics involved are quite tedious, and a very good subject for a computer program. Leeson (Ref 964) addresses the issue in detail in his book, and he made a spreadsheet type of program available for calculating elements. As the element-strength analysis is always done with the wind blowing broadside to the elements, the issues of variable area or cross-flow principle don't have to be taken into consideration.

The Element Strength module of the *Yagi Design* software is a dedicated software program that allows the user to calculate the structural behavior of Yagi elements with up to nine tapering elements. This module operates in the English measurement system as well as in the metric system (as do all other modules of the integrated *Yagi Design* software). A drag factor of 1.2 is used for the element calculations (as opposed to 0.66 in the older RS-222-C standard).

Interactive designing of elements enables the user to achieve element sections that are equally loaded. Many published element designs show one section loaded to the limit, while other sections still exhibit a large safety margin. Such unbalanced designs are always inefficient with respect to weight, wind area and load, as well as cost.

Each change (number of sections, section length, section diameter, wind speed, aluminum quality, ice load, etc) is immediately reflected in a change of the moment value at the interface of each taper section, as well as at the center of the element. When a safe limit is exceeded, the unsafe value will blink. The screen also shows the weight of the element, the wind area, and the wind load for the specified wind speed.

It is obvious that the design in the first place will be dictated by the material available. Material quality, availability and economical lengths are discussed in Section 3.3.6.

A 40-meter Yagi reflector is approximately 23 meters long. This is twice the length of a 20-meter element. Designing a good 40-meter element can be done starting from a sound 20-meter element, which is then lengthened by more tapered sections toward the boom, calculating the bending stresses at each section drop.

When designing a Yagi element you must make sure that the actual bending moments (LM_t) at all the critical points match the maximum allowable bending moments (RM) as closely as possible. LM_v is the bending moment in the vertical plane, created by the weight of the element. This is the moment that creates the sag of the element. LM_t is the sum of LM_v and the moment created by the wind in the horizontal plane. Adding those together may seem to create some safety, although it can be argued that turbulent wind may in actual fact blow vertically

Fig 13-5 — No, this is not a 15-meter Yagi on top of a 6-meter Yagi, as you might judge from the very small degree of sag in the elements, but the 3-element 40-meter Yagi at ON4UN. The antenna is 30 meters high, 5 meters above a 20-meter Yagi (15-meter boom). The amount of sag in an element is a very good indicator of the mechanical strength (read: wind resistance) of the element.

in a downward direction.

The reflector element for my 40-meter Yagi uses material with metric dimensions available in Europe. The design was done for a maximum average wind speed of 140 km/h, using F22 quality (Al Mg Si 0.5%) material. This material has a yield strength of 22 kg/mm^2 (31,225 lb/inch^2). For material specifications see Section 3.3.6.

All calculations are done for a static condition. Dynamic wind conditions can be significantly different, however. The highest bending moment is at the center of the element. Inserting a 2-meter long steel tube (5 or 7-mm wall) in the center of the center element will not only provide additional strength but also further reduce the sag.

Whether 140 km/h will be sufficient in your particular case depends on the following factors:

- The rating of the wind zone where the antenna is to be used. The latest EIA/TIA-222 standard lists the recommended wind speed by county in the US.
- Whether modifiers or safety factors are recommended (see EIA/TIA-222 standard).
- Whether you will expose the element to the wind or put the boom into the wind (see Section 3.3.4).
- Whether you have your Yagi on a crank-up tower, so that you can nest it at protected heights during high wind storms.

Fig 13-5 shows the 3-element full-size 40-meter Yagi placed 5 meters above my 5-element 20-meter Yagi, which has a similar taper design. Note the very limited sag on the elements. The telescopic fits are discussed in Section 3.3.7. **Figs 13-6** and **13-7** show the section layout of the 40-meter reflector element, calculated for both metric and US (inch) materials.

3.3.3.1. Element Sag

Although element sag is not a primary design parameter, I included the mathematics to calculate element sag in the Element Strength module of the *Yagi Design* software. While designing, it is interesting to watch the total element sag. Minimal element sag is an excellent indicator of a good mechanical design. Too much sag means there is somewhere along the element too much weight that does not contribute to the strength of the element. The sag of each of the sections of an element depends on:

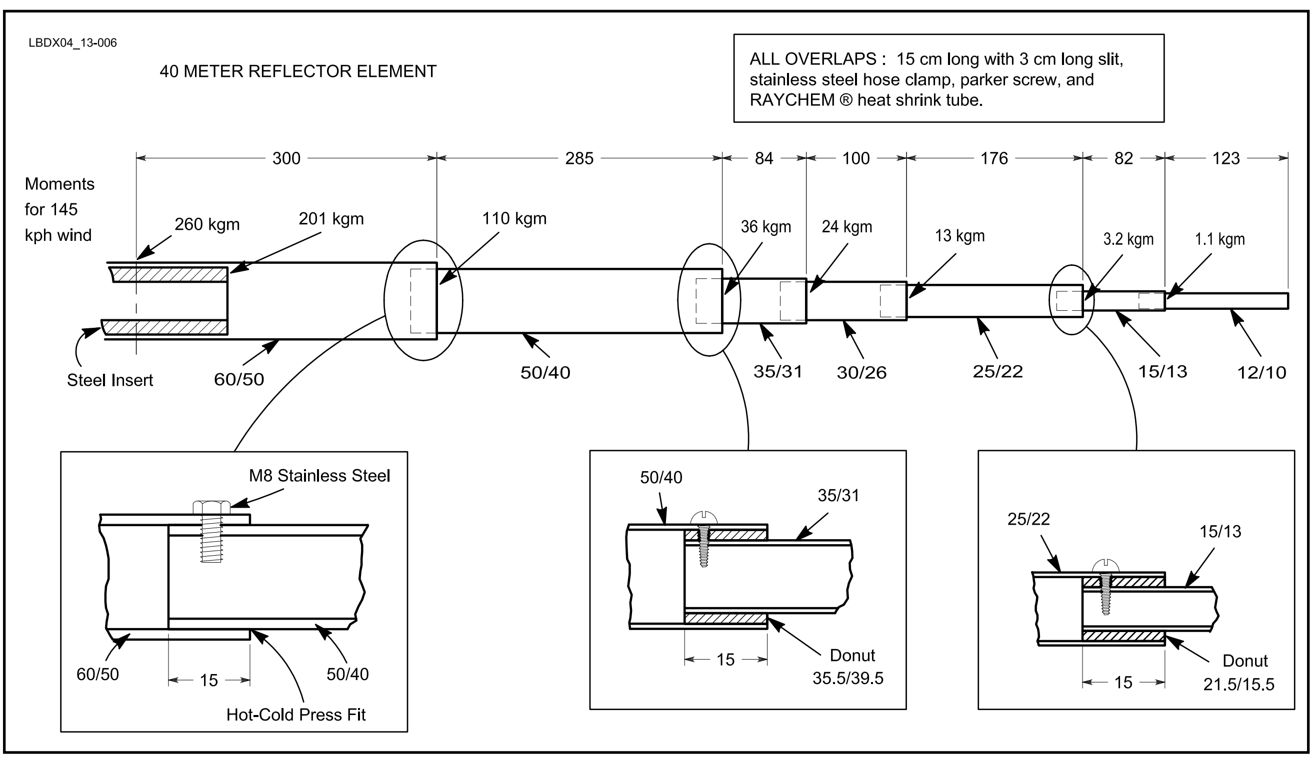


Fig 13-6 — Mechanical layout of a 40-meter full-size reflector using metric materials.

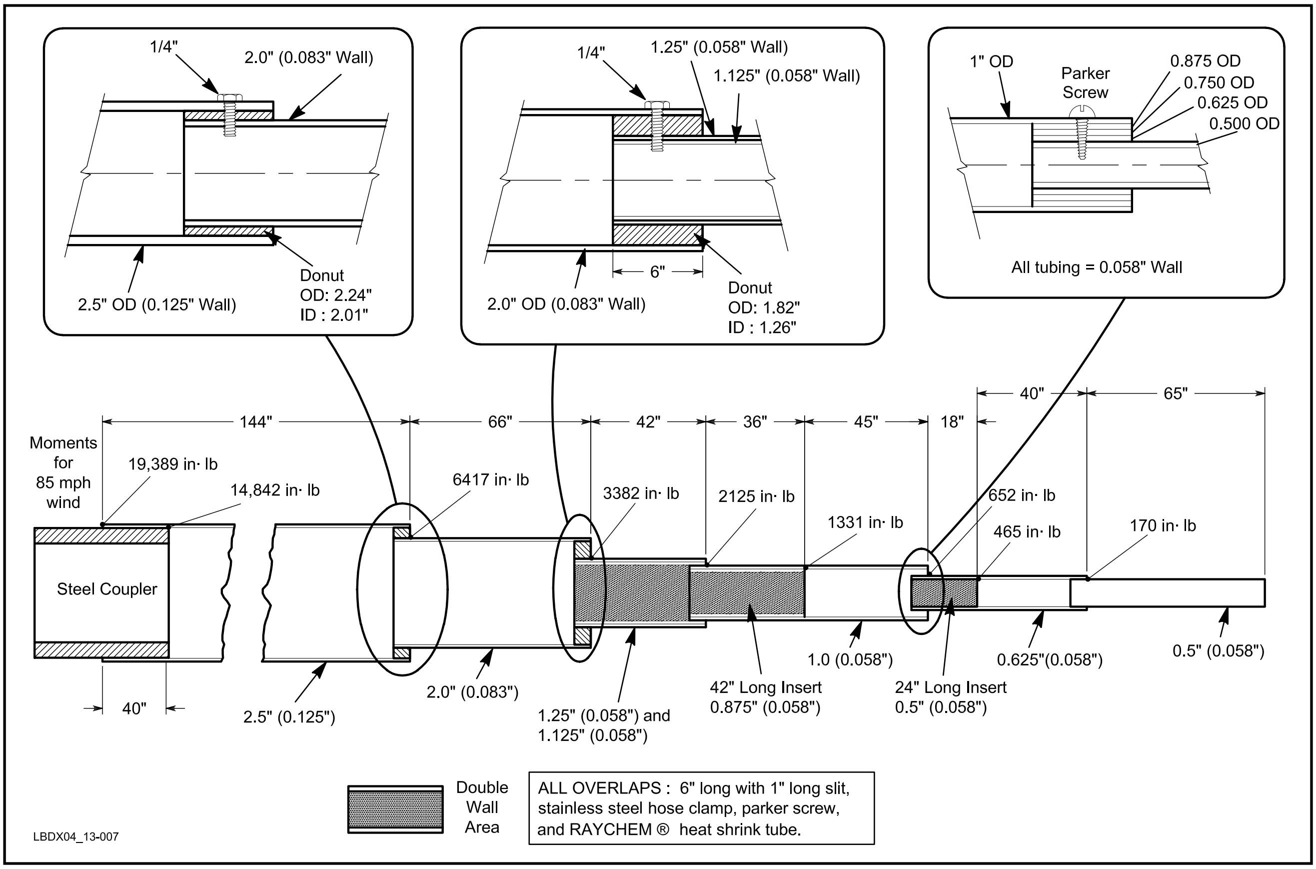


Fig 13-7 — Layout of the 40-meter reflector using US materials (inch dimensions).

- The section's own weight.
- The moment created by the section(s) beyond the section being investigated (toward the tip).
- The length of the section.
- The diameter of the section.
- The wall thickness of the section.
- The elasticity modulus of the material used.

The total sag of the element is the sum of the sag of each section. The elasticity modulus is a measure of how much a material can be bent or stretched without inducing permanent deformation. The elasticity modulus for all aluminum alloys is 700,000 kg/cm^2 (9,935,000 lb/inch^2). This means that an element with a stronger alloy will exhibit the same sag as an element made with an alloy of lesser strength.

The 40-meter reflector designed above has a calculated sag of 129.5 cm, not taking into account the influence of the steel insert (coupler). The steel coupler reduces the sag to approximately 91 cm. These are impressive figures for a 40-meter Yagi. With everything scaled down properly, the sag is comparable to that of most commercial 20-meter Yagis. After mounting the element, the total element sag was that calculated by the software.

3.3.3.2. Alternative Element Designs Using US Materials

The US design is made by starting from standard tubing lengths of 144 inches. The standard dimensions commonly available in the US and availability of aluminum tubes and pipes is discussed in Section 3.3.6.

For the two larger-diameter tubes, I used aluminum pipe. The remaining sections are from the standard tubing series with 0.058-inch wall thickness. From the design table we see that for some sections I used a wall thickness of 0.11 inch, which means that we are using a tight-fit section of ⅛-inch less diameter as an internal reinforcement.

The design table shows that the center sections would marginally fail at a 90-mi/h design wind speed. In reality this will not be a problem, since this design requires an internal coupler to join the two 144-inch center sections. This steel coupler must be strong enough to take the entire bending moment. The section modulus of a tube is given by Eq 13-1:

$$S = \pi \times \frac{OD^4 - ID^4}{32 \times OD}$$

where
S = section modulus
OD = tube outer diameter
ID = tube inner diameter

The maximum moment a tube can take is given by:

$$M_{max} = YS \times S \qquad \text{(Eq 13-3)}$$

where
YS = yield strength of the material
S = section modulus as calculated above
or

$$M_{max} = YS \times \pi \times \frac{OD^4 - ID^4}{32 \times OD} \qquad \text{(Eq 13-4)}$$

The yield strength varies to a very large degree (Ref 964 p 7-3). For different steel alloys it can vary from 21 kg/mm^2 (29,800 lb/inch^2) to 50 kg/mm^2 (71,000 lb/inch^2).

A 2-inch OD steel insert (with aluminum shimming material) made of high-tensile steel with a YS = 55,000 lb/inch^2 would require a wall thickness of 0.15 inch to cope with the maximum moment of 19.622 inch-lb at the center of the 40-meter reflector element.

Note that the element sag (42.1 inches with a 2 × 40-inch-long steel coupler) is very similar to the sag obtained in the previous metric design example. It is obvious that for an optimized Yagi element (and for a given survival wind speed), the element sag will always be the same, whatever the exact taper scheme may be. In other words, a good 40-meter Yagi reflector element, designed to withstand a 140 km/h (87 mi/h) wind should not exhibit a sag of more than 40 inches (100 cm) when constructed totally of tubular elements. More sag than that proves it is a poor design.

3.3.3.3. The Driven Element and the Director

Once we have designed the longest element, we can easily design the shorter ones. We should consider taking the "left over" lengths from the reflector for use in the director. The lengths of the different sections for the 3-element Yagi number 10 from the *Yagi Design* database, according to the metric and US systems, are shown in **Table 13-1**. Typically, if the reflector is good for 144 km/h, the director and the driven element will withstand 160 to 170 km/h.

3.3.3.4. Final Element Tweaking

Once the mechanical design of the element has been finalized, the exact length of the element tips will have to be calculated using the Element Taper module of the software. You can also use a modeling programs such as *EZNEC* or *YW* and enter all the tapered sections directly.

Table 13-1

Element Design Data for the 3-Element 40-Meter Yagi Reflector, Driven Element and Director

Section	*OD/Wall*	*Dir.*	*Dr. Ele.*	*Refl.*
1	60/5	300	300	300
2	50/5	285	285	285
3	35/2	60	85	84
4	30/2	60	112	100
5	25/1.5	135	135	176
6	15/1	60	80	82
7	12/1	111	80	113
Total length (cm)		1011	1077	1150

Section	*OD/Wall*	*Dir.*	*Dr. Ele.*	*Refl.*
1	2.375/0.154	144	144	144
2	2.00/0.109	55	66	66
3	1.25/0.11	34	42	50
4	1.00/0.11	30	30	30
5	1.00/0.058	30	38	42
6	0.625/0.11	18	15	21
7	0.625/0.058	28	30	34
8	0.50/0.058	60	63	65
Total length (inches)		399	428	452

Note: This design assumes a boom diameter of 75 mm (3 inches) and U-type clamps to mount the element to the boom (L = 300 mm, W = 150 mm, H = 70 mm). Availability of materials will be the first restriction when designing a Yagi antenna.

3.3.4. Boom Design

Now that we have a sound element for the 40-meter Yagi, we must pay attention to the boom. When the wind blows at a right angle to the boom, the maximum pressure is developed on the boom area. At the same time, the loading on the Yagi elements will be minimum. There is no intermediate angle at which the loading on the boom is higher than at a 90° wind angle, when the wind blows broadside onto the boom.

3.3.4.1. Pointing the Yagi into the Wind

We all have heard the question, "Should I point the elements into the wind, or should I point the boom into the wind?" The answer is simple. If the area of the boom is smaller than the area of all the elements, then put the boom perpendicular to the wind. And vice versa.

Let me illustrate this with some figures for a 40-meter Yagi. Calculations are done for a 140 km/h wind, with the boom-to-mast plate in the center of the boom. The figures below were calculated in the Mechanical Yagi Balance module of the *Yagi Design* software.

Zero-degree wind angle (wind blowing broadside to the elements):

- Boom moment in the horizontal plane: Zero
- Thrust on tower/mast 323 kg (force)
- Maximum bending moment in the elements

90° wind angle (wind blowing broadside to the boom):

- Boom moment 114 kgm (bending moment)
- Thrust on tower/mast: 87 kg (force)
- Minimum bending moment in the elements

In this case it is obvious that we should at all times try to put the boom perpendicular to the wind during a storm with high winds. For calculating and designing the rotating mast and tower, I recommend, however, that you take into account the worst-case wind pressure of 323 kg.

What about a long-boom HF Yagi? For a 6-element 10-meter Yagi, putting the elements perpendicular to the wind would be the logical choice. But relying on the exact direction of the Yagi as a function of wind direction is a dangerous practice and I don't want to encourage this. This does not mean that in case of high winds you couldn't take advantage of the best wind angle to relieve load on the Yagi, mast or tower, but what is gained by doing so should only be considered as extra safety margin only!

3.3.4.2. Weight Balancing

In **Fig 13-8**, I assumed that the mast is at the physical center of the boom. As the driven element is offset toward the reflector, the Yagi will not be weight balanced. A good physical design must result in a perfect weight balance, since it is extremely difficult to handle an unbalanced 40-meter monster on a tower when trying to mount it to the rotating mast. The obvious solution is to shift the mast attachment point in such a way that a perfect balance is achieved.

The Mechanical Yagi Balance module of my software calculates weight-balancing for a Yagi. It automatically calculates the area of the required boom dummy plate to reestablish

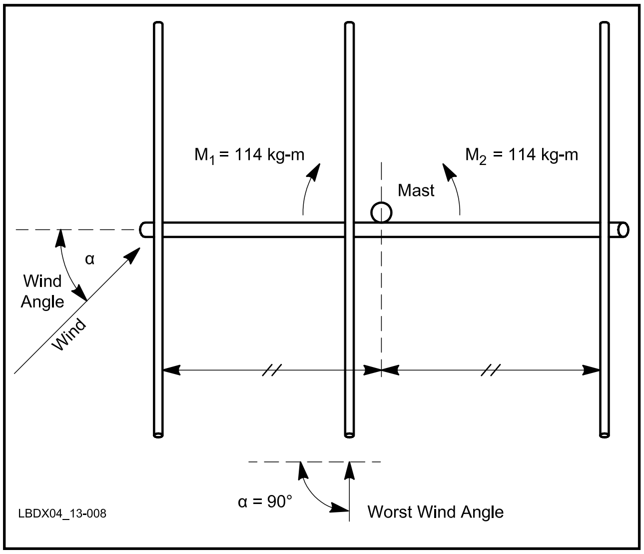


Fig 13-8 — Boom moments in the horizontal plane as a result of the wind blowing onto the boom and the elements. The forces produced by the wind on the Yagi elements do not contribute to the boom moment; they only create a compression force in the boom (see text). The highest boom moments occur when the wind blows at a 90° angle, broadside to the boom.

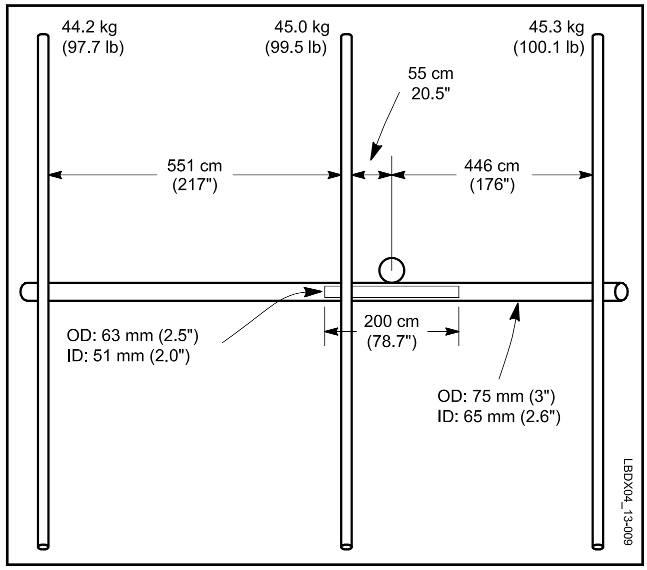


Fig 13-9 — Weight-balanced layout of the 3-element 40-meter Yagi, showing the internal boom coupler. The net weight, without a match box (containing the Gamma or Omega matching capacitors) and without the boom-to-mast plate is 183 kg.

torque balance. Components taken into account for calculating the weight balance are:

- The Yagi elements
- The boom
- The boom coupler (if any)
- The boom dummy (see Section 3.3.4.3 below)
- The match box (box containing gamma/omega matching components).

Fig 13-9 shows the layout that produces perfect weight balance. In our example I have assumed no match box. Slightly offsetting the driven element of the 3-element Yagi avoids the conflict between the location for the mast and for the driven element attach point.

3.3.4.3. Yagi Torque Balancing

The cause of mast torque has been explained in Section 3.3.2. If the bending moment in the boom on one side of the mast is not the same as the bending moment at the other side of the mast, we have a net *mast torque*. One moment is trying to rotate the mast clockwise, while the other tries to rotate the mast counterclockwise.

Only when the boom areas on both sides of the mast are identical will the Yagi be perfectly torque-balanced. The wind area of the elements and their placement on the boom do *not* play any role in the mast torque, as the direction of the force developed by the wind on an element is always perpendicular to the element itself, which means in-line with the boom. As such, element wind area cannot create a boom moment, but merely loads the boom with compression or elongation.

It is the mast torque that makes an antenna *windmill* in high winds. A good mechanical design must be torque-free at all wind angles. During our weight-balancing exercise earlier, we shifted the mast attachment point somewhat to reestablish weight balance. This causes the boom moments on both sides of the mast to become different. To reestablish balance, we mount a small *boom dummy plate* near the end of the shorter boom half. This plate has an area of 133 cm^2 and should be mounted 50 cm from the reflector for torque-balance.

3.3.4.4. Boom Moments

I calculated the boom moments after torque-balancing and found that the boom bending moments have increased slightly, from 114 kgm for the "non-weight-balanced Yagi" to 120 kgm after weight balancing and adding the boom dummy. This is a negligible price to pay for having a weight-balanced Yagi.

The software calculates everything related to the boom design. The material stresses are computed for the coupler, as well as for the boom. The boom stress is only meaningful if the boom is not split in the center. With a split boom it is the coupler that takes the entire stress.

Even for a 140 km/h wind, the stresses in the boom are low. But as we will likely point the boom into the wind in windstorms (Section 3.3.4.1), we should build in a lot of safety. Also, as mentioned before, the 140 km/h does not include any safety factors or modifiers, as may be prescribed in the standard EIA/TIA-222.

To me, it is proof of *poor* engineering to design a boom that needs support guys to make it strong enough to withstand the forces from the wind and the bending moments caused by it. If guy wires are employed to provide the required strength, guying will have to be done in both the horizontal as well as the vertical plane. Guy wires can be used to eliminate boom sag. This will only be done for cosmetic rather than strength reasons.

Three-way guying may be necessary where vertical gusts can be expected (hilltop QTHs) to prevent the boom from dancing up and down due to vertical updrafts.

3.3.4.5. Boom Sag

The boom as now designed will withstand 140-km/h

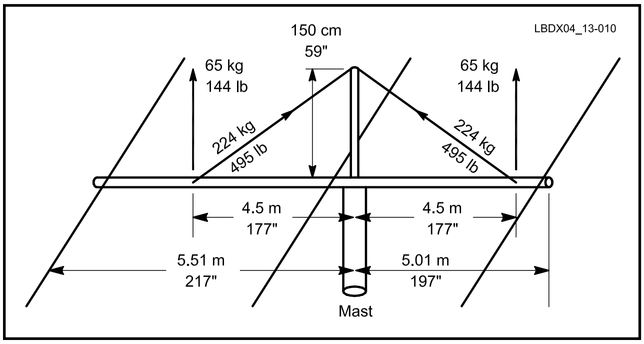


Fig 13-10 — Layout of the boom-support cables (trusses) with the forces and tensions involved. The truss cables are not installed to provide additional strength to the boom; they merely support the boom in order to compensate for the sag from the weight of the elements on the boom.

winds, with a good safety factor. The same boom, however, without any wind loading will have to endure a fair bending moment in the vertical plane, caused by the weight of the elements and the weight of the boom itself.

Fig 13-10 shows the forces and dimensions that create these bending moments. The weight moments were obtained earlier when calculating the Yagi weight balance.

Weight moments to the "left" of the mast:
Element no. 1: –226.6 kgm
Element no. 2: –24.5 kgm
Boom left: –37.1 kgm
Boom insert left: –4.2 kgm
Boom dummy: –0.3 kgm
Total: –292.7 kgm

Weight moments to the "right" of the mast:
Element no. 3: 243.6 kgm
Boom right: 45.2 kgm
Boom insert right: 4.2 kgm
Total: 293 kgm

The weight moment to the left of the mast is the same as to the right of the mast since the Yagi is weight-balanced. Here comes another surprise: The boom is loaded almost *three times* as much by weight loading in the vertical plane (293 kgm) than it is by wind loading at 140 km/h in the horizontal plane (120 kgm).

The maximum allowable bending moment for the boom steel insert with a diameter of 63 mm and 6 mm wall is 619 kgm as calculated with Eq 13-2 for a material yield strength of 20 kg/mm^2. This steel coupler has a safety factor of *two* as far as the weight-loading in the vertical plane is concerned. Boom stress by weight will usually be the condition that will specify the size of the boom with large low-band Yagis using heavy elements.

The boom, using the above calculated coupler, does not require any guying for additional strength. However, the high weight loading of the very long elements sitting at the end of the boom halves will cause a very substantial sag in the boom. For my 40-meter beam the sag amounts to nearly 65 cm, which is really excessive from a cosmetic point of view. A sag of 10 cm is due to the boom's own weight and 55 cm is due to the weight of the elements at the tips of the boom.

Again, I consider it a proof of good engineering to eliminate sag by supporting the boom using truss cables. The two boom halves are supported with two sets of dual parallel guy wires attached on the boom at a point 4.5 meters from the mast attachment point. The guy wires are supported from a 1.4-meter high support mast made of a 35-mm OD stainless steel tube, which is welded to the boom-to-mast plate. See Fig 13-10.

The weight that is supported is given by the previously calculated moment divided by the distance of the cable attachment point to the mast attachment point. Assuming the two boom halves are hinged at the mast, each support cable would have to support the total weight as shown above, divided by the sine of the angle the truss support cable makes with the boom.

Leeson (Ref 964) covers guyed booms well in his book. In the case above we are *not* guying the boom to give it additional strength, we do it only to eliminate sag. Guying a boom is not a simple problem of moments, but a problem of a compressed column, where the slenderness of the boom and the compression force caused by the guy wire (usually in three directions) come into the picture. In our case these forces are so low that we can simplify the model as done above. In the above case we implicitly assumed that the boom has enough lateral strength (which we had calculated). For solving the wire-truss problem we assume that the boom is a "nonattached" cantilever. The fact that the boom is attached introduces an additional safety factor.

If a single steel cable is used, a 6-mm OD cable is required to safely support this weight. I use *two* cables of 4-mm OD Kevlar (also known as Phillystran in the US). I use this because it was available at no cost, and it does not need to be broken up with egg insulators (Kevlar is a fully dielectric material which has the same breaking strength as steel and the same elongation under load). Note that turnbuckles may prove to be the weak link in the system and stainless-steel turnbuckles can be very expensive. If two parallel cables are used, a tension equalizer must be used to ensure perfect equal stress in both cables. In the case of two truss cables without equalization, one of the cables is likely to take most of the load.

Let me go into detail why I use two parallel support guys. **Fig 13-11** shows the top of the support mast, on which two

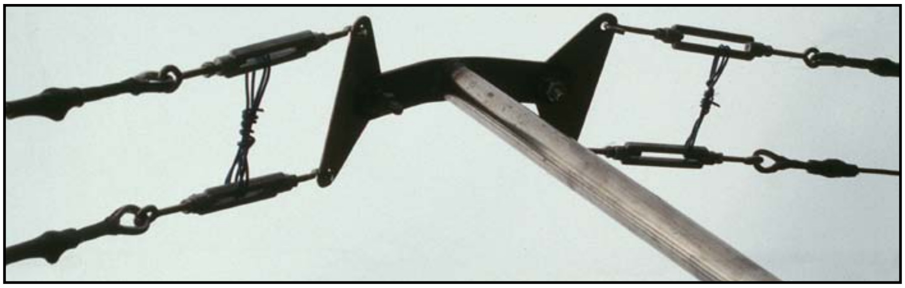


Fig 13-11 — Details of the tension-equalizing system at the top of the support mast, where the two boom-support trusses are attached. The triangular-shaped plate can rotate freely around the 10-mm bolt, which serves to equalize the tensions in the two guy wires. See text for details.

triangular-shaped stainless-steel plates are mounted. These plates can pivot around their attachment point, which consists of a 1-cm diameter stainless-steel bolt. The two guy wires are connected with the correct hardware (very important — consult the supplier of the cable!) at the base of these triangular pivoting plates. The pivoting plates now serve a double purpose:

- They equalize the tension in the two guy wires.
- They serve as a visual indicator of the status of the guy wires.

If something goes wrong with one of the support wires, the triangular plate will pivot around its attachment point. At the same time the remaining support (if properly designed) will still support the boom, although with a greatly reduced safety factor.

To install the support cables and adjust the system for zero or minimum boom sag if you don't use turnbuckles, place the beam on two strong supports near the end of the boom so as to induce some inverse sag in the boom. Lift the center of the boom to control the amount of inverse sag. Now adjust the position of the boom attachment hardware to obtain the desired support behavior.

Make sure you properly terminate the cables with thimbles. The loads involved are not small, and improper terminations will not last. This is especially true when Kevlar rope is used.

3.3.5. Element-to-Boom and Boom-to-Mast Clamps

With an element weighing well over 40 kg, attaching such a mast at the end of a 5-meter arm must be done with great care. The forces involved when we rotate the Yagi (start and stop) and when the beam swings in storm winds are impressive.

After an initial failure, I designed an element-to-boom mounting system that consists of three stainless-steel U-channel profiles (50 cm long) welded together. The element is mounted inside the central channel profile using four U bolts with 12-mm wide aluminum saddles. Four double-saddle systems are used to mount the unit onto the boom (see **Fig 13-12**). U bolts must be used together with saddles and you must use saddles on both sides. The bearing strength of U bolts is far too low to provide a durable attachment under extreme wind loads without saddles on both sides. Never use U bolts made of threaded stainless-steel rods directly on the boom; if they can move but a hair, they become like perfect files that will machine a nice groove in the boom in no time!

At the center of the boom I mounted a 60-cm wide, 1-cm thick stainless-steel plate to which the 1.5-meter long support mast for the boom guying is welded. The boom is bolted to the boom-to-mast plate using eight U bolts with saddles matching the 75-mm OD boom (see **Fig 13-13**). On the tower, this plate

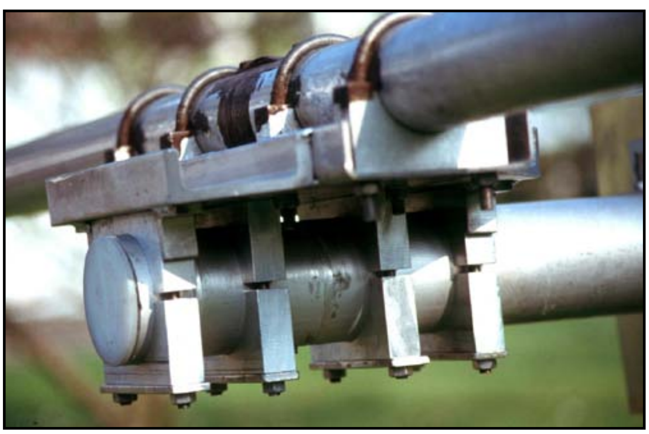


Fig 13-12 — The element-to-boom mounting system as used on the ON4UN 40-meter Yagi.

Table 13-2

Dimensions and Weight of Aluminum Tubing in F22 Quality

OD mm	*Wall mm*	*Weight g/m*	*OD mm*	*Wall mm*	*Weight g/m*
10	1	76	40	1.5	489
12	1	93	44	2	541
13	1	103	48	1.5	603
14	1	110	50	5	1923
15	1	127	50	2	820
19	1.5	227	52	1.5	654
20	1.5	235	57	2	940
22	2	339	60	5	2350
22	1.5	261	60	3	1460
25	2.5	477	62	2	1040
25	2	398	70	5	2757
25	1.5	298	70	3	1718
28	1.5	336	80	5	3181
30	3	687	80	4	2579
30	2	484	84	2	1385
32	1.5	387	90	5	3605
35	2	564	100	5	4029
36	1.5	438	100	2	1676
40	5	1495	110	5	4485
40	2	644			

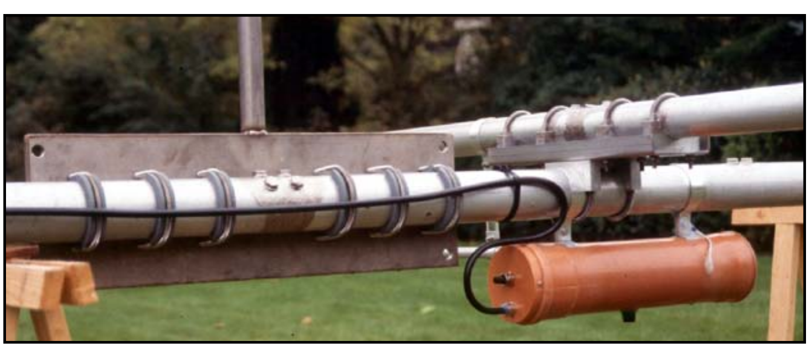


Fig 13-13 — The Omega matching system and plastic "drainpipe" box containing the two variable capacitors. Note also the boom-to-mast mounting plate made of 1-cm thick stainless steel. The boom is attached to this plate with eight U bolts and double saddles.

Table 13-3
List of Currently Available Aluminum Tubing in the US

OD inches	*Wall inches*	*Weight lb/foot*	*OD inches*	*Wall inches*	*Weight lb/foot*
0.250	0.058	0.04	1.625	0.058	0.34
0.375	0.058	0.07	1.750	0.058	0.36
0.500	0.058	0.10	1.750	0.083	0.51
0.625	0.058	0.12	1.875	0.058	0.39
0.750	0.058	0.15	2.000	0.065	0.45
0.875	0.058	0.18	2.000	0.083	0.59
1.000	0.058	0.20	2.000	0.125	0.83
1.125	0.058	0.23	2.500	0.065	0.59
1.250	0.058	0.26	2.500	0.083	0.74
1.375	0.058	0.28	2.500	0.125	1.06
1.500	0.058	0.31	3.000	0.065	0.71
1.500	0.065	0.34	3.000	0.125	1.30
1.500	0.083	0.43			

Table 13-4
List of Currently Available Aluminum Pipe in the US

OD	*Wall*	*OD*	*Wall*
1.050	0.113	1.900	0.109
1.050	0.154	1.900	0.145
1.315	0.133	1.900	0.200
1.315	0.179	2.375	0.065
1.660	0.065	2.375	0.109
1.660	0.109	2.375	0.154
1.660	0.140	2.375	0.218
1.660	0.191	2.875	0.203
1.900	0.065	2.875	0.276

is bolted to an identical plate (welded to the rotating mast) using four 18-mm OD stainless-steel bolts.

3.3.6. Materials

In the metric world (mainly Europe), aluminum tubes are usually available in 6-meter sections. **Table 13-2** lists dimensions and weights of a range of readily available tubes. Aluminum tubing in F22 quality (Al Mg Si 0.5%) is readily available in Europe in 6-meter lengths. The yield strength is 22 kg/mm^2.

Tables 13-3 and **13-4** show a range of material dimensions that are available in the US. *The ARRL Antenna Book* also lists a wide range of aluminum tubing sizes. Make sure you know which alloy you are buying. The most common aluminum specifications in the US are:

6061-T6: Yield strength = 24.7 kg/mm^2
6063-T6: Yield strength = 17.6 kg/mm^2
6063-T832: Yield strength = 24.7 kg/mm^2
6063-T835: Yield strength = 28.2 kg/mm^2

When designing the Yagi elements, a maximum effort should be made to use full fractions of the 6-meter tubing lengths, in order to maximize the effective use of the material purchased. A proper section overlap is 10 to 15 cm. The effective net lengths of fractions of a 600 cm tube are 285, 185, 135, 85 and 60 cm.

In the US, aluminum is available in 12-ft lengths. The effective economical cuts (excluding the 6-inch overlap) are 66, 42, 30, 22.8 inches, etc.

3.3.7. Telescopic Fits

You can make well-fitting telescopic joints as follows: With a metal saw, make two slits of approximately 30-mm length into the tip of the larger section. To avoid corrosion, use plenty of Penetrox or other suitable contact grease when assembling the sections. A stainless-steel hose clamp will tighten the outer element closely onto the inner one (with shimming material in between if necessary). A stainless-steel Parker screw will lock the sections lengthwise. For large diameters and heavy-wall sections, a stainless steel 6- or 8-mm bolt is preferred in a pre-threaded hole.

Metric tube sections do not provide as snug a telescoping fit as do the US series with a 0.125-inch-diameter step and 0.058-in. wall thickness. At best there is a 1mm difference between the OD of the smaller tube and the ID of the larger tube. A fairly good fit can be obtained, however, by using a piece of 0.3-mm-thick aluminum shimming material. The slit, hose clamp, Parker screw and heat-shrink tube make this a reliable joint as well.

Sometimes sections must be used where the OD of the smaller section is the same as the ID of the larger section. To achieve a fit, make a slit approximately 5 cm long in the smaller tube. Remove all burrs and then drive the smaller tube inside the larger to a depth of 3 times the slit length (eg, 15 cm). Do this after heating up the outer tube with a flame torch and cooling down the inner tube in ice water. The heated-up outer section will expand, while the cooled-down inner section will shrink. Use a good-sized plastic hammer and enough force to drive the inner tube quickly inside the larger tube before the temperature-expansion effect disappears. A solid unbreakable press fit can be obtained. A good Parker screw or stainless-steel bolt (with pre-threaded hole) is all that's needed to secure the taper connection.

Under certain circumstances a very significant drop in element diameter is required. In this case a so-called *doughnut* is required. The doughnut is a 15-cm long piece of aluminum tubing that is machined to exhibit the right OD and ID to fill up the gap between the tubes to be fit. Often the doughnut can be made from short lengths of heavy-wall aluminum tubing.

I always cover each taper-joint area with a piece of heat-shrinkable tube that is coated with hot-melt glue on the inside. This protects the element joint and keeps the element perfectly watertight.

3.3.8. Material Ratings and Design Conditions

All the above calculations are done in a static environment, assuming a wind blowing horizontally at a constant speed. Dynamic modeling is very complex and falls out of the scope of this book. If all the rules, the design methodology and the calculating methods as outlined above and as used in the mechanical design modules of the *Yagi Design* software are closely followed, a Yagi will result that will withstand the forces of wind, even in a normal dynamic environment, as has been proved in practice. My 40-meter Yagi was designed to be able to withstand wind speeds of 140 km/h, according to the EIA/TIA-222-E standard. The 140-km/h wind does *not* include any safety factors or other modifiers.

The most important contribution of all the above calcula-

Table 13-5
Ice Loading Performance of the 40-Meter Beam

Radial Ice		*Max Wind Speed*		*Sag*	
mm	*inch*	*kph*	*mph*	*cm*	*inch*
2.5	0.1	116	72	132	52
5.0	0.2	96	60	183	72
7.5	0.3	79	49	242	95
10	0.4	64	40	310	122
12.5	0.5	47	29	386	152
15	0.6	25	15	435	171
16	0.63	0	0		Break

Note: As designed, the Yagi element will break with a 16-mm (0.63 inch) radial ice thickness at zero wind load, or at lower values of ice loading when combined with the wind. The design was not optimized to resist ice loading. Optimized designs will use elements that are overall thicker, especially the tip elements.

tions is that the stresses in all critical points of the Yagi are kept at a similar level when loading. In other words, the mechanical design should be well-balanced, since the system will only be as strong as the weakest point.

Make sure you know exactly the rating of the materials you are using. The yield stress for various types of steel and especially stainless steel can vary with a factor of three! Do not go by assumptions. Make sure.

3.3.9. Element Finishing

As a final touch I always paint my Yagi beams with three layers of transparent metal varnish. It keeps the aluminum nice and shiny for a long time.

3.3.10. Ice Loading

Ice loading greatly reduces an antenna's wind-survival speed. Fortunately, heavy ice loading is not often accompanied by very high winds, with an exception for the harshest environments (near the poles).

Although we are almost never subject to ice loading here in Northern Belgium, it is interesting to evaluate what the performance of the Yagi would be under ice loading conditions. **Table 13-5** shows the maximum wind survival speed and element sag as a function of radial ice thickness. As the ice thickness increases, the sections that will first break are the tips. The reflector of our metric-design element will take up to 16 mm of radial ice before breaking. At that time the sag of the tips of the reflector element will have increased from 100 cm without ice to approximately 500 cm with ice. If the Yagi must be built with heavy ice loading in mind, you will have to start from heavier tubing at the tips. The Element Strength module will help you design an element meeting your requirements in only a few minutes.

K5GO informed the ham community that there now is a special paint that helps reducing ice loading. It is Wearlon Super F1, icephobic paint (**www.wearlon.com**). This paint has a high content of silicone. If ice starts to form, it slides off pretty quickly, according to K5GO, who's been using the paint for a number of years. He said, "The elements and the end pieces of the boom for my 40 meter Yagi are painted with this stuff and this antenna accumulates very little ice as compared to the rest of the Yagi antennas."

3.3.11. Material Fatigue

Many have observed that light elements (thin-wall, low wind-survival designs) will oscillate and flutter under mild wind conditions. Element tips can oscillate with an amplitude of well over 10 cm. Under such conditions a mechanical failure will be induced after a time. This failure mechanism is referred to as *material fatigue*.

Element vibrations can be prevented by designing elements consisting of strong heavy-wall sections. Avoid tip sections that are too light. Tip sections of a diameter of less than about 15 mm are not recommended, although they may be difficult to avoid with a large 40-meter Yagi. Through the entire length of each element I run an 8-mm nylon rope, which lies loosely inside the element. This rope dampens any self-oscillation that might start in the element.

At both ends, the rope is fastened at the element tips by injecting a good dose of silicone rubber into the tip of the element and onto the end of the rope. The tip is then covered with a heat-shrinkable plastic cable-head cover with internal hot-melt glue. And at both ends of the element you must drill a small hole (3 mm) at the underside of the element about 5 cm from the tip of the element to drain out any condensation water that may accumulate inside the element.

Make sure the rope lays loosely inside the element. The method is very effective, and not a single case of fatigue element failure has occurred when these guidelines were followed. A simple test consists of trying to hand excite the elements into a vibration mode. Without internal rope this can usually be done quite easily. You will become really frustrated trying to get into an oscillation mode when the rope is present. Try for yourself!

3.3.12. Matching the Yagi

The only thing left to do now is to design a system to match the antenna feed-point impedance (28 Ω to the feed-line impedance (50 Ω). The choice of the omega match is obvious:

- No need for a split element (mechanical complications).
- No need to adjust the length of a gamma rod.
- Fully adjustable from the center of the antenna.

Only a true "plumber's delight" construction, with no floating or insulated elements, can guarantee proper operation as a top loading device (eg for 160 meter) on top of a tower. With floating elements the insulation may flash over and be destroyed when the Yagi is near or at the top of the tower, and cause destruction of the beam and erratic functioning as a loading device on 160 meters.

The two Omega-match capacitors are mounted in a housing made of a 50 cm long piece of plastic drainpipe (15 cm OD), which is mounted below the boom near the driven element (Fig 13-13). This is a very flexible way of constructing boxes for housing Gamma and Omega capacitors. The drain pipes are available in a range of diameters, and the length can be adjusted by cutting to the required length. End caps are available that make professional-looking and perfectly watertight units.

The design of the Omega match is described in detail in Section 3.10.2. **Fig 13-14** shows the SWR curve of my 40-meter Yagi. The 1.5:1 SWR bandwidth turned out to be approximately 250 kHz.

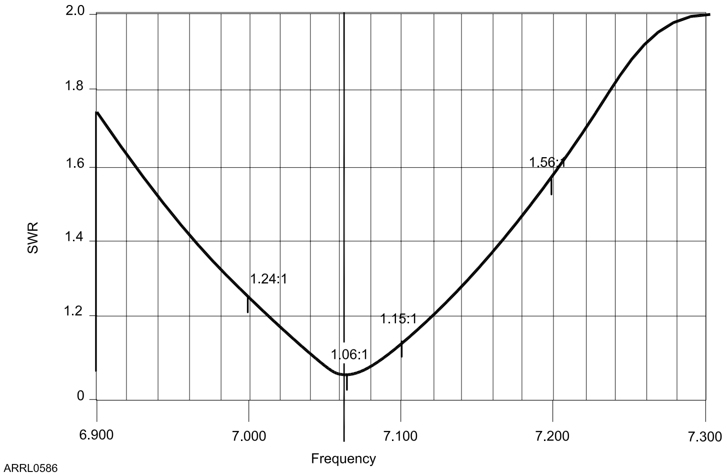


Fig 13-14 — SWR curve of the ON4UN 40-meter Yagi, as copied from a screen dump running the AIM 4170 antenna analyzer. The antenna was designed to cover 7.0 (SWR 1.24:1) to 7.2 MHz (SWR 1.56:1). The 2:1 SWR bandwidth is 300 kHz.

3.3.13. Tower, Mast, Mast Bearings, Drive Shaft and Rotator

If you want a long-lasting low-band Yagi system, you must pay attention to:

- The tower.
- The rotating mast.
- The mast bearings.
- The rotator.
- The drive shaft.

3.3.13.1. The Tower

Your tower supplier or manufacturer will want to know the wind area of your antenna. Or maybe you have a tower that's good for 2 meters^2 of top load. Will it be okay for the 40-meter antenna?

Specifying the wind area of a Yagi is an issue of great confusion. Wind-thrust force is generated by the wind hitting a surface exposed to that wind. The thrust is the product of the dynamic wind pressure multiplied by the exposed area, and with a so-called *drag coefficient*, which is related to the *shape* of the body exposed to the wind. The resistance to wind of a flat-shaped body is obviously different (higher!) than the resistance of a ball-shaped or tubular-shaped body.

This means that if we specify or calculate the wind area of a Yagi, we must always specify if this is the equivalent wind area for a flat plate (which really should be the standard) or if the area is simply meant as the sum of the projected areas of all the elements (or the boom, whichever has the largest projected area; see Section 3.3.4.1).

In the former case we must use a drag coefficient of 2.0 (according to the latest EIA/TIA-222-E standard) to calculate the wind load, while for an assembly of long and slender tubes a coefficient of 1.2 is applicable. This means that for a Yagi consisting only of tubular elements (Yagi elements and boom), the flat-plate wind area will be 66.6% lower (2.0/1.2) than the round-element wind area.

The 40-meter Yagi, excluding the boom-to-mast plate, the rotating mast and any match box, has a flat-plate equivalent wind area of 1.65 meters^2. As the projected area of the three elements is 2.5 times larger than the projected area of the boom, the addition of the boom-to-mast plate and the match box will not change the wind load, which for this Yagi is only determined by the area of the elements. The round-element equivalent wind area for the Yagi is 2.74 meters^2.

The wind thrust generated by this Yagi at a wind speed of 140 km/h is 302 kg. This is for 140 km/h winds, without any safety margins or modifiers. Consult the EIA/TIA-222-E standard or your local building authorities to obtain the correct figure you should use in your specific case.

Let me make clear again that the thrust of 302 kg is only generated with the elements broadside to the wind. If you put the boom into the wind, the loading on the tower will be limited to 90 kg. However, I would not advise using a tower that will take less than 300 kg of top load. Consider the margin between the boom in the wind and the elements in the wind as a safety margin.

3.3.13.2. The Rotating Mast

Leeson (Ref 964) covered the issue of masts very well. Again, what you use will probably be dictated in the first place by what you can find. In any case, make sure you calculate the mast.

My 3-element 40-meter beam sits on top of a 5-meter long stainless-steel mast, measuring 10 cm in diameter with a wall thickness of 10 mm. This mast is good for a wind load of 579 kg at the top. I calculated the maximum wind load as 302 kg. At the end of a 5-meter cantilever the bending moment caused by the beam is 1,510 kgm. Knowing the yield strength of the tube, we can calculate the minimum required dimensions for our mast using Eq 13-2.

$$M_{max} = YS \times \pi \times \frac{10^4 - 8^4}{32 \times 10} = YS \times 58$$

where YS = yield strength. The stainless-steel tube I used has a yield strength of 50 kg/mm2.

M_{max} = 5000 × 58 = 290,000 kg-cm (kilogram-centimeter)
= 2900 kgm

It appears that we have a safety factor of 75% versus the moment created by the Yagi (1510 kgm). I have not included the wind load of the mast itself, but the safety margin is more than enough to cover the bending moment caused by the mast.

In my installation I welded the boom to mast plates on the mast at the heights where the beam needs to be mounted. These plates are exact replicas of the stainless-steel plates mounted on the booms of the Yagis (the boom-to-mast coupling plates). When mounting the Yagi on the mast, you do not have to fool around with U bolts; the two plates are bolted together at the four corners with 18-mm-OD stainless-steel bolts.

One word of caution about stainless-steel hardware. Do not tighten stainless-steel bolts as you would do with steel bolts. Stainless-steel bolts gall when over tightened and are very difficult to remove later. It is always wise to use a special grease before assembling stainless-steel hardware. Also, where safety is a concern, use one normal bolt, doubled up with a special safety self-locking bolt (with plastic insert). Between the two plates a number of stairs have been welded in order to provide a convenient working situation when installing the antennas.

3.3.13.3. The Mast Bearings

The mast bearings are important parts of the antenna setup. Each tower with a rotating mast should use two types of bearings:

- The thrust bearing — it should take axial weight as well as a radial load.
- The second bearing should only take a radial load.

The thrust bearing should be capable of safely bearing the weight of the mast and all the antennas. The thrust-bearing assembly must be waterproof and have provisions for periodic lubrication. **Fig 13-15** shows the thrust collar welded on the stainless-steel mast inside my top tower section. Notice the stainless-steel housing of the thrust bearing. The bearing is a 120-mm ID, FAG model FAG30224A (T4FB120 according to DIN ISO 355). In my tower the thrust bearing is 2 meters below the top of the tower.

My tower's second bearing is mounted right at the top of the tower and consists of a simple 10-cm long nylon bushing with approximately 1-mm clearance with the mast OD. Note that the thrust bearing could instead be at the top, with the radial bearing at the lower point. The choice is dictated by practical construction aspects.

The mast and antenna weight should never be carried by the rotator. In my towers I have the rotator sitting at ground level, with a long drive shaft in the center of the self-supporting tower. The drive shaft is supported by the thrust bearing near the top of the tower. The fact that the heavy drive shaft hangs in the center of the tower adds to the stability of the tower. I can easily replace the rotator. The coupling between the rotator and the drive shaft is a cardan axle from a heavy truck, as shown in Fig 13-15.

3.3.13.4. The Rotator

I would not dare to suggest using one of the commercially available rotators with antennas of this size. Use a prop-pitch or a large industrial-type worm-gear reduction with the appropriate reduction ratio and motor. For example, the Prosistel "Big Boy" rotator is available from Array Solutions at **www.arraysolutions.com**.

3.3.13.5. The Drive Shaft

The drive shaft is the tube connecting the rotating mast with the rotator. The drive shaft must meet the following specifications:

- It must act as a torque absorber when starting and stopping the motor. This effect can be witnessed when you start the rotator and the antenna actually starts moving a second later. This relieves a lot of stress on the rotator. Leeson (Ref 964) uses an automotive transmission damper as a torque absorber spring.
- The drive shaft should not have too much spring effect, so that the antenna points in the right direction even in high winds. If there is too much springiness, excessive swinging

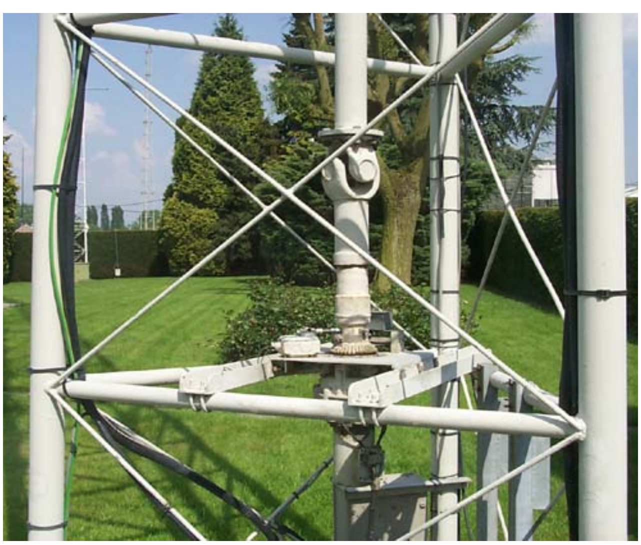


Fig 13-15 — Top: The thrust bearing for the 100-mm OD mast inside the top section of the 24-meter tower at ON4UN. Bottom: base of the self-supporting 25-meter tower (measuring 1.5 meters across), with the prop-pitch motor installed 1 meter above the ground. The drive shaft is coupled to the prop-pitch motor via a cardan axle from a heavy truck. Having the motor at ground level facilitates service, and takes torque load off the tower. In addition, the long drive shaft acts as a shock (momentum) absorber, greatly reducing strain on the motor.

of the antenna could damage the antenna. The acceleration and the forces induced by the swinging of the elements could induce failure at the element-to-boom mounts.

The torque moment will deform (twist) the drive shaft (hollow tube). The angle over which the shaft is twisted is directly proportional to the length of the shaft. In practice, we should not allow for more than ±30° of rotation under the worst torque moment.

In an ideal world, the Yagi is *torque-balanced*, which means that even under high wind load there is no mast torque. In practice nothing is less true: Wind *turbulence* is the reason that the large wind capture area of the Yagi always creates a large amount of momentary torque moment during wind storms.

When rotation is initiated, the inertia of the Yagi induces twist in the drive shaft. The same is true after stopping the rotator, when the antenna overshoots a certain degree before coming back to its stop position.

In practice you will have to make a judicious choice between the length of the drive shaft and the size of the shaft. Using a long drive shaft and the rotator at ground level has the following advantages (in a non-crank-up, self-supporting tower):

- No torque induced on the tower above the point where the rotator is installed.
- Motor at ground level facilitates maintenance and supervision.
- Long crank shaft works as torsion spring and takes torque load off the motor.
- The disadvantage is that you will need a sizable shaft to keep the swinging under control.

3.3.13.6. Calculating the Drive Shaft

It is difficult, if not impossible, to calculate the torque moment caused by turbulent winds. I have estimated the momentary maximum torque moment to be three times as high as the torque moment on one side of the boom, as calculated before for a wind speed of 140 km/h. This is 360 kgm. This means that the wind turbulence momentarily causes the antenna to rotate in only one direction, and that we disregard the forces trying to rotate the antenna in the opposite direction. In addition, I added a 200% safety factor. I use this figure as the maximum momentary torque moment to calculate the requirements for the drive shaft. I have not found any better approach yet, and it is my practical experience that this approach is a fair approximation of what can happen under the worst circumstances with peak winds in a highly turbulent environment.

Assumed momentary maximum torque moment T = 360,000 kgmm (kilogram-millimeter), calculate the section shear modulus (Z):

$$Z = \pi \times \frac{D^4 - d^4}{16 \times D}$$

Assume the following:

D = 8 cm
d = 6.5 cm
Z = 56.7 cm^3

Calculate the shear stress (ST):

$$ST = \frac{T}{Z}$$

where

Z = modulus of section under shear stress
T = applied torque moment

$$ST = \frac{36{,}000 \text{ kgcm}}{56.7 \text{ cm}^3} = 635 \text{ kg / cm}^2 = 6.35 \text{ kg / mm}^2$$

This is a low figure, meaning the tube will certainly not break under the torque moment of 36,000 kgcm. Calculate the maximum twist angle (TW). The twist angle of the shaft is directly proportional to the shaft length. In my case the rotator is 21 meters below the lower bearing, which makes the shaft 21 meters long. The critical part of the whole setup is the shaft-twist angle under maximum mast torque, where:

T = applied torque (360,000 kgmm)
L = length of shaft (21,000 mm)
G = rigidity modulus of the material = 8000 kg/mm^2
J = section modulus × radius of tube = 56,700 mm^3 × 40 mm = 2,268,000 mm^4
TW = 0.44 radians = 25°

A twist angle of 25° is an acceptable figure. The twist should in all cases be kept below 30° to keep the antenna from excessively swinging back and forth in high winds. It is clear that the same result could be obtained with a much lighter tube, provided it was much shorter in length.

3.3.14. Raising the Antenna

A 3-element full-size 40-meter Yagi, built according to the guidelines outlined in the previous paragraphs, is a "little monster." Including the massive boom-to-mast plate, it weighs nearly 250 kg (500 lbs). A few years ago I befriended a man who has his own crane company. He has a whole fleet of

Fig 13-16 — The ON4UN 40-meter Yagi is lowered on top of the rotating mast at a height of 30 meters using a 48-meter hydraulic crane.

hydraulic cranes that come in very handy for mounting large antennas on towers.

Fig 13-16 shows the crane arm extended to a full 48 meters, maneuvering the 40-meter Yagi on top of the 30-meter self-supporting tower. With the type of boom-to-mast plates shown in Fig 13-13, it takes but a few minutes to insert the four large bolts in the holes at the four corners of the plates and get the Yagi firmly mounted on the mast.

3.3.15. Conclusion

Long-lasting, full-size low-band Yagis are certainly not the result of improvisation. The 40-meter Yagi I've described here has been up for many years now, without any repairs. Long lasting Yagis, especially for the low bands are the result of a serious design effort, which is 90% a mechanical engineering effort. Software is now available that will help design mechanically sound, large low-band Yagis. This makes it possible to build a reliable antenna system that will out-perform anything that is commercially available by a large margin. It also brings the joy of home-building back into our hobby, the joy and pride of having a no-compromise piece of equipment.

3.4. A Super-Performance, Super-Lightweight 3-Element 40-Meter Yagi

Nathan Miller, NW3Z, designed a very novel and attractive 3-element Yagi that was featured in *QST* (Ref 979). See **Fig 13-17**. It weighs only a tiny fraction of the battleship described in Section 3.3. The NW3Z antenna can be turned with a run-of-the-mill good-quality rotator. The antenna is based on a similar 2-element design by Jim Breakall, WA3FET.

This 3-element Yagi also uses the principle of instantaneous pattern reversal, which I described in a previous edition of this book (see also Section 3.5). Basically this Yagi uses two directors, symmetrically located with respect to the driven element. By using small relays an inductance is inserted in the middle of the parasitic element to turn it into a reflector to make instantaneous direction switching possible. You must have experienced this feature in order to fully appreciate it!

3.4.1. Electrical Performance

I modeled the antenna using *EZNEC*, both in free space as well as over real ground. The dimensions shown in **Fig 13-18** are very close to those published by NW3Z. In free space the antenna exhibits 7.34 dBi gain, with a feed impedance of 40.5 Ω, using a loading inductance with a reactance of 138 Ω as shown in the *QST* article. The gain is more than 7.1 dBi, across the whole 40-meter band. The F/B performance in free space is illustrated in **Fig 13-19**.

This is a fairly low-Q antenna, yielding a feed-point impedance of nearly 50 Ω, which means that the antenna is split fed with a current balun. The computed SWR values are shown in **Table 13-6**.

I also modeled the antenna over real ground, at a height of 21 meters. We learned in Chapter 5 that for most DX paths an elevation angle between 10° and 15° seems to be optimum. If the angle is 15°, a Yagi at 0.6 λ will lose about 1.5 dB compared to its brother at 1 λ, but it will have much better high-angle rejection. The high antenna rejects a signal at a wave angle

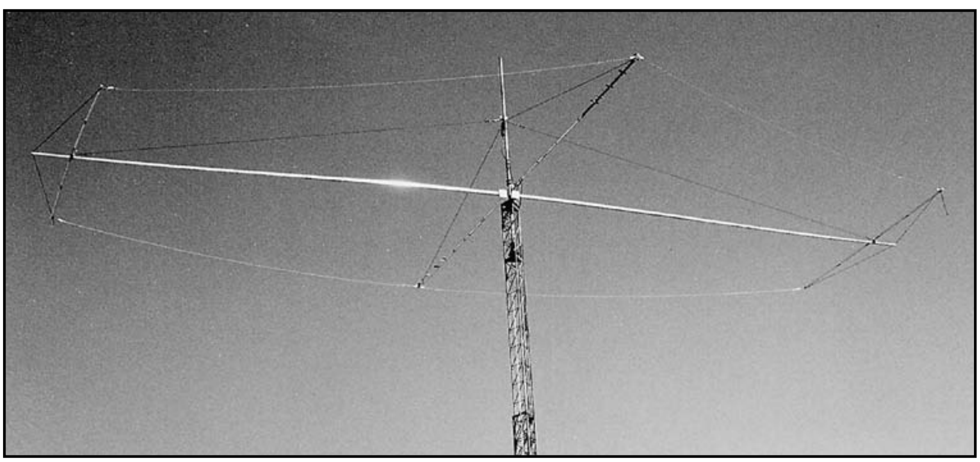


Fig 13-17 — The 3-element 40-meter NW3Z Yagi is mounted on a 21-meter crank-up tower at the Penn State University Dept of Electrical Engineering research facility at Rock Springs.

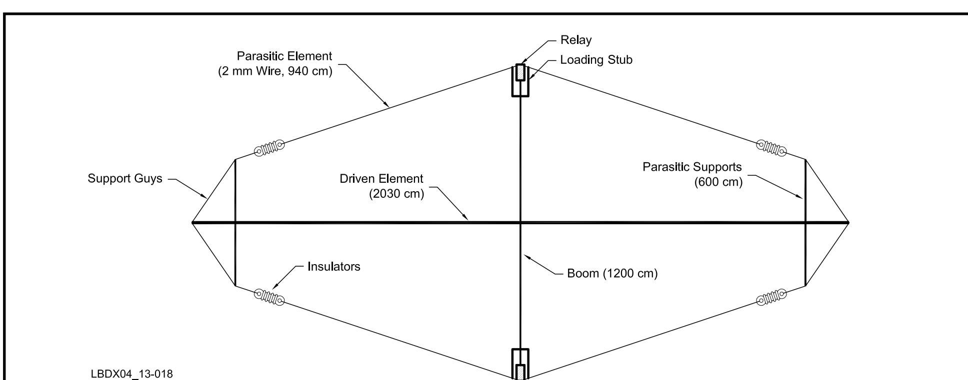


Fig 13-18 — Layout of the NW3Z/WA3FET lightweight 3-element 40-meter Yagi.

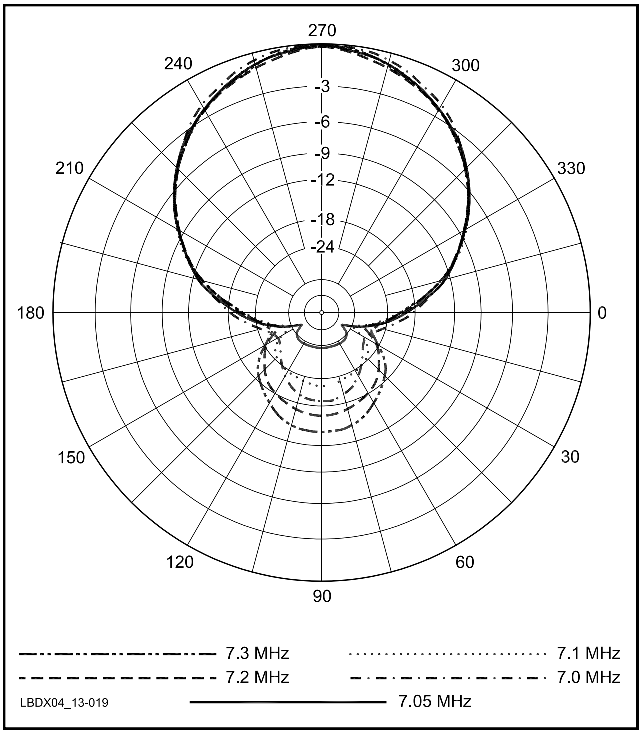


Fig 13-19 — Horizontal radiation pattern in free space for the NW3Z/WA3FET 40-meter Yagi. The F/B is 20 dB or better from 7.0 to 7.1 MHz, and still a usable 17 dB at 7.2 MHz.

Table 13-6

SWR Performance of the WA3FET/NW3Z Yagi Modeled in Free Space, Using X_L = 138 Ω.

7.0 MHz	*7.05 MHz*	*7.1 MHz*	*7.2 MHz*	*7.3 MHz*
1.4:1	1.2:1	1.1:1	1.3:1	1.7:1

Table 13-7

SWR Performance of the WA3FET/NW3Z Yagi at 21 Meters Over Average Ground, Using X_L = 132 Ω.

7.0 MHz	*7.05 MHz*	*7.1 MHz*	*7.2 MHz*	*7.3 MHz*
1.6:1	1.3:1	1.2:1	1.3:1	1.8:1

of 60° in the forward direction by about 8 dB. The antenna at 0.6 λ will reject the same signal, about 18 dB! Computer SWR values are shown in **Table 13-7**.

This antenna is within reach of many and it can perform quite outstandingly at a 0.5 to 0.6-λ height. **Fig 13-20** shows the radiation patterns for 7.05 MHz. Note that in order to obtain best F/B at that frequency, the reactance of the loading coil should be changed from 138 Ω to 130 Ω. The SWR curve becomes a little steeper, especially on the low side.

3.4.2. Mechanical Design

The prototype was made using two types of aluminum tubing: The 2.5 and 2.25-inch-OD tubing is an extruded 6061-T6 alloy with 0.125-inch walls. All other tubing is 6063-T832 with 0.058-inch wall. The parasitic elements are made of #10 aluminum plated steel wire. Copper-clad steel wire or bronze wire would also be appropriate.

The boom, for which the taper schedule is shown in **Fig 13-21** weights only 9 kg. The entire Yagi weighs well under 50 kg, which makes this a really super lightweight 3-element full-size 40-meter antenna!

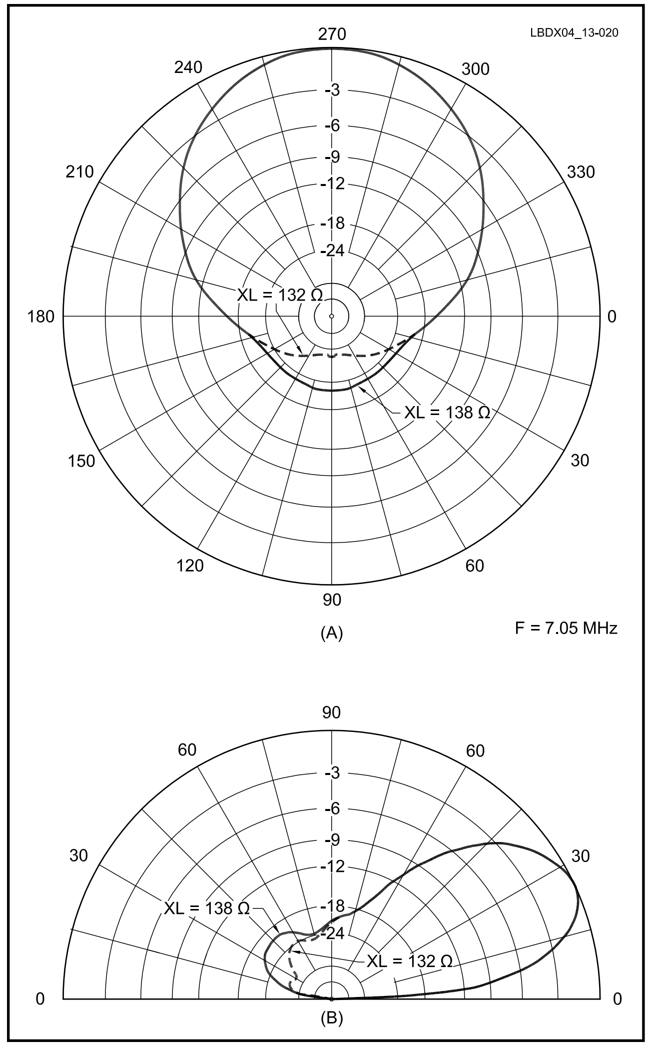


Fig 13-20 — At A, horizontal radiation pattern at 21-meter height over average ground for the NW3Z/WA3FET 40-meter Yagi. The patterns are for 7.05 MHz. Reducing the value of the loading reactance from 138 to 132 Ω improves the F/B performance.

3.4.3. The Parasitic Element Supports

Miller used an aluminum spreader (1.5-inch OD) tubing broken up with fiberglass rods to minimize loading of the director. Also, where the element supports are attached to the driven element, he uses a 30-cm long fiberglass rod, to keep the metal of the support far enough from the driven element. A valid alternative would of course be to use fiberglass poles along the entire length.

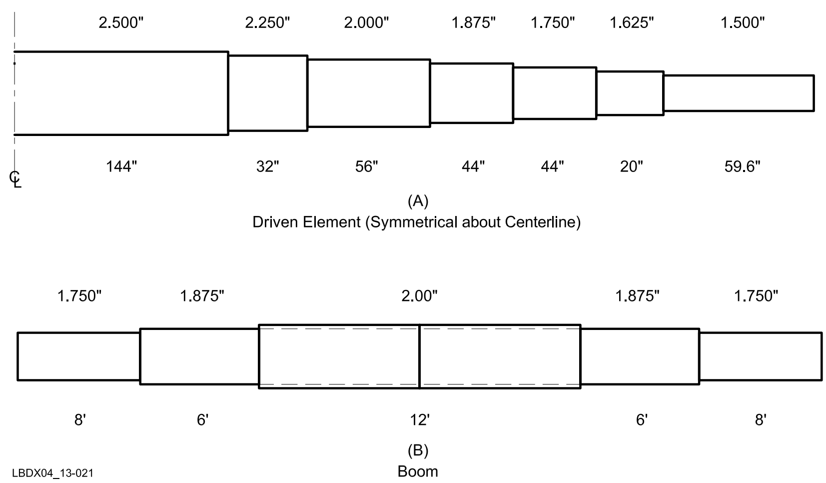


Fig 13-21 — Taper schedule for the driven element (A) and boom (B) of the NW3Z/WA3FET Yagi. The driven-element taper schedule is quite different from an ordinary Yagi element taper, since the element supports the 6-meter (20-foot) long cross bars at the end, which in turn support the ends of the parasitic wires. In this design the driven element is more like a boom, while the boom can be much lighter because it only supports the centers of the parasitic wires. The driven element weighs about 22 kg.

3.4.4. Truss Wiring

Because of the additional cross-arm and parasitic-wire weight loading on the tips, the full-size driven element requires a supporting truss. The antenna uses Phillystran (PVC coated Kevlar rope) for this purpose. The boom is also guyed. Both sets of guy wires are attached to a support about 2 meters above the antenna. Horizontal support guys are used from the driven-element tip to the ends of the parasitic supports to counter the tension in the parasitic wires, as shown in **Fig 13-22**.

3.4.5. Tuning the Yagi

The shorted-stub loading reactance of 132 to 139 Ω represents an inductance of 3 to 3.1 μH. You can achieve an unloaded Q of more than 500 with a well-designed coil compared to a Q of about 100 with a linear-loading stub. So I designed a high-Q coil for this application:

Required inductance: 3.1 μH
Coil diameter: 7.5 cm (3 inches)
Coil length: 11.3 cm (4.5 inches)
Number of turns: 8 air-wound
Conductor: 6 mm (¼ inch) copper tubing

To tune the coil to the required exact value, just stretch or squeeze the turns. For a constant diameter (7.5 cm) and for 8 turns the inductance will vary from 2.8 μH to 3.8 μH by changing the coil length from 11.3 to 15 cm.

The *NEC* model shows that the director is resonant on 6.7 MHz, and the reflector on 7.0 MHz by themselves. The best way to make sure that the parasitic elements are resonant on 6.7 MHz would be to feed the elements temporarily with λ/2 feed lines and cut them for zero reactance using a network analyzer or an antenna analyzer. While doing this the driven element and the second parasitic element must be left "open." Adjusting the loading coil or stub can be done the same way: Connect the feed line in series (not in parallel!) with the loading coil.

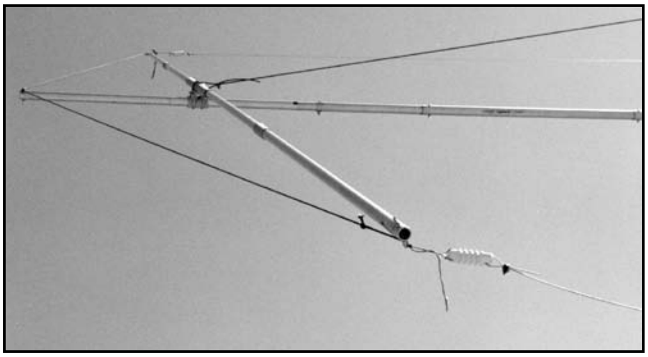


Fig 13-22 — The parasitic-wire, cross support, parasitic elements, horizontal and vertical supports in place on the NW3Z/WA3FET Yagi.

3.4.6. Feeding the Yagi

The original design uses a very simple split-element feed, since the antenna impedance is around 40 Ω. This requires a split driven element. A fiberglass rod used as an element insert can be used for the purpose. When direct feed is used, a choke balun is required. Alternatively, you could use any of the other matching systems described in Section 3.8.

3.4.7. Conclusion

Considering that the NW3Z antenna only trades 0.2 dB of forward gain vs my heavy-weight 3-element Yagi, and given its additional feature of instant direction reversal, this antenna is one of the most interesting designs that has been published for a long time, and deserves great popularity. When will we see the first 80-meter version of this design?

3.5. Design Considerations for a 3-Element Full-Size 80-Meter Yagi

In this section we will do a design of a 3-element full-size 80-meter Yagi, and point out all the details that require our attention.

One of the disadvantages of a large rotatable Yagi is the fact that switching directions takes a while. Very large and heavy Yagi antennas should not be rotated at speeds of more than 0.5 to 1 rev/min maximum. In order to overcome this problem, it is possible to design a Yagi where, by means of relays, the director is instantly transformed into a reflector, and vice versa. This means that at 1 rev/min it would never take

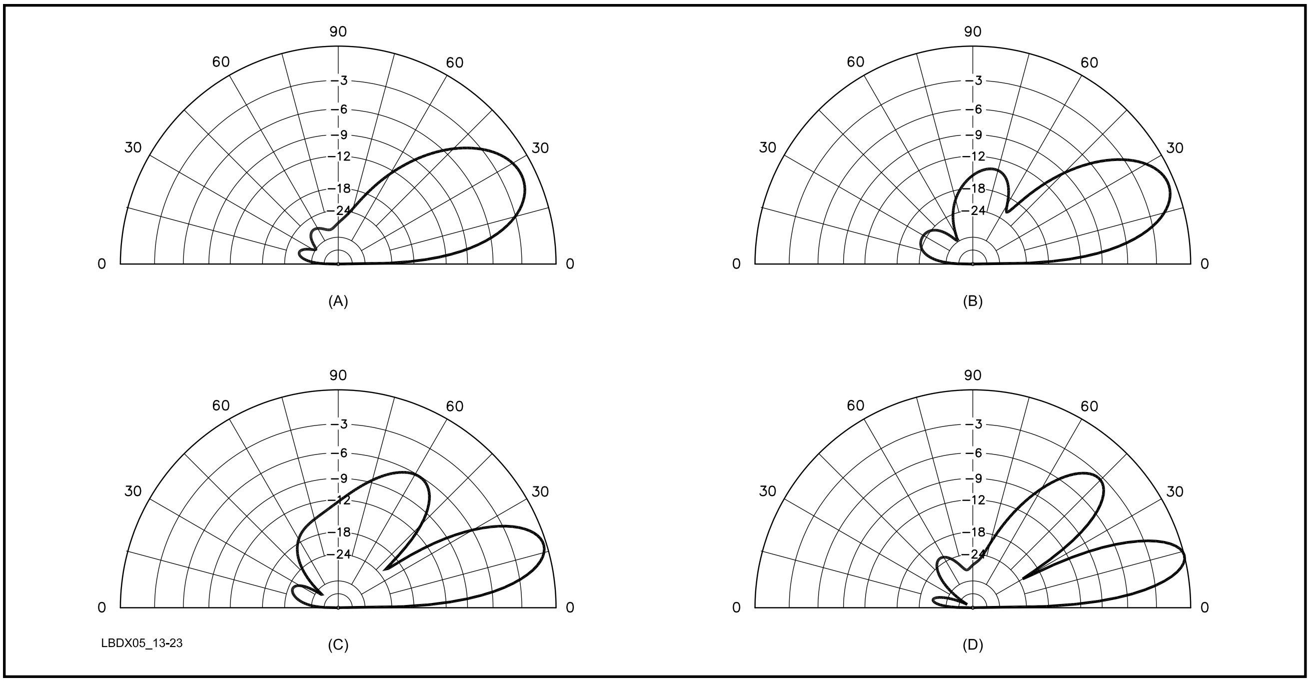


Fig 13-23 — Vertical radiation pattern of a 3-element Yagi at various heights. It is clear that the 0.5 λ height is by far the most suitable height for general DXing on 80 meters. The high-angle secondary lobes and the narrow first lobe plus the minimum (dip) between the first and the second lobe make higher heights a bad choice for 80 meters, where the bulk of DX signals come in at angles between 25 and 50°. The patterns shown are generated for flat ground with good ground conductivity. A: 0.5 λ height, B: 0.6 λ height, C: 0.8 λ height and D: 1.0 λ height.

more than 15 seconds to point the antenna in any direction. On average, it would take 7.5 seconds.

Also, due to its high relative bandwidth (8.2% as compared to 2.4% for the 20-meter band), it is impossible to design a Yagi that will exhibit good gain, good F/B and an acceptable SWR at the high end (3.8 MHz) as well as the low end (3.5 MHz) of the band, without resorting to our bag of special tricks.

3.5.1. Antenna Height for an 80-Meter Yagi

Fig 13-23 shows the radiation patterns for a "standard" 3-element Yagi at heights ranging from λ/2 to 1 λ. Above λ/2, an annoying high angle lobe appears, and a lot of RF is wasted at that angle. At λ/2 height, the radiation angle is approximately 25° to 30° (depending on the ground quality), with a reasonably broad lobe (29° at –3 dB).

Chapter 5, Section 1.1.1.2 showed us that wave angles as low as 10° are not unusual, and that the bulk of DX happens at angles between 10 and 20°. This means that with horizontally polarized antennas you can't really get high enough: 40 meters (λ/2) seems to be a bare minimum (radiation angle approximately 25°), but I know of 80-meter Yagis at 30 meters that perform very well also.

If you are tempted to put the Yagi much higher, eg, at 1 λ (78 meters), the main lobe is as low as 14°, which is really too low for most cases. In addition, the lobe will be quite narrow (only 14° at –3 dB) and you have a null at 30°, which happens to be the angle where you will have a lot of DX coming in. The second lobe is at 45°, which in turn is already too high for serious DX work. I know very high antennas are like a status symbol, but this time too high is no good! It is true that at 1-λ height the Yagi exhibits 1.0 dB more gain than at λ/2, but what's the point of concentrating more energy at the wrong elevation angle?

The 3-element full-size Yagi design that I will cover in this Section is developed to be installed at a height of λ/2 (38 meters). It is obvious that the considerations outlined above also apply for loaded (coil or linear loading) Yagis described in Section 3.6.

3.5.2. Electrical Design

The Yagi has been developed to be physically fully symmetrical. This means that the driven element is right at the center of the boom, with two parasitic elements of equal physical length. Both parasitic elements are the director length at the highest operating frequency. The reflector is then loaded in the center (by an inductance) in order to lower its resonant frequency. This means that both parasitic elements need to be split at the boom. By a set of relays it is possible to either short the split (turn the element into a director) or insert the required inductance (turn it into a reflector). This is the approach that also was adopted by WA3FET and NW3Z for designing their super-lightweight 40-meter Yagi (Section 3.4).

I also set out to design a Yagi which should be switchable from the SSB to the CW portion of 80 meters without any compromise in performance (gain, F/B).

The constant-element-diameter design, shown in **Fig 13-24A** (3.5 MHz data), uses a constant diameter of 100 mm (4 inches) which (later) turned out to be the equivalent diameter of the tapering diameter element of our mechani-

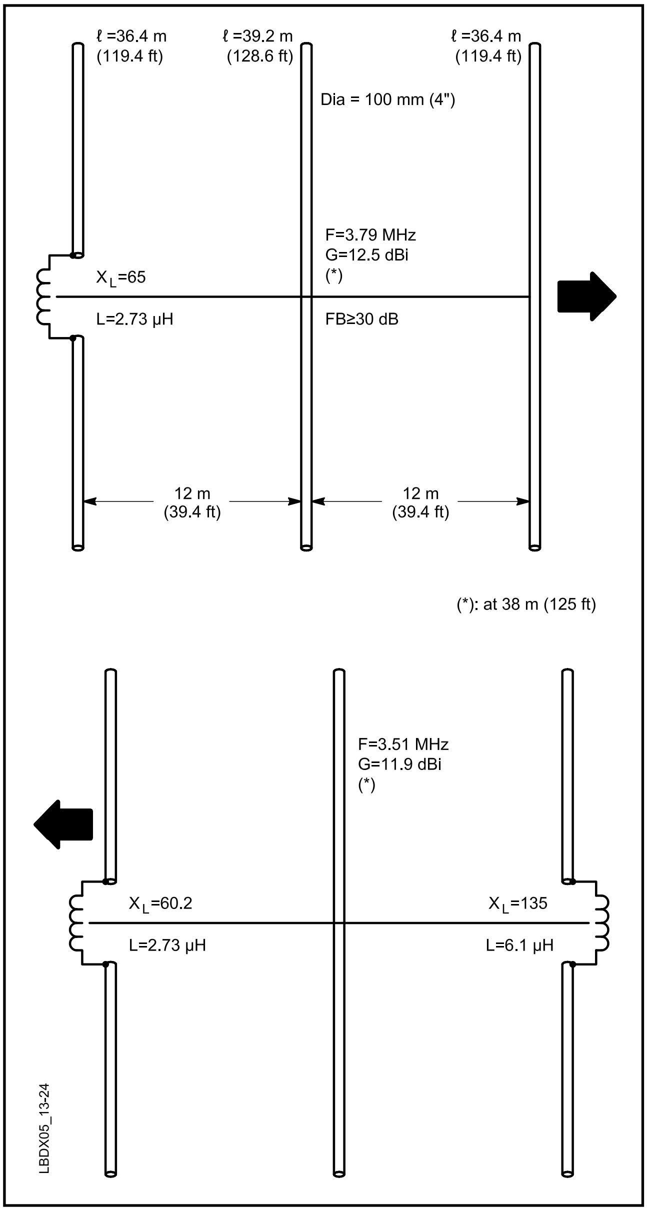


Fig 13-24 — Design of the equally spaced 3-element 80-meter Yagi. The element lengths shown are for a constant element diameter of 100 mm. The loading coils make this an excellent Yagi for both the phone end of the band as well as the CW end of the band. Note that the same coil (2.73 µH) is used as a loading element for the reflector on 3.8 MHz and for the director on 3.5 MHz.

cal design (see Section 3.5.8). Using this element diameter and inserting a coil with X_L= 65 Ω in the reflector turns this into a Yagi with a very good gain and F/B ratio. Note that with a smaller element diameter the Q factor of the element would be higher, which in turn means that more inductance would be required to tune the element to the same frequency. For a constant diameter of 22.225 mm (⅞ inch), the required reactance would be 85 Ω.

To make the same Yagi also work on 3.5 MHz, all that is required is a coil in the director element, and a second (larger) coil in the reflector. It turns out that on 80 meters, an element length that makes a perfect reflector for 3.8 MHz is a perfect director on 3.5 MHz. In other words, the same coil that is used for loading the reflector on 3.8 MHz can be used as a loading coil for the director on 3.5 MHz.

In our example, the coil that has a reactance of 65 Ω on 3.79 MHz (2.73 µH) has a reactance of (65 × 3.51/3.79) = 60.2 Ω on 3.51 MHz. Together with a loading coil having a reactance of + *j*135 Ω at 3.51 MHz (6.1 µH), this value results in a very good 3-element Yagi for the CW end of the 80-meter band. If the antenna is erected at a height of λ/2, the F/B ratio is between 25 and 30 dB at any wave angle between 0 and 90°, at both design frequencies (3.79 and 3.51 MHz).

The exact length of the driven element does not influence the directivity or gain of the antenna, but it is important when it comes to designing a matching system (see Section 3.10).

The design was modeled with *ELNEC*. Modeling and optimizing of the Yagi for best gain and F/B was done over real (good) ground at a height of λ/2. This is the ideal height for such an antenna.

Under these conditions the gain is calculated as 12.5 dBi at 3.79 MHz and 11.9 dBi at 3.51 MHz. The horizontal and vertical radiation patterns for the 3-element Yagi are shown in **Figs 13-25** and **13-26**.

3.5.3. Parasitic Parallel Capacitance with Split Elements

Split elements cannot be realized without introducing some parallel capacitance between the inside end of the half-element and the boom, or between the two element halves (in case you have no boom or have a dielectric boom). The ends of the insulated elements have a certain capacitance with the boom because of the mechanical construction of the insulating material and all the mounting hardware. If we were to use the loading coils as modeled above, without taking into account the "parasitic" capacitance, the loading effects could be way off.

The parasitic capacitance is the value of the series connection of the capacitances of each half element to the boom. In other words, the values shown in **Fig 13-27** are half the values as measured on one of the element legs. It is essential that this capacitance be measured. This can easily be done before the Yagi is raised. However, you cannot measure it on a finished element, because the self-capacitance (from one side to the other and also to ground) of the full element itself would upset the results.

I made a mockup of the center insulator consisting of the boom and the mounting hardware, but no element. Then I measured the capacitance at the Yagi operating frequency. The capacitance can range from just a few pF, if special care has been taken to reduce it, to several hundred pF.

Once the mechanical design of the element-to-boom mounting system is finished, a full-size model must be built and the capacitance measured. These capacitance values are then plugged into *ELNEC* to continue developing the mathematical model.

One should take the necessary measures to keep the value of the capacitance between each element half and the boom below 100 pF, to avoid problems. (This was a serious problem for building the 160 meter Yagi covered in Section 3.11.)

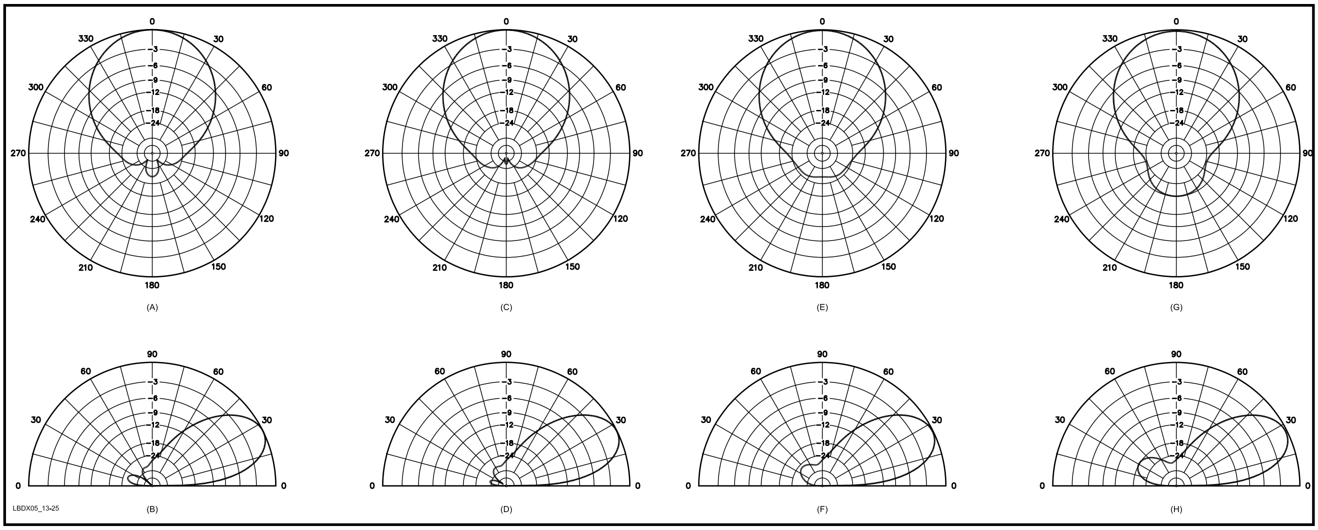


Fig 13-25 — Horizontal and vertical radiation patterns for the 3-element 80 meter Yagi on the SSB end of the band. These patterns are generated for a design frequency of 2.70 MHz. The azimuth patterns are generated for main vertical radiation angle of 27°. A and B: 3.8 MHz, C and D: 3.79 MHz, E and F: 3.775 MHz and G and H : 3.75 MHz.

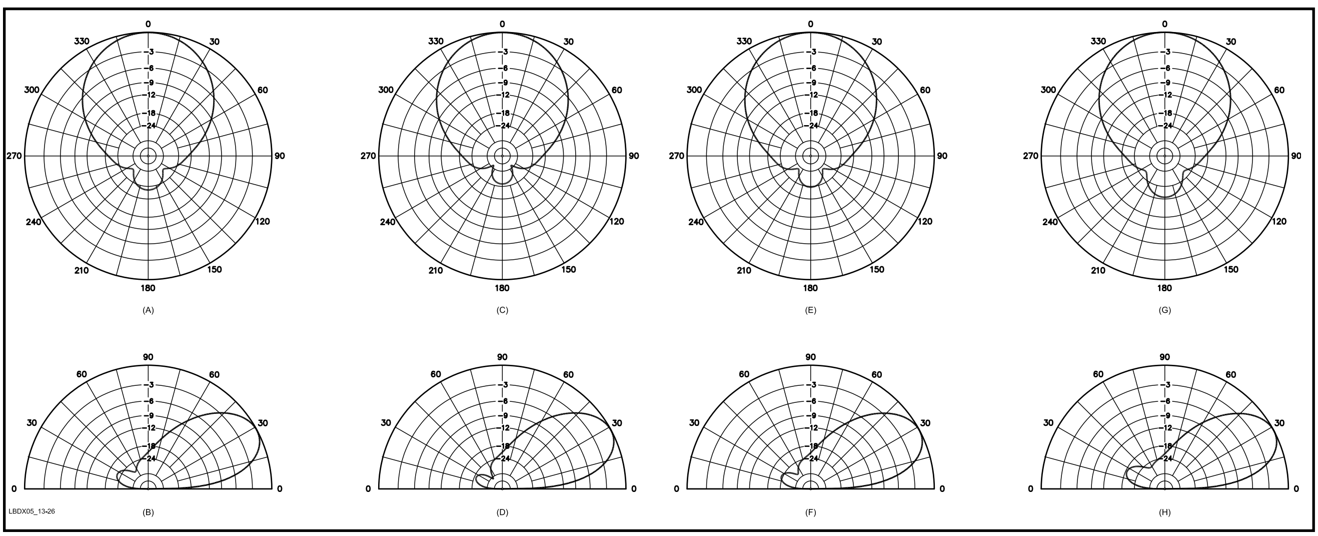


Fig 13-26 — Horizontal and vertical radiation patterns for the 3-element 80-meter Yagi on the CW end of the band. These patterns are generated for a design frequency of 2.70 MHz. The azimuth patterns are generated for main vertical radiation angle of 27°. A and B: 3.5 MHz, C and D: 3.51 MHz, E and F: 3.53 MHz and G and H: 3.55 MHz.

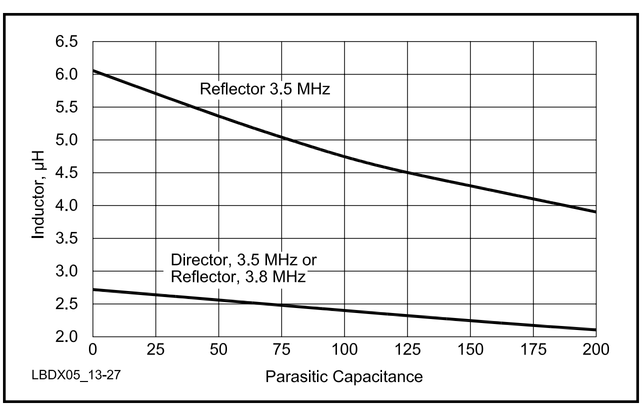


Fig 13-27 — Value of the tuning coil for the 80-meter Yagi as a function of parallel capacitance. Center-insulated elements always suffer from parasitic parallel capacitance (capacitance between the element and the boom). This means that the loading coils are in fact part of parallel circuits. The values must be adjusted in order to obtain the desired reactance. This chart shows the required reactance (in µH) as a function of total capacitance (which is equal to half of the capacitance between one element side and the boom). See text for details.

3.5.4. Modeling the Yagi Including the Parasitic Parallel Capacitance

Let us assume we have measured a capacitance of 32 pF across the split elements where the loading coils will be connected. We now must model the Yagi using a parallel tuned circuit as a loading element, instead of just an inductor. The parallel capacitor of the tuned circuit is 32 pF. We must find the required inductance to achieve the desired loading as modeled before in our simplified model without parallel capacitance.

The following design methodology was used.

- The Yagi was first modeled and optimized without taking into account the parasitic capacitance.
- When the model was optimized, the resonant frequency of the director and the reflector was determined. This can be easily done as follows.

  1) Delete all elements from the model, except the element whose resonant frequency we want to know.
  2) Keep the loading device (if any), and excite the center of the element. The loading device can be simply in series with the excitation.
  3) Change the resonant frequency until you find a feed-point impedance where the reactive part is zero (this is the definition of resonance). In our Yagi the director for the SSB design (F_{design}= 3.79 MHz) is resonant at 4.005 MHz; the reflector is resonant at 3.745 MHz. The CW design (F_{design} = 3.51 MHz) has a director that is resonant at 3.745 MHz, and a reflector that is resonant at 3.465 MHz.

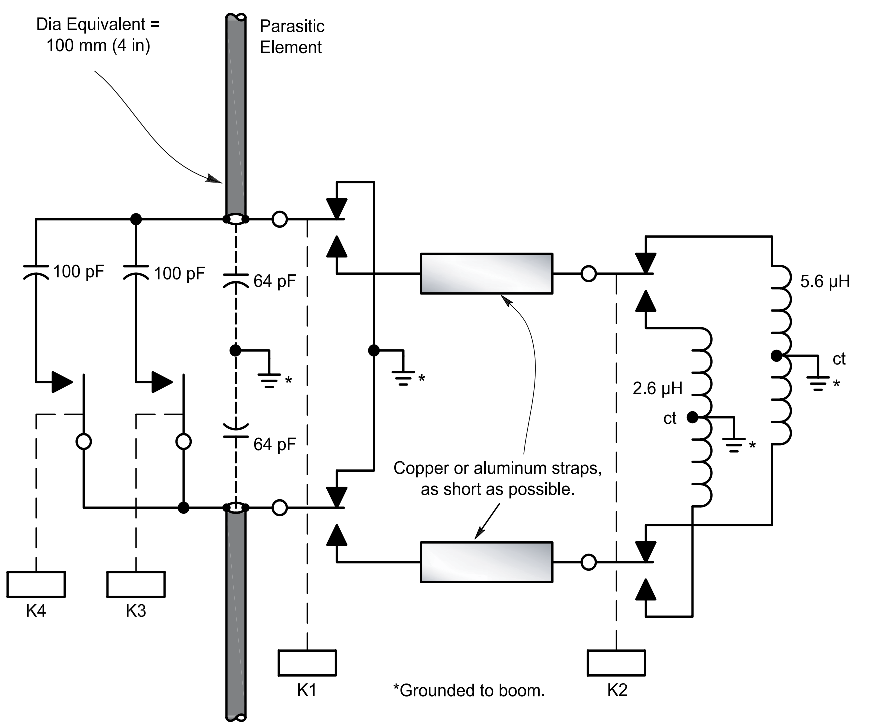
- Now the loading inductors are replaced in the modeling program by a parallel tuned circuit ($C_{parallel}$ = 32 pF), and the inductance values are found that produce the same resonant frequencies as found in our simplified (no parallel capacitance) model.

All of these steps can easily be done using the *EZNEC* modeling program.

The 3.745-MHz element turns out to require a loading inductance of 2.6 µH (in parallel with the 32 pF of parallel capacitance). This is + *j* 62 Ω at 3.79 MHz, or 57.3 Ω at 3.51 MHz. Compare these values with the values of 65 Ω and 60.2 Ω (L = 2.73 µH) as required when there is no parasitic parallel capacitance.

The 3.465-MHz CW-band reflector requires a loading coil of 5.6 µH (in parallel with 32 pF). This represents a reactance of + *j* 123 Ω at 3.51 MHz. Without the parallel capacitance the required loading inductance was 6.1 µH.

Fig 13-27 shows the adapted values of inductive reactance as a function of the parasitic capacitance. This chart is only valid for the Yagi with a given Q factor. In our design case, this is for a Yagi with an equivalent constant element diameter of 100 mm (4 inches). A Yagi with smaller diameter



	3.5 REF	3.5 DIR	3.8 DIR	3.8 REF	3.75 REF	3.70 REF
K1	X	X	0	X	X	X
K2	0	X	0	X	X	X
K3	0	0	0	0	X	X
K4	0	0	0	0	X	X

LBDX05_13-28

Fig 13-28 — Switching harness for the parasitic elements. To make a director at 3.8 MHz the two half elements are strapped. As a reflector on 3.8 MHz, a coil measuring 2.6 µH is inserted. When operating as a reflector element on 3.5 MHz the 5.6 µH coil is selected, while the director now selects a coil of 2.6 µH. The 64 pF shown is the parasitic capacitance between the elements and the boom. The 100 pF parallel capacitors are switched in parallel with the coil to enhance the F/B on 3.75 and 3.7 MHz (see text).

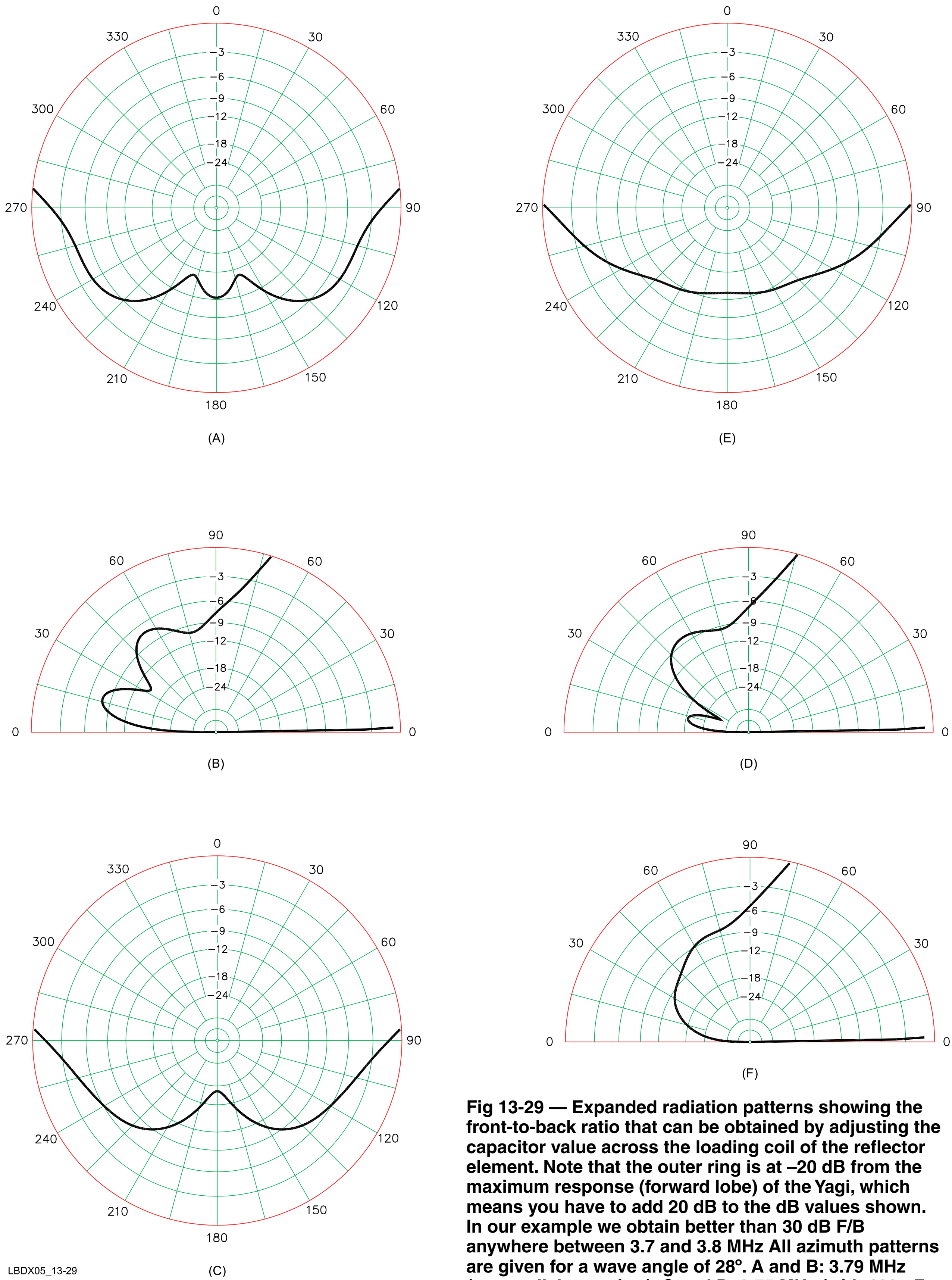


Fig 13-29 — Expanded radiation patterns showing the front-to-back ratio that can be obtained by adjusting the capacitor value across the loading coil of the reflector element. Note that the outer ring is at –20 dB from the maximum response (forward lobe) of the Yagi, which means you have to add 20 dB to the dB values shown. In our example we obtain better than 30 dB F/B anywhere between 3.7 and 3.8 MHz All azimuth patterns are given for a wave angle of 28°. A and B: 3.79 MHz (no parallel capacitor), C and D: 2.75 MHz (with 100 pF parallel), E and F: 3.7 MHz with 200 pF parallel.

elements will require more loading inductance and vice versa. A similar chart can be constructed easily for any element Q factor by modeling the combinations using software such as *ELNEC*.

3.5.5. The Loading Coils

The coil data for a high-Q 2.6-μH loading coil (reflector loading in the SSB band, director loading in the CW band) are:

Inductance: 2.6 μH
Coil diameter: 10 cm (4 inches)
Coil length: 6.9 cm (2.7 inches)
Total lead length: 15 cm (6 inches)
Turns: 5

For the 5.6-μH coil (reflector loading in CW band) the data are:

Inductance: 5.6 μH
Coil diameter: 10 cm (4 inches)
Coil length: 12.2 cm (4.8 inches)
Total lead length: 15 cm (6 inches)
Turns: 10

These coils are wound using 6 mm OD copper tubing and are air wound using no form (see Fig 13-75 later in this chapter).

Fig 13-28 shows the loading and switching layout, which is identical for both parasitic elements. The switching has two purposes, switching from SSB to CW band as well as instantaneous direction reversal.

3.5.6. Remote Tuning for Optimum F/B

The radiation patterns are shown in Figs 13-25 and 13-26. Note that the F/B deteriorates quite rapidly in the SSB band below 3.76 MHz.

We can tune the Yagi for a high F/B ratio over quite a wide spectrum by connecting a capacitor in parallel with the loading coil at the center of the reflector element. As we are on the "slope" of the parallel tuned circuit formed by the loading coil and the parallel capacitor, we just change the value of the impedance (which is a positive reactance) by changing the value of the parallel capacitor. In practice, this will not be needed on the CW band, where an excellent F/B is obtained from 3.5 to 3.53 MHz. In the phone band, however, adding a variable capacitor across the hairpin of the reflector will allow us to tune the Yagi for an F/B ratio of better than 25 dB at any frequency between 3.68 and 3.8 MHz!

Without the extra capacitance, the F/B is better than 22 dB from 3.76 to 3.8 MHz. With 100 pF in parallel, the F/B is better than 23 dB from 3.73 to 3.78 MHz, and with 200 pF in parallel, an F/B of better than 24 dB can be achieved between 3.69 and 3.73 MHz. These are worst-case F/B values over the entire 90° wave angle in the back of the Yagi. **Fig 13-29** shows the back patterns of the Yagi (on a very much stretched scale — outer ring equals –20 dB referenced to the maximum response) when tuned for maximum F/B using the variable capacitor across the reflector element.

In practice we can mount two transmitting-type 100-pF ceramic capacitors right at the center of the reflector element and switch these capacitors in parallel with the loading hairpin with vacuum relays. Fig 13-28 shows the switching and loading arrangement which must be provided at both parasitic elements.

3.5.7. Feeding the Yagi

Several methods are described in Section 3.10.3.

3.5.8. Mechanical Design of the Elements

Doing a fancy design on paper (on screen, really) is one thing. Doing the physical design, constructing it, and keeping it up in the air is another thing!

The half-element lengths for our theoretical model (see Fig 13-24), with a constant diameter of 100 mm are 18.2 meters for the director (reflector) and 19.6 meters for the driven element. In terms of wavelength (f = 3.79 MHz) these dimensions are: director/reflector: 0.46 λ, driven element: 0.4955 λ and spacing: 0.1517 λ.

The final element lengths, when using a tapering schedule, are much longer than for the constant reference diameter, and can be calculated using the Element Taper module of the *Yagi Design* software. Depending on the exact taper configuration, a full-size reflector will be approximately 42 meters long. There are two practical approaches for constructing elements that are that long:

- All tubular construction.
- Tubular tips and lattice construction for central section.

One of the successful constructions according to the first principle was done by OZ8BV (see Section 3.5.8.1). Using the Enter Your Own Design module of the *Yagi Design* software, the Element Taper module, and the Element Strength module, we can make the physical design of tubular elements.

3.5.8.1. The OZ8BV Design

In the early 1990s, Ben OZ8BV, first used an original KLM 3-element shortened Yagi, which was blown to pieces in the first gale wind at the Danish coast. Ben subsequently reinforced the design, using extensive side and top bracing, but he still was not fully happy with the electrical performance of the antenna.

This made him build his own 3 element full-size Yagi, according to the design outlined above, including the F/B tuning (see Section 3.5.6). OZ8BV used extensive guying and bracing

Table 13-8
Element Taper for 80-Meter Yagi

	Sec#1	*Sec#2*	*Sec#3*	*Sec#4*	*Sec#5*	*Sec#6*	*Sec#7*	*Sec#8*	*Sec#9*
OD (mm)	90	75	60	35	30	25	18	14	10
Wall (mm)	5	3	3	2	2	1.5	1.5	1	1
Length (cm)	782	160	395	74	115	105	98	65	100

Note: The taper was calculated using the Element Strength module of the ON4UN *Yagi Design* program. The element weight is approximately 80 kg. The element starting section 2 will survive wind speeds of 120 km/h without trussing. The center section (782 cm long) needs to be securely trussed.

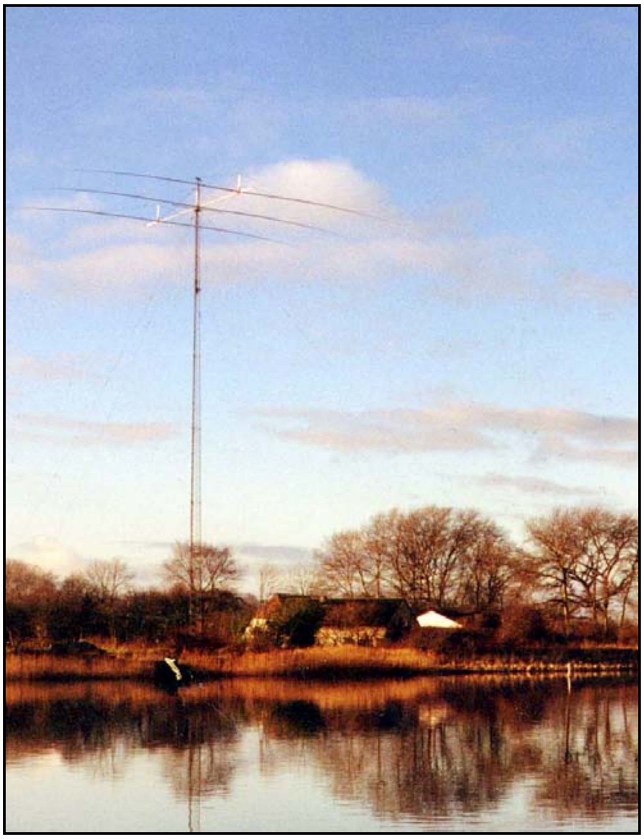


Fig 13-30 — OZ8BV's 3-element full-size 80-meter Yagi atop of his 48-meter tower on the Danish Baltic sea shore (in the 1990s)

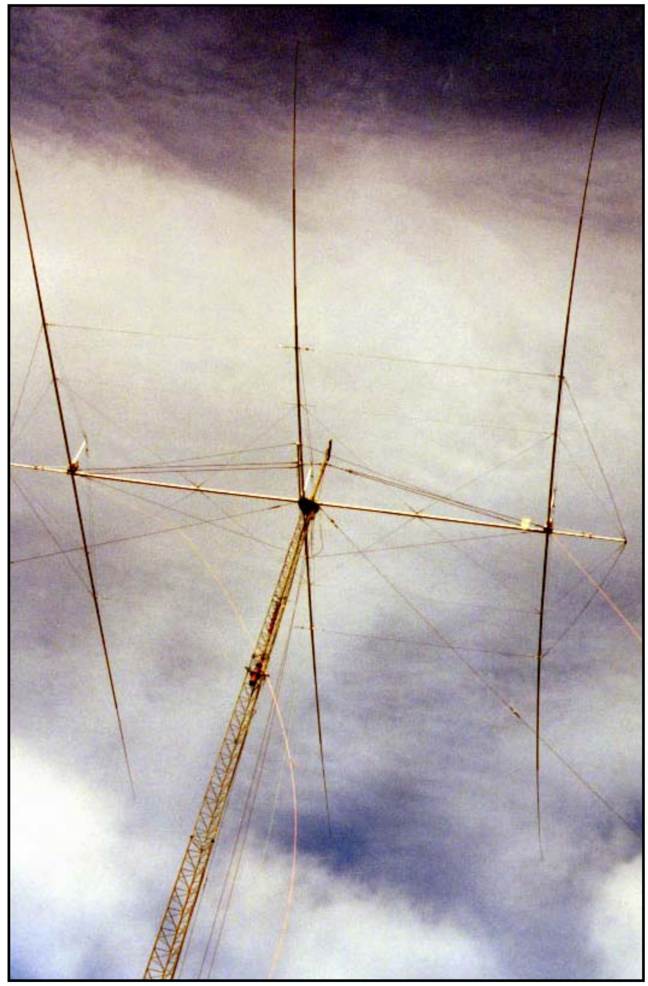


Fig 13-31 — This picture taken in 1998 clearly shows the rigging used by OZ8BV that kept his 3-element full-size 80-meter Yagi on the tower for over five years before it was taken down.

to achieve a 3 element full-size Yagi (**Fig 13-30**) that stayed up over five years without any damage, until it was taken down when Ben moved to another QTH.

Remote fine tuning of the F/B was achieved through the used of vacuum variables, fine tuning the loading coils (hair-pin coils) that resonate the parasitic elements, as explained in Section 3.5.6. Instantaneous direction reversal was also incorporated. This facility also proved to be very useful for measuring F/B. Using calibrated test equipment, Ben measured 30 dB F/B at any frequency in the 80-meter band

The antenna is made out of all aluminum tubular elements. The taper schedule is shown in **Table 13-8**. The elements are split, with a fiberglass bar (not rod), measuring 83 mm OD inserted in the center.

To keep the weight of the element within practical limits, and in order to obtain the required strength, it is necessary to guy the central part (the section with the OD of 90 mm) of the elements toward the boom.

In this construction you can calculate the inner section (the guyed section) using the approach as outlined by Leeson (Ref 964) for guyed booms. The unguyed section can be calculated using the ON4UN *Yagi Design* software.

The 38-meter-long elements are guyed 4 meters and 7.82 meters out, using 6 mm OD and 4 mm OD Phillystran cable. The element with this dimension is resonant on 4.035 MHz. The resonant frequency of one of the elements must be measured at the final operating height before calculating the required loading inductances. The elements must be brought to the required resonant frequencies using small high-Q coils at the center.

The boom is 18 meters long and has a diameter of 112.5 mm with 6 mm wall thickness. The boom extends beyond the attachment points of the parasitic elements using 2.7 meter long fiberglass tubes, measuring 60 mm OD, with 4 mm wall. These extensions are required to do the side-trussing of the elements, as is seen in **Fig 13-31.** The boom is also side trussed with 10 mm Phillystran, and vertically trussed with 15 mm Phillystran.

3.5.8.2. The DJ6JC Mechanical Design

Heinrich Lumpe, DJ6JC (SK), who had his own company (WIBI) making commercial radio towers and antenna systems, developed a mechanical design for a 3-element full-size Yagi using lattice construction for the boom and the major part of the elements. A similar technique was also used for the OH8X antenna described in Section 3.11.

The boom as well as the elements are made of a tapering rectangular lattice construction, measuring 60 cm in the center and 40 cm at the ends. Made of steel, the boom weighs no less than 1500 kg. The elements are made using aluminum. Here

Fig 13-32 — NO8D has a stack of two 3-element full-size 80-meter Yagis that use a lattice structure as boom. This approach makes it possible to make a strong boom that weighs much less than a boom made of large diameter heavy wall aluminum tubing.

too, a rectangular lattice construction tapers from 40 cm at the boom to just a few centimeters 16 meters out from the boom. At that point aluminum tubing (tapered diameter sections) make up for the remaining length (approximately 3.5 meters). The element does *not* require any guying, and exhibits a sag of only 30 cm over its entire length! The total element also weighs 114 kg. Compare that with the full size element of the 40 meter Yagi described in Section 3.3.3) which weighs about half as much. The total weight of the antenna is less than 1900 kg.

Whereas DJ6JC had planned to have this antenna up a long time ago, the local authorities have decided otherwise. It turned out that Heinrich, DJ6JC (now a Silent Key), would never see his masterwork up in the air. Too bad!

3.5.8.3. Conclusion of Full-Size 80-Meter Yagis

A project such as the construction of a full-size 3-element Yagi for 80 meters is not a simple task. Very few of the full-size 80-meter Yagis built so far have had a long life. Depending on what wind speed you want the "monster" to be able to survive, an 80-meter Yagi weighs between 700 and 2000 kg. The material cost is substantial, not to talk about the many hundreds of hours of labor that will go into such a project.

The design of the Yagi described in Section 3.5 includes three special features that makes this an especially interesting design:

- Instantaneous 180° directional switching with no compromises.
- Instantaneous SSB to CW switching with no compromises.
- Optimum F/B ratio over a wide bandwidth by capacitor-controlled compensation.

Nevertheless, some amateurs take the challenge. NO8D has a stack of two 3-element 80 meter Yagis shown in **Fig 13-32**. Recently, Juha OH8NC made his dream come true and built the biggest low band antenna in the world, known as Radio Arcala (See Section 3.11.).

3.6. Loaded 80-Meter Yagi Designs

Full-size 80-meter Yagis are obviously not for everyone. The investment is very important, and they are, let it be said,

Fig 13-33 — Three-element 80-meter Yagi using shortened elements at W6KW. The Yagi uses the high-Q loading coils developed by W6ANR. See text for details. The 55-meter high antenna sits on a knoll surrounded by flat terrain. To work on the antenna the platform visible along the tower is unfolded, and the Yagi is tilted 90° so that the center of the element is easily accessible from the platform.

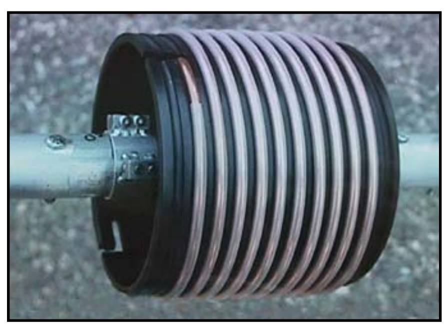


Fig 13-34 — The W6ANR/K7ZV loading coil, wound using ¼-inch copper tubing on a grooved ABS coil form, measures 10 inches long and 7 inches in diameter. Note the husky, large-area contact clamp used to connect the coil to the element. The ¼-inch copper tubing is covered with a plastic heat-shrink tube to protect it from surface contamination and surface leakage.

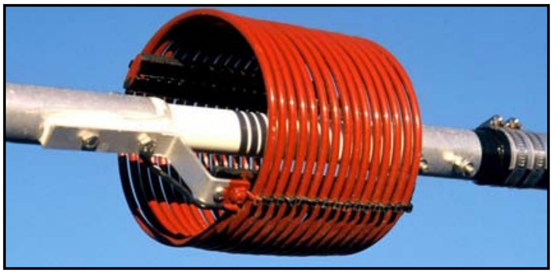


Fig 13-35 — Steve Babcock, VE6WZ, also makes his own high-Q loading coils. The coils are air-wound except for the two lateral strips of insulating material to provide mechanical rigidity. Note the redundant connection made with insulated wire.

Fig 13-36 — It is obvious that unprotected coils can get severely detuned by ice loading as shown in this photo at VE6WZ.

difficult to keep up. If carefully designed and well-made, Yagis with shortened elements can perform almost as well as a full-size Yagi. A three-element Yagi with shortened elements can be made to have just as good a directivity as its full-sized brother. It may, however, give up a fraction of a dB in gain.

3.6.1. Loading Coils Instead of Linear Loading

Until the mid 1990s it was commonly accepted that linear-loading devices (such as used for many years by KLM, Force 12 and M^2) were the best solution for low-loss loading of shortened low-band Yagis. Linear loading was assumed to have very low losses. Until then, for some strange reason, coil loading was generally assumed to be a lossy affair. Was that another tall tale?

As a consequence of a lot of experimenting and modeling in recent years our knowledge in this matter has improved a great deal. We now know that coil loading can actually be much better than linear loading.

David Padrick, W6ANR, decided to analyze loading coils in detail and found out that most commercial coils did indeed exhibit poor unloaded Q. But David also came to the conclusion that it was not all that hard to make coils with Qs of 650 or even more. (Ref 694). However, David found that the dimensions are critical, and high-Q coils able to handle high power should be quite large! (See **Fig 13-33**.)

A high-Q coil is not enough by itself. It is equally important to make very low loss RF connections. This is where many commercial designs failed in the past. Good RF connections mean connections that are wide with respect to their length! RF conductors also are crucial. The location of the coils on the elements is important (see Chapter 7). With very high-Q coils we can afford putting them out further on the elements, which increases the radiation resistance without introducing additional losses, provided the Q remains high.

Padrick found that commercially made Yagis using elements loaded with sloping stubs exhibit two sorts of problems: Inferior F/B because of radiation off the sloping loading stubs, and accumulated resistive losses of loading stubs. He pointed out that the length of the linear-loading wire is 2 to 3 times longer than the wire or tubing in a high-Q coil providing the same degree of loading.

In addition, the gauge of a loading stub is usually only a fraction of what's used for high-Q coils. Poor connections to the element and the relays and jumpers used to switch band-segments add to these problems, and all of these critical items have been found to deteriorate over time.

Changing from a linear-loading stub to a coil actually yields a net increase in R_{rad}. Tom, W8JI, tested linear-loading stubs and equivalent inductance coils. He found it was possible to build coils with Qs up to 800 (Ref 694), but also that the linear stubs never exceeded 100 (see Chapter 9)!

Peter Dalton, W6KW's new Yagi uses elements that are approximately 66% of full-size, and he has incorporated the loading coils out at about 55% of the total element length. Loading coils with an inductance of 17 µH were required to resonate the elements. See **Fig 13-34**. The final fine tuning of the parasitic elements was done by varying the length of the element tips. If these tips are not easily accessible, the fine tuning can of course be done by using a small central loading coil. A Q of 650, as quoted by W6ANR, means a series loss resistance of only 0.624 Ω per coil. If you want to know all the details of loading coil design visit **www.w8ji.com/loading_inductors.htm**.

I calculated the influence of the Q factor on the gain and directivity. The influence on gain can be seen in **Table 13-9.** In a typical 3-element coil-loaded Yagi, with the reflector spaced closer than the director and tuned for best F/B, the current magnitude in the reflector is almost the same as for the driven element, while the director only carries 15 or 20% that much current. As a consequence, the influence of coil losses on an-

Table 13-9
Influence of Coil Losses on Antenna Gain

Loss R	*0 Ω*	*0.5 Ω*	*1.0 Ω*	*1.5 Ω*	*2.0 Ω*	*2.5 Ω*	*3.0 Ω*
Q factor	—	800	400	333	200	160	133
Gain, dB	0 dB	–0.34	–0.66	–0.97	–1.27	–1.55	–1.82

Reference: 12.3 dBi

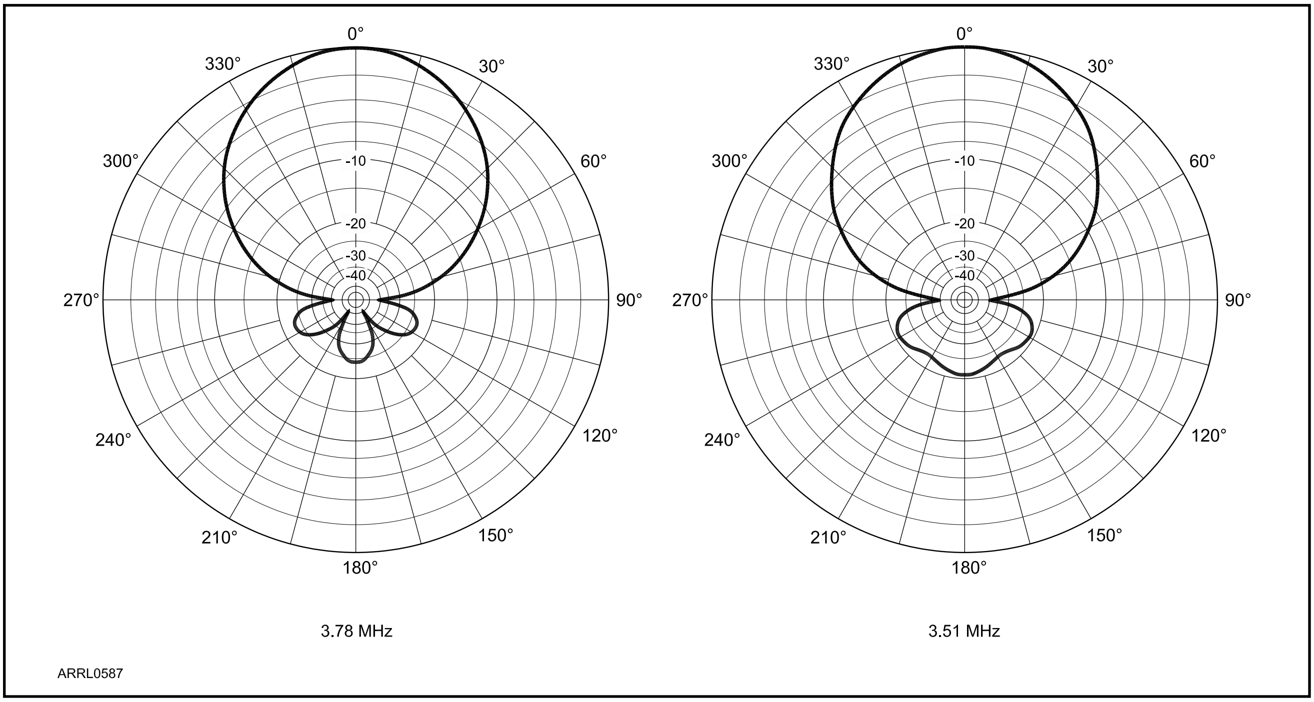


Fig 13-37 — Horizontal patterns of the high-Q coil-loaded 3-element 80-meter Yagi on 3.78 MHz and on 3.51 MHz.

tenna gain is the same for the driven element and the reflector, but much less for the director. In other words, if you have a lossy coil, better put it in the director element.

The Q factor has very little impact on directivity. An antenna that has been optimized for an excellent F/B with no coil losses, may actually show even marginally better directivity when small losses (1 or 2 Ω) are introduced! High losses reduce the mutual coupling to a degree that proper current magnitude and phase can no longer be set up in the elements to achieve a good F/B. As far as directivity is concerned, in the model I used it was possible to achieve a little deeper null with Qs of 200, compared to 400. This is actually irrelevant and only says something about the model, not about what can be achieved with a real antenna.

So far as directivity is concerned, you can say that there is very little to be gained by going for Q factors above 150 or 200, *but* from the point of view of total antenna gain, the higher the Q, the higher the gain. The difference in antenna gain between a Q of 700 and a Q of 175 is about 0.5 dB, which is like losing 15 % of your power!

3.6.1.1. How to Make High-Q Coils

The picture of the loading coils made by K7ZV in Fig 13-34 gives you part of the answer. Heavy gauge copper-tubing conductors (6 mm or ¼ inch) and good low-loss contacts. But there is more. There are not only series losses involved, but also parallel losses.

Any leakage current between adjacent turns of the coil causes parallel losses. If we keep the Q high (800), it means that we have an equivalent series resistance (for an inductive reactance of 400 Ω) of 400/800 = 0.5 Ω. This is the equivalent of a parallel loss resistance of 400 × 800 = 320 kΩ. If we allow dirt, smoke deposits, etc, to accumulate on the surface of the bare copper tubing of an unprotected coil, we can expect parallel losses to drop well below 320 kΩ, especially when it gets wet. Therefore the coil must be properly protected. One neat way to protect the copper tubing used to wind the coil with a heat-shrink tube of a material that has a good dielectric for HF and is UV-resistant (consult Raychem for such material). Later loading coil models used very heavy AWG 3 enameled copper wire. It doesn't pay to reduce the serial losses to almost zero if you don't take care of the parallel losses too.

Rich, K7ZV, who has one of the better 80-meter signals from the West Coast in Europe (together with W6KW and W6RJ), was the man who designed and made the coil forms as shown in Fig 13-33 and 13-34. Both W6RJ (shown later in Fig 13-41) and W6KW (Fig 13-33) use the same kind of loading coils on their beams. Until quite recently K7ZV had his own high-precision machine shop to construct the coils. Rich is now retired and that source of first class loading coils has dried up. It appears that David, W6ANR, also no longer has the coils available and is no longer able to help those who want to rebuild their Yagis using linear loading with high Q coils.

There are two manufacturers that still sell 80 meter Yagis with linear loading: M^2 (2 and 3 element) and Force 12 (2 element). I understand neither one of them intends to improve their products by using high-Q loading coils, as explained above.

However, making high-Q coils yourself is not such a formidable task. On his superb Web site (**www.qsl.net/ve6wz/**) Steve Babcock, VE6WZ, describes the design and construction details for his 2-element 80-meter Yagi,which he uses from his city lot on a 28-meter crank up tower to produce the most dominant signal from VE6-land into Europe. See **Figs 13-35** and **13-36**.

Steve used *coil.exe* (a DOS program from Brian Beezley, K6STI) to design his low-loss loading coils. He found that not only conductor size and coil dimensions were important, but that form loss plays a significant role in determining the unloaded Q. You must be very careful with the program, however, as it sometimes predicts unreasonably high Qs. The

Table 13-10
Full-Size 3-Element 80-Meter Yagi Data

Freq (MHz)	*Gain (dBi)*	*Rrad (Ω)*	*F/B (dB)*	*SWR*
3.74	7.3	31.1 – *j* 9.5	23 - 28	1.4
3.76	7.3	30.5 – *j* 5	23 - 40	1.2
3.78	7.4	29	25 - 30	1
3.80	7.5	27 + *j* 6	23 - 25	1.3
3.82	7.6	24.4 + *j* 12	19 - 30	1.6

Note: This performance data is for an 80-meter Yagi with full-size elements that serves as a reference for the development of a Yagi with reduced size elements.

Table 13-11
Reduced Size 3-Element 80-Meter Yagi Data

Freq (MHz)	*Gain (dBi)*	*Rrad (Ω)*	*F/B (dB)*	*SWR*
3.76	6.3	24.3 – *j* 9.6	15	1.6
3.78	6.1	32.4	24	1
3.80	5.9	38.6 + *j* 6.7	30	1.3
3.82	5.8	42.4 + *j* 12	20	1.5
3.84	5.7	44 + *j* 18	18	1.8

Note: After plugging in the coil loading values in our reference model, the performance data shown were obtained in our reduced size Yagi.

final coil measures 15 cm in diameter and 15 cm long, and uses 15.5 turns of 3⁄16-inch (~5 mm) copper tubing. The final physical coil design at VE6WZ is mostly air core, but uses two black ABS strips to give the required mechanical rigidity to the coil. Steve estimated a final Q of around 500 to 1000 (a loss resistance of 0.7 to 1.5 Ω). To prevent copper corrosion and to ensure maximum surface insulation (especially important near the ABS strips), Steve painted the copper winding with red electrical varnish (Q dope).

Steve also took all possible precautions to minimize contact and connection resistance. Notice in Fig 13-35 how the wide and thick aluminum strip is used as mechanical support and electrical connection as well. Steve also used a redundant connection made of an insulated wire.

Other suggestions and details about home brewing high Q coils can be found on W8JI's Web site (**www.w8ji.com/loading_inductors.htm**) and on the Web site of Greg, W8WWV (**www.seed-solutions.com/gregordy/Amateur%20Radio/Experimentation/HiQCoil.htm**).

Table 13-12
Optimized Size 3-Element 80-Meter Yagi Data

Freq (MHz)	*Gain (dBi)*	*Rrad (Ω)*	*F/B (dB)*	*SWR*
3.74	6.4	23.7 – *j* 10.6	13	1.6
3.76	6.15	31.3	22 - 30	1.0
3.78	5.95	38.0 + *j* 7.1	22 - 30	1.3
3.80	5.8	42.1 + *j* 12.9	20	1.6
3.50	5.9	37 – *j* 3	20	1.2
3.51	5.75	32	22	1
3.52	5.6	36 + *j* 2	19	1.1
3.53	5.5	38.8 + *j* 4.3	17	1.3
3.55	5.3	40.9 + *j* 7.9	14	1.4

Note: Performance data for the optimized design in both the phone as well as the CW section of the 80 meter band.

3.6.1.2. Replacing the Linear Loading Devices with High Q Coils

Building the high-Q coils for this job may not be the most the most difficult part. Determining the required value of the coils and tuning the Yagi appears to be more challenging. Without going into nitty gritty details, I would like to explain the general procedure to be followed for doing the conversion.

1) Hardware configuration: Carefully measure all the dimensions of the Yagi — tubing lengths (tapered sections), tubing diameter, boom length, element spacing etc.

2) Element-to-boom capacitance: Measure the element-to-boom capacitance. For that you should only have a short section of the element mounted (on one side is okay). The capacitance between the element and the boom should not be more than 100 pF (50 pF per side). If the value is more than 100 pF you are likely to have problems fine tuning the Yagi (see also Section 3.11.1). If the capacitance is larger, take appropriate measures to lower it (increase insulation distance).

3) Model a full size equivalent antenna: Using a *NEC2* or *NEC4* based modeling program (eg *EZNEC*), model a Yagi with full size elements on the available boom length, and with the available inter-element spacing. Optimize it according to your likings. You can consult the database that comes with the ON4UN *Yagi Design* software (see modeling file *ON4UN-80-YAG-fullsize.ez* on the CD). As an example I chose design #9 (Birgit). The model can be made with constant element diameter (eg 25 mm OD). Example: Boom length = 19 meters, spacing to reflector = 8 meters, spacing to director = 11 meters. Five minutes of modeling gave me a beautiful design (see **Table 13-10**).

4) Element resonant frequencies: Using the modeling program, we determine the resonant frequency of each element in the model (decouple the two other elements by inserting a high impedance in the center). The driven element by itself appears to be resonant on 3.765 MHz, the director on 3.98 MHz and the reflector on 3.58 MHz. We now know exactly on which frequencies the elements of our coil-loaded Yagi need to be resonant.

5) Model the reference element: As we want to use identical loading coils in the three element, we must first design the director. Start by entering all the tapered sections. Put a loading coil in each half element. Let us assume the driven element is 29 meters long (approximately 70% of full size) with the coils 8 meters out on the elements (a little over half way out). Using an element taper going from 60 mm OD in the center to 12 mm OD at the top, this element requires a loading coil of approximately 11.1 µH to resonate the element on 3.98 MHz (see modeling file *ON4UN-80-YAGI-1el.ez*). **Table 13-11** shows the data for a Yagi with reduced size elements.

6) Real life model verification: Now we want to verify our loaded dipole model versus real life. To do that we build this director element, mount it on a piece of boom (to include the element to boom parasitic capacitance), raise it to its final height on the tower and measure its resonant frequency and impedance with an accurate antenna analyzer. If you're slightly off the desired frequency (3.98 MHz), trim the tips of the elements. If you are far off you may have to change the loading coil inductance. Once this element is resonant at the design frequency, this will be your *reference element*, which means: *don't touch it from now on*. Let's assume the element is resonant exactly where we want it (3.98 MHz).

7) The reflector element: Now we run the loaded element model on 3.58 MHz, and see how much of a central loading coil is required (the 11.1 μH coils represent an inductance of 249 Ω on this frequency). Our modeling program tells us that a coil with an inductance of 200 Ω (8.9 μH on 3.58 MHz) will be required to resonate the element on 3.58 MHz. That is a reasonable value (coil approximately 14 turns, diameter 7.5 cm, length 9 cm). You can of course do the same model verification with the reflector, as describe above.

8) The driven element: Finally we run the loaded element model on 3.78 MHz, where the 11.2 μH loading coils represent an inductance of 266 Ω. It is obvious that the driven element will also require some loading, but its resonant frequency is irrelevant as to the performance of the Yagi. We can make the loading part of the matching system.

9) The complete Yagi model: Now the time has come to complete the model which so far consists only of a loaded dipole with two more elements tuned to the frequencies we have calculated above. Plugging the five coils with 11.7 μH inductance (277 Ω on 3.78 MHz) into the model yields a very nice pattern (see **Fig 13-37**) with a gain of 6.11 dBi.

10) Optimizing: It appeared that we could best lower the passband by approximately 20 kHz, which resulted in a slightly higher value for the loading coils (12.1 μH). The reflector loading coil value was 6.4 μH (file: *ON4UN-80-YAGI-3el-SSB.ez*).

11) CW band: Reflector center loading coil: 12.1 μH, director central loading coil: 6.4 μH (file: *ON4UN-80-YAGI-3el-CW.ez*). **Table 13-12** shows performance data for an optimized design in both the CW and phone bands.

12) Putting the antenna at real height: Most 80-meter Yagis are *not* several wavelengths above ground. A typical height is 30 meters. The proximity of the ground will slightly change the values of the loading coils. For this example, the changes were minimal, but further model optimization can be done.

Fig 13-38 — Loading coil as manufactured by Optibeam and used in their 2- and 3-element reduced size Yagis.

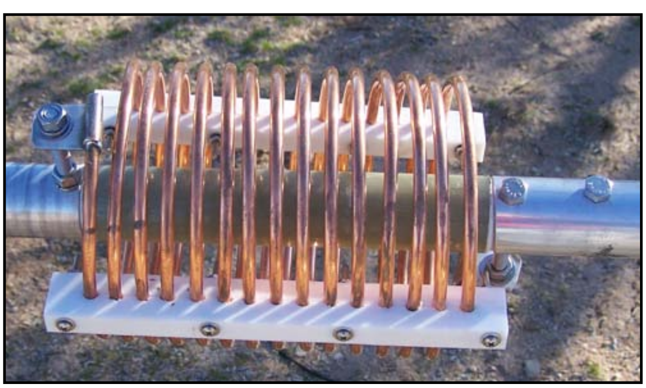


Fig 13-39 — These beautifully crafted loading coils for converting an old KLM 3-element 80-meter Yagi were built by Larry, N7DD, after a design by N7RT.

The Loading Coils

Using the *ON4AA Coil Calculator* (available online at **hamwaves.com/antennas/inductance.html**), we find the required inductance of 11.2 μH using an air-wound coil with a diameter of 100 mm, using 6 mm OD tubing as the conductor. The coil is formed from 15.4 turns with a coil length of 154 mm (1 turn per 10 mm). The calculated unloaded Q is 1130 (equivalent loss resistance = 0.24 Ω).

Using a coil with a diameter of 75 mm and using 3 mm OD enameled wire, the coil (now ~17 turns with a length of 102 mm) has a Q of approximately 800 (R_{loss} = 0.35 Ω), which is still very good.

The German Yagi manufacturer Optibeam makes reduced-size 80-meter Yagis (models OB3-80 and OB2-80) where the element length is approximately 58% of full-size, using loading coils of 18 μH with a fairly large diameter (135 mm). The coils are wound with fairly thin aluminum wire, 2 mm OD. This yields a very acceptable Q factor of approximately 750 (equivalent loss resistance = 0.7 Ω). See **Fig 13-38**. Another way of making loading coils is shown in **Fig 13-39**.

We should not forget that in all the above calculations we have only considered the series type losses, but there are also the parallel losses. Most, if not all of the high-Q loading coil designs are "in open air" (unprotected), because open (dry, clean) air is a very good dielectric. These large open coils are also subject to significant inductance changes in wet and snowy/frosty weather In addition they get covered by all kinds of contaminants and impurities that float abundantly in the air, especially in and near cities, or in the neighborhood of smoke stacks. A black deposit of smoke particles causes the insulating quality of the coil form to deteriorate over time, increasing shunt losses, and will eventually also change the inductance of the coils. Therefore it's a good idea, if possible, to periodically clean the loading coils. Several techniques can be used to protect the coil wire in order to reduce parallel losses: use double coated enameled copper wire, protect the copper tubing with heat shrink tubing (with hot melt) or cover it with Q-dope (red insulating varnish type 4228, available from MG Chemicals), as suggested by VE6WZ.

How important is this Q-factor (see also Section 3.6.1)? It has little influence on the directivity of the Yagi, but it mainly influences gain. How much? On our 12.1 μH coils, going from a Q-factor of 1130 (6 mm copper tubing coils) to a Q of 660 (2 mm aluminum wire) reduces the gain of the array less than 0.1 dB! This does not mean that any Q is okay. A very popu-

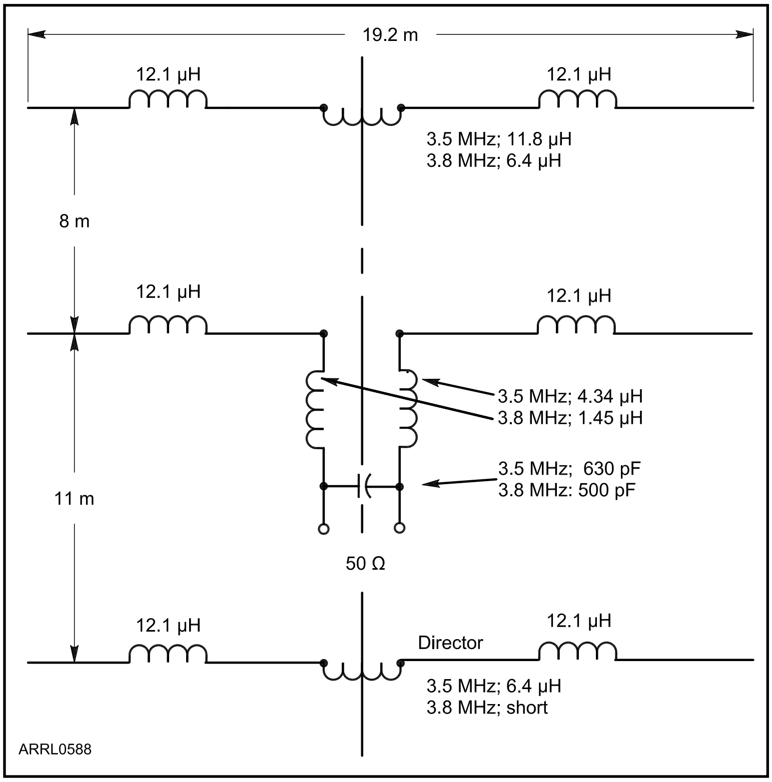


Fig 13-40 — Values of the loading devices for both the phone and the CW end of the band. The relay switching is not shown.

Fig 13-41 — This 3-element shortened 80-meter Yagi of W6ANR design with K7ZV coils is installed at W6RJ's mountaintop QTH.

lar commercial 40-meter reduced size Yagi uses loading coils with a Q-factor of approximately 80, resulting in equivalent serial losses of 8 Ω (see **www.qsl.net/ve6wz/CC_coil.html**). If we used such low-Q coils on our example design, we would lose almost 2 dB of gain, which means that 37% of the applied power would be used to heat (and maybe burn) the loading coils.

Note also that going from a full-size 3-element Yagi to a Yagi with the same general design, but using elements that are 70% of full size results in a gain loss of approximately 1.5 dB. Maybe 5.7 dBi gain (in free space) does not sound like a lot? After all a Four Square has 6.5 dBi gain over good ground. To compare apples with apples, we need to calculate the gain of the Yagi over real ground, which also means we have to include ground reflection gain (see Chapter 8, Section 1.2.1.4). With the Yagi at a height of 30 meters, the gain, at the main radiation angle (33°) has increased to 9.75 dBi, caused by ~4 dB ground reflection gain (over "average" ground). This means that the 3-element reduced-size Yagi has an operational gain of approximately 3 dB over a Four Square.

Matching and SSB / CW Switching

On 3.78 MHz the feed impedance without any matching or loading applied to the driven element is 39.5 – j 47.1 Ω. Transformation to 50 Ω can be done with a balanced L-network, for which we can calculate the values using the L-Network module of the *ON4UN Low Band Software*. Plugging these values (L = 1.45 µH and C = 500 pF) into our model (see file *ON4UN-80-YAGI-3el-SSB-2.ez*), the SWR in the SSB band is 1.4:1 at 3.8 MHz, 1:1 at 3.77 MHz and 1.5:1 at 3.75 MHz. One could also use a Beta (hairpin) matching system and use a parallel coil instead of a parallel capacitor (see Section 3.10.3). Using a parallel capacitor however makes it easier to fine-tune the matching system by using a variable capacitor.

On 3.510 MHz the feed impedance without any matching or loading applied to the driven element is 33.2 – j 178 Ω. with a balanced L-network, for which we can calculate the values using the L-Network module of the *ON4UN Low Band Software*. Plugging these values (L = 4.34 µH and C = 630 pF) into our model (see file *ON4UN-80-YAGI-3el-CW-2.ez*), the SWR in the CW band is 1.2:1 at 3.5 MHz, 1:1 at 3.51 MHz going up to 1.3:1 at 3.54 MHz.

Fig 13-40 shows the coil and capacitor values for the Yagi. It is obvious that the feed impedance is balanced, and that a current balun must be used.

Conclusion

I have tried to explain all the steps one has to go through when planning to do the conversion for a 2 or 3 element Yagi with linear loading devices to a Yagi with high-Q loading coils. This is just an example. The aluminum tubing section lengths and diameters I used in this example are likely not what you have, so the results you will obtain will be slightly different. Also, you might want to base your design on another full-size model than what I took (design #9 from my *Yagi Design* software). It may look complicated, but the results will certainly be rewarding.

3.6.1.3. Other Big Guns on 80 Meters Using Shortened Yagis

Rich, K7ZV,who made those beautiful loading coils, lives in Oregon, close to the California border, in a county with lots of small mountains. One of those is his mountaintop. From his house the terrain slopes down in all directions. This is a real dream QTH, although there must be an important degree of signal scattering from the many other hilltops in just about all directions. Rich says it does not make much difference whether he has his 80-meter Yagi antenna retracted to 16 meters or at its full 40 meters of height. The effective height is impressive in both cases. Fully retracted the antenna is about 12 meters high. His 3-element Yagi is based on the W6RJ version of M^2 aluminum, with the W6ANR/K7ZV loading coils.

Using *Yagistress* (by K7NV), K7ZV redid the mechanical design of the standard 80-M3 80-meter Yagi by M^2, which was very similar to the original KLM design. The original boom was lengthened from 17.5 to 20 meters and strengthened as well. The new design uses a totally different boom and element guying system, which allows tipping the boom vertical to access the director or reflector without the need for a crane. This makes assembly, installation and maintenance much easier. The elements are different as well, and while maintaining a similar taper schedule they are much stronger, with double-wall tubing throughout most of the of the span. With guying just inboard of the coils at the center of the half element they are also much stronger.

Fig 13-42 — OH2BH's 3-element 80-meter Yagi was flown by helicopter from the assembly area to the tower, which is in the middle of the woods about 1 km away.

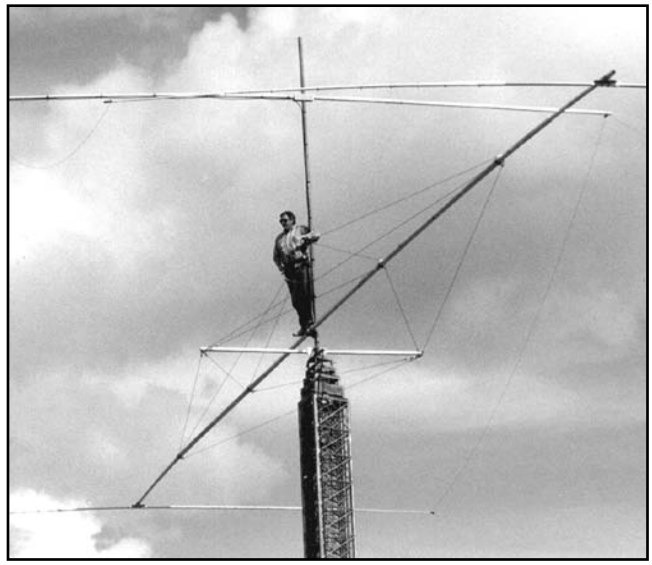


Fig 13-43 — Rod, W7CY, stands proudly on the boom of his 2-element Moxon-type loaded 80-meter Yagi.

Rich's most impressive tool is his bucket truck (cherry picker), which is permanently parked at the base of the antenna tower. Need to work on the antenna? Need to measure something? Just start the truck, up goes the basket and within minutes you can reach any part of the antenna, at a height where it still works! Rich's 3-element Yagi has been designed for 3.8 MHz but can be switched to the CW end of the band by inserting the appropriate loading coils in the three elements.

Bob Ferrero, W6RJ, eminent low band DXer and contester and owner of HRO, was formerly K6AHV. Bob is well-known in the world of 80-meter DXers. He built his dream station on a 1000-meter high mountaintop about 70 km from his home on the east side of the San Francisco Bay. No neighbor within maybe 5 or 10 km, which means no noise, just nature and the sky.

In **Fig 13-41** you can see his 40-meter tall microwave tower, topped with the heaviest duty tower that US Towers makes. When fully extended, this can make the 3-element 80-meter Yagi almost disappear in the sky.

The crank-up tower and tilting boom allow all antenna work to be done from the large platform at the 40-meter level on the microwave tower. What an amazing sight! The tower and the shack sit right on top of the ridge, sloping quite steeply in most directions. This makes the effective height of the antennas very high. Bob told me it hardly makes any difference whether he has the crank up extended or not. I think it does not make a bit of difference, given his QTH!

Since winter 2003/2004 Bob now also runs a wire Four Square for 160 meters from the top of the tower, and that does the trick for him on Top Band. He also operates the station on his mountaintop from his home by UHF remote control.

World famous Martti Laine, OH2BH, is another owner of a 3-element 80-meter Yagi according to the W6ANR design with K7ZV coils. See **Fig 13-42**. The antenna was built using the W6RJ version of M^2 aluminum, but was heavily reinforced with three-dimensional truss wires on the elements, to help them cope with Scandinavian wind and ice loads. Martti can also operate his station in the middle of the woods remotely from his Helsinki downtown QTH.

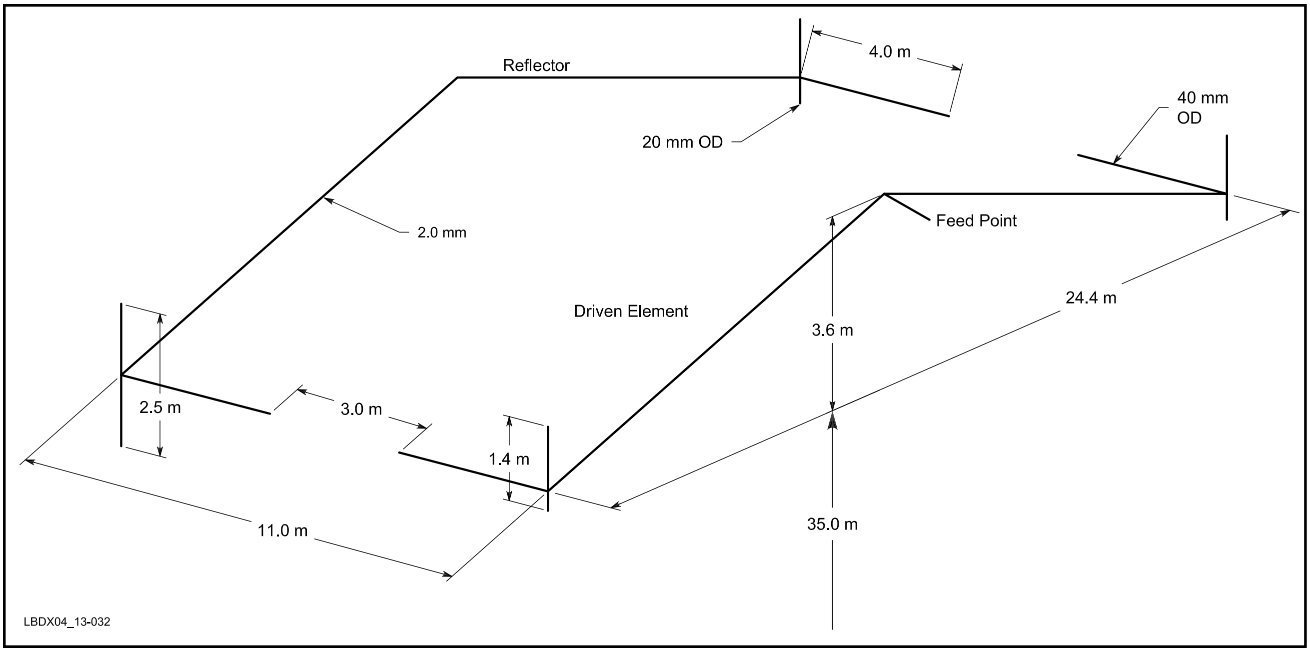


Fig 13-44 — Approximate dimensions of the W7CY 2-element 80-meter capacitively coupled Yagi.

Steve Babcock, VE6WZ, was obviously inspired by the work done by W6ANR and K7ZV when he set out to build a 2-element 80-meter Yagi that would fit into his city lot on a 28-meter crank-up tower. Steve told me, "You must also know that it was your book that inspired me to build it." Steve has a Web site where he describes the design and construction of his antenna in great detail (**www.qsl.net/ve6wz/**). Steve also said, "This 2-element Yagi is by far the best homebrew antenna I have ever built. It has substantially exceeded my expectations! I have built many, many other homebrew antennas over the years, but none have ever performed so well." Whereas K7ZV uses his bucket truck to get access to his antenna for measuring and experimenting, Steve uses the roof of his house as a work platform. His heavy-duty 28-meter crank-up tower is just adjacent to his house. From the roof he has access to the whole antenna. To achieve the results Steve does, his antenna must be very carefully tuned. It certainly is not a plug-and-play design. But Steve is usually the first, if not the only one, I hear in Europe when the band decides to open up from his northerly location. It is amazing what Steve does from a residential area.

Charley, WØYG, is another addict of large loading coils. After a disappointing experience with a Yagi using linear loading stubs, he rebuilt his antenna with high-Q loading coils. He now says, "If you doubt the efficiency of this design listen some time in an 80-meter pileup. The guys who really rule the roost use those W6ANR converted beams. They are all over the West Coast, some in the interior of the States and some in Europe. The design is practical, tuning is a snap and results are, well, fantastic!"

3.6.2. W7CY 2-Element 80-Meter Yagi

Rod Mack, W7CY, developed an interesting 2-element, capacitively end-loaded 2-element Yagi for 80 meters, as shown in **Fig 13-43**. He published pictures and some raw dimensions on the Internet Web pages at **www.ulio.com/ants.html**. He claims in excess of 20 dB F/B, which is in

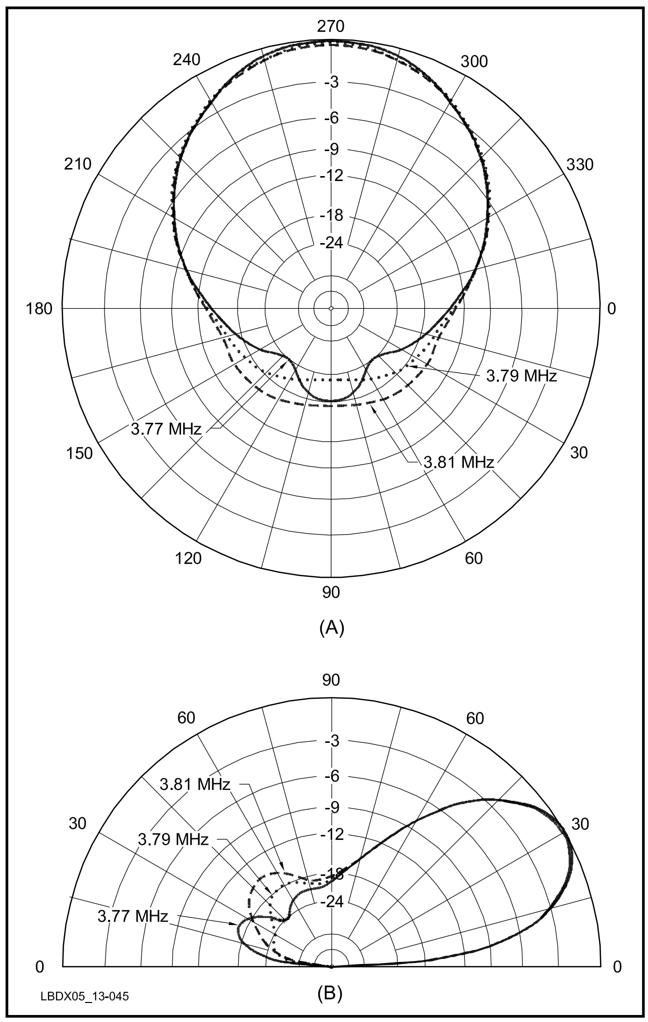


Fig 13-45 — Horizontal and vertical radiation patterns for the W7CY Yagi at a boom height of 35 meters. The F/B is just short of spectacular for a 2-element Yagi.

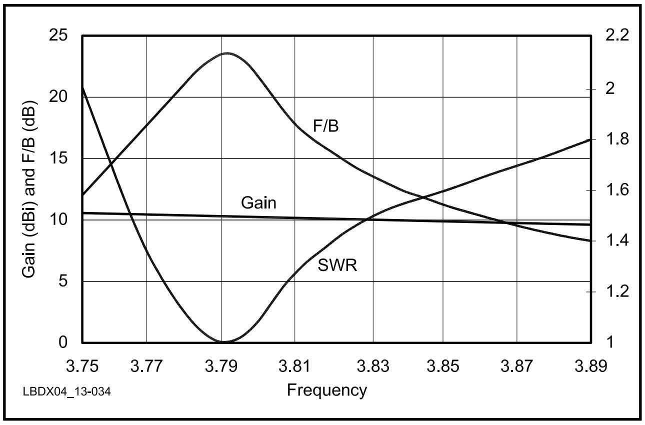


Fig 13-46 — Gain over average ground, F/B and SWR for the W7CY 2-element shortened 80-meter Yagi as a function of frequency

line with other antennas using capacitively coupled element tips, similar to the Moxon-type Yagis popularized by L. B. Cebik, W4RNL (SK).

I modeled W7CY's array using *EZNEC* and noticed some very interesting things. The layout of the array is shown in **Fig 13-44**. It uses a 24.4-meter long boom. At both ends of the boom he mounted 11-meter long spreaders, which serve as capacitive loading devices for the sloping elements. The center of the two elements is supported by an 11-meter boom, which is 3.6 meters above the main boom. This means that the central part of the elements is an inverted-V.

Rod stated that the capacitive loading spreaders are, of course, insulated from the boom, since they carry very high voltages at their ends. This is where he tunes the array by varying their length. Modeling demonstrates that the distance between the tips of the loading devices near the boom is a very critical item! By varying this distance, you can control the coupling between the two elements to a fine degree. Again, this is similar to the "Moxon" type of Yagi design.

For an element spacing of 11 meters, the ideal current relationship in the two elements is 1:1 for current magnitude and 125° for phase shift. You can model this easily by "planting" two verticals spaced 11 meters (on 3.8 MHz) and feeding them that way. If the tips are too close together you will have too much current in the reflector. With parts near the element tips so close together you can create a great deal of capacitive coupling.

I developed a system by which I loaded the sloping elements in two different ways (both capacitively) — by changing the length of the horizontal loading wires (the support structure), and by adding some vertical aluminum tubing at the same point. By judiciously weighing the ratio of these two capacitive loading devices, I arrived to a point where the required current ratios in both elements were obtained. At that point the F/B was over 24 dB, together with a feed-point impedance of very close to 50 Ω.

I could not obtain these results by loading the elements with only the horizontal "spreader" loading tubes; they gave me too much coupling between the two elements, as their tips came too close together. I found this out by inserting a resistor in the reflector, which reduced the current magnitude with a sacrifice of gain. The same result was obtained by spacing the tips of the horizontal "loading wires" further apart.

The array has a very nice bandwidth as well. **Fig 13-45** shows the radiation patterns for the array over a span of 40 kHz, without retuning the reflector. It is of course possible to tune the reflector by installing a variable capacitor in the center of the element and changing its capacitance as you move around on the band. By doing so the same F/B can be achieved (20 to 25 dB) anywhere in the band. Even without doing any retuning, this array has an SWR of less than 2:1 over more than 150 kHz. **Fig 13-46** shows some essential array data for a frequency range of 3750 to 3890 kHz.

3.6.2.1. Duplicating the W7CY Antenna

First of all, it is important to know that the dimensions shown in Fig 13-44 are ballpark figures. These are by no way "build-and-forget" dimensions. I modeled this antenna with several modeling programs: *AO* (*MININEC* based), *ELNEC* (*MININEC*), *EZNEC* (*NEC-2* based) and *EZNEC-PRO* (*NEC-4* based). Although all of these programs achieved essentially the same radiation patterns after fine-tuning, these results were all obtained with slightly different dimensions. The main reason for this is the inability of some of these programs to handle wires with vastly different diameters. The problem lies in modeling the capacitance hat, which is made out of aluminum tubing, while the rest of the element is made of a much thinner wire. As an example, with the dimensions optimized for 3.79 MHz using *NEC-2*, the frequency shifted down approximately 80 kHz using *NEC-4*. The dimensions shown in Fig 13-44 are the results of modeling with the *NEC-4* engine, which is supposed to give the most accurate results in this case.

Whereas these models may not give us the exact lengths for a precise operating frequency, they give us a good idea of what can be achieved so far as directivity is concerned. Further, it is very important to model in order to determine the resonant frequency of the parasitic reflector and the driven element. We can use this information to tune the array in real life.

The models tell us that for an array optimized for 3.79 MHz, the reflector is resonant on 3.80 MHz (yes, higher than the design frequency!), and the driven element by itself is resonant on 3.94 MHz. This clearly shows what mutual coupling does!

You must determine the resonant frequency of the driven element and the director with the other element removed from

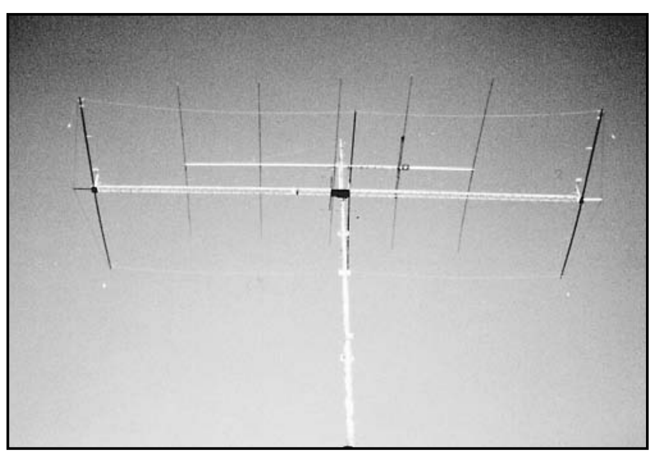


Fig 13-47 — The K6UA 2-element 80-meter array mounted on a Telrex rotating pole, just under a 5-element 20-meter Yagi, which is dwarfed in comparison.

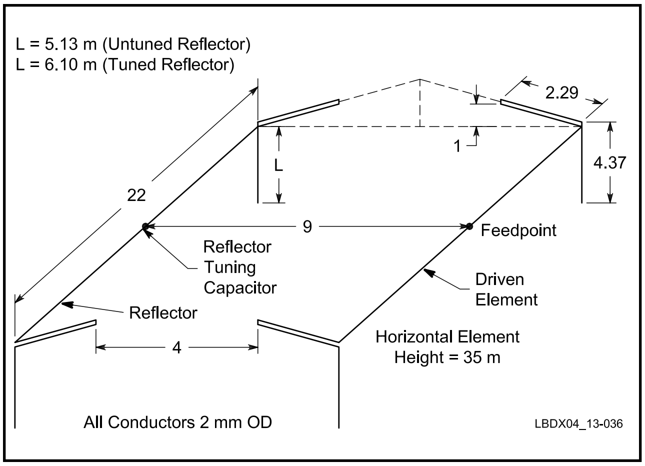


Fig 13-48 — Sketch of a 2-element 80-meter Yagi similar to K6UA's design. The dimensions are approximate, and were used to calculate the patterns of the array. The wires of the loading stubs in this model are separated 20 cm.

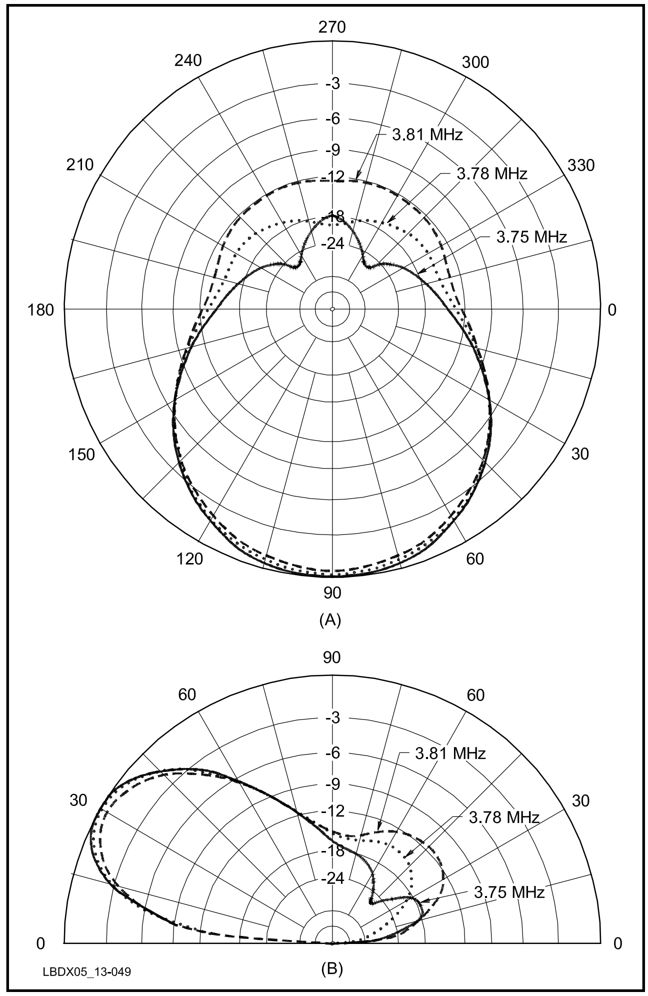


Fig 13-49 — Radiation patterns for the K6UA Yagi. The F/B is 23 dB at the design frequency for an elevation angle of 30°.

the model. For example, I decoupled the other element by inserting a load of R = 9999 Ω and X = 9999 Ω in the center of the element. Armed with this information, here is how we can tune the actual array:

1) Build the array according to the dimensions of your model. Be prepared, however, to change the dimensions of the loading devices.
2) Cut a feed line that is λ/2 at the resonant frequency of the reflector (in our case 3.8 MHz). For RG-213 cable the length is (0.66 × 299.8) / (3.8 × 2) = 26.03 meters. If you cannot reach the end of the feed line, you can use a full wavelength feed line as well (52.06 meters).
3) Connect this feed line to the reflector, raise the antenna to final height, decouple the driven element (leave the center open) and now adjust the loading devices symmetrically on both sides until you get resonance on 3.8 MHz. That takes care of tuning the reflector. Remove the feed line and close the reflector.
4) Now connect the feed line to the driven element, and prune the length of the loading devices for minimum SWR at the design frequency (3.79 MHz).

3.6.3. The K6UA 2-Element 80-Meter Yagi

Dale Hoppe, K6UA (SK), must have been around 80-meter DXing almost as long as the band has been there. Dale was quite a character, and too bad "K Six United America" became a Silent Key in December 2007. Some old timers may remember Dale as "W6 Very Strong Signal." With his beautiful hilltop QTH on an avocado plantation, not only avocados grew well, but also antennas!

Although from this way-above-average QTH almost any antenna would work, Dale was an avid antenna experimenter and builder. The latest of his designs was a 2-element shortened 80-meter Yagi, which he described in *CQ Magazine* (Ref 978). It is clear that the tower and the boom, which I have seen at Dale's place for ages, were what set him on his way to develop the array (see **Fig 13-47**).

The boom of the array is a 22-meter long triangular tower, 30-cm wide. Fiberglass vaulting poles, measuring 4.5 meters long, were mounted at both ends of the boom, providing the 9-meter spacing between the driven element and the reflector. The 22-meter long horizontal elements are loaded at both ends by loading stubs, as shown in **Fig 13-48**. The linear-loading stubs also serve as vertical bracing for the vaulting poles.

Dale reports raising and lowering the antenna about five times and cutting the length of the vertical trim wires to tune the array (see Ref 978). Tuning the reflector is quite critical, and changes of a few centimeters can make an important difference in antenna Q. If the reflector is too short, the feed-point impedance will be much lower than 50 Ω and the bandwidth will be very narrow. When properly tuned, the array exhibits a gain, F/B and SWR pattern as shown in **Fig 13-49**. The antenna has a fairly high bandwidth above its design frequency, and the gain remains fairly constant as well. Depending on the wave angle considered the F/B is >20 dB over approximately 50 kHz. This is quite a good figure for a 2-element Yagi.

This array is very similar to the W7CY array, the difference being slightly shorter elements, closer spacing and partial inductive loading of the elements. As with the W7CY array, the position of the tips of the loading elements facing one another (the tips of the stubs) determine the degree of coupling of the two elements in the array. The amount of coupling is quite critical to obtain maximum F/B ratio. In the model I used, a physical spacing of approximately 4 meters between the tips of the loading stubs gave the best results.

Instead of using aluminum-tubing capacitance hats, which cause a modeling problem due to the vast difference in diameter between the wires in the antenna (2 mm) and the tubes, I decided to keep the original K6UA approach and use "dangling" wires to tune the array. The length of these wires can be trimmed to change the frequency of the elements. To keep them more or less in place in the breeze, you could hang small weights at the end of those wires. The dimensions given in Fig 13-48 were obtained by modeling through *EZNEC* (*NEC-2* engine).

One way of tuning the reflector is to watch the SWR about 30 kHz below the design frequency and adjust the reflector (shorten it) until the SWR is about 2:1 (see **Fig 13-50**). Tuning of the driven element to obtain lowest SWR at the design frequency is the last step in tuning the array. The antenna can be fed directly with a 50 Ω feed line via a choke balun.

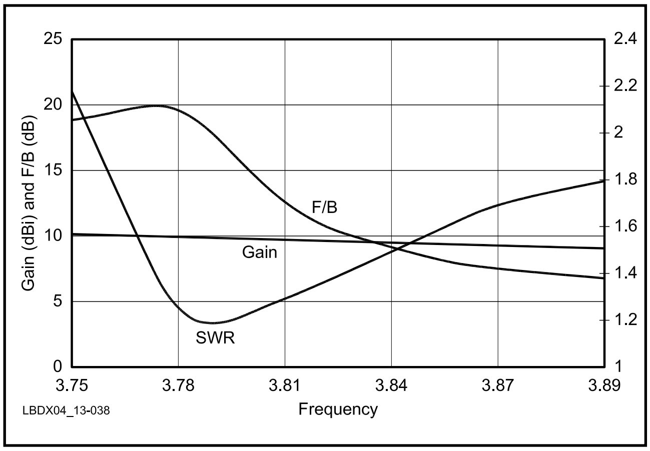


Fig 13-50 — Gain, F/B and SWR for the 2-element K6UA 80-meter Yagi over average ground.

3.6.3.1. An Alternative Tuning Method

Raising the Yagi repeatedly in order to tune the antenna for best F/B may not be the most attractive job. There is an alternative way though, that brings you an additional advantage. I intentionally lengthened the reflector a substantial degree, and made the vertical tuning wires about 1 meter longer (6.0 meters instead of 5.13 meters). Now we have a reflector that is way too long and we can electrically tune it to where we want it, by simply inserting a capacitor in the center of the element. In the model case I achieved 23 dB F/B on any frequency between 3.75 MHz and 3.87 MHz, simply by adjusting the value of the capacitor from 663 pF on 3.75 MHz and 354 pF on 3.87 MHz (see **Table 13-13**)! **Fig 13-51** shows the vertical radiation patterns obtained at various frequencies in that range.

The same approach for remotely tuning the reflector can of course be used on the W7CY 2-element Yagi. In conclusion, this design is a good example of what can be achieved using locally available materials and a good deal of knowledge, insight and imagination. Dale's big signal on 80 meters was the best proof of it.

3.7. Horizontal Wire Yagis

Yagis require a lot of space and electrical height in order to perform well. Excellent results have been obtained with

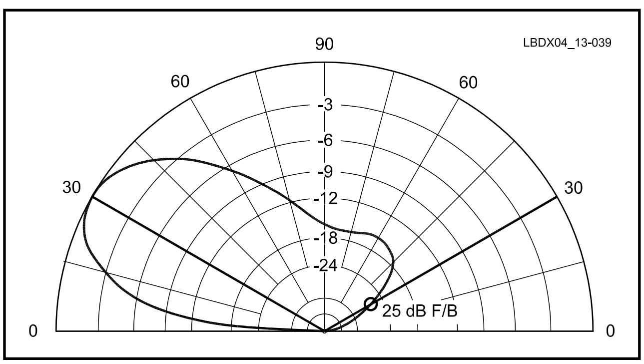


Fig 13-51 — Vertical radiation pattern for the K6UA 2-element 80-meter Yagi, showing the directivity that can be obtained across 150 kHz of bandwidth by remotely tuning the reflector with a variable capacitor.

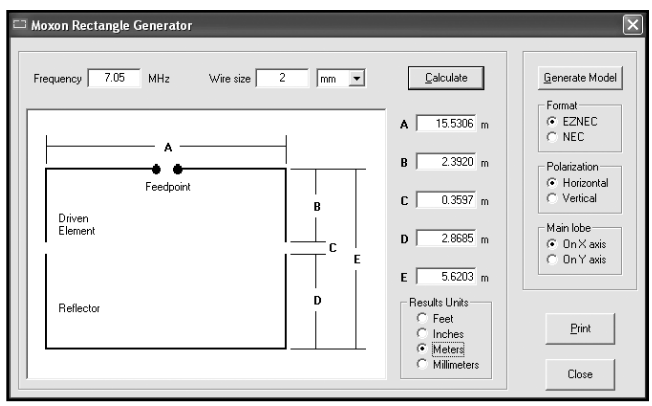


Fig 13-52 — The *Moxon Rectangle Generator* developed by AC6LA based on the work of L.B. Cebik, W4RNL (SK).

Table 13-13

Values of the Capacitive Reactance (X_C) and Corresponding Capacitance (pF) to Tune the K6UA for Best F/B Across a Wide Spectrum. Curve SWR (1) is for the Array Tuned for Best F/B at >23 dB (R_{rad} About 37 Ω). SWR (2) is for the Array Adjusted for 50-Ω Impedance (F/B Approximately 15 dB).

Freq (MHz)	*3.75*	*3.78*	*3.81*	*3.84*	*3.87*	*3.9*
X_C (Ω)	–64	–77	-90	–103	–116	–129
C (pF)	663	547	464	403	354	317
SWR (1)	2.0	1.5	1.3	1.5	1.9	>2.0
SWR (2)	1.4	1.1	1.2	1.6	2.0	>2.0

fixed-wire Yagis strung between high apartment buildings, or as inverted-V shaped or sloping elements from catenary cables strung between towers. There are few circumstances, however, where supports at the right height and in a favorable direction are available. When using wire elements, it is easy to determine the correct length of the elements using a *MININEC* or a *NEC*-derived modeling program (eg, *EZNEC*). Wire Yagis have been described in detail in Chapter 12 (Other Arrays).

3.8. The Moxon Rectangle

The rectangular-shaped antenna, which in the recent years has become popular under the name "Moxon Rectangle" is in fact a further development of the VK2ABQ antenna. It was L.B. Cebik, W4RNL (SK) who later undertook a very large modeling project to determine the dimensions for Moxon rectangles with uniform-diameter elements. His work resulted in design equations that were used by AC6LA to develop a simple *Windows* design program, the Moxon Rectangle Generator (**www.qsl.net/ac6la/moxgen.html**). See **Fig 13-52.** Details of Cebik's work are available at **www.antennex.com/Sshack/moxon/moxon.html**

In a nutshell: the Moxon rectangle is a 2-element parasitic array that uses reduced element lengths (approximately 75% of full size). The ends of the elements are brought very close to each other, resulting in a much tighter coupling between the elements than is the case with a regular Yagi. The net result is that the magnitude of the current in the reflector is almost as high as for the driven element. Properly designed, one can achieve a much higher F/B ratio than is possible with a standard 2-element Yagi, the gain being almost the same. The design equations developed by L.B. Cebik also result in a very convenient 50 Ω feed impedance.

I have not seen a design tool yet that makes it possible to use a mixture of tapered diameter elements and wires. This clearly must be related to the fact that a Moxon antenna is not a "forgiving" design. It is a critical design, because of the very critical coupling between the two elements. To achieve the best possible modeling results (>30 dB F/B) in an actual antenna, the spacing between the folded back wires of the driven element and the reflector are critical to within a few millimeters. In real life though, the variable influence of a real ground makes such figures unattainable. This is probably also the reason why — to my knowledge — only one commercial antenna manufacturer so far has used the Moxon rectangular array concept. The Optibeam OB2-40 Moxon antenna for 40 meters uses coil loading to further reduce the element length. All of this degrades the Moxon rectangle in many cases to a reduced size 2-element Yagi without exceptional directivity characteristics.

3.9. Vertical Arrays with Parasitic Elements (Vertical Yagis)

Do vertical arrays with parasitic elements work on the low bands? If you are a Top Bander, look in your log for KØHA. He's either there long before anyone else from his area, or he's there all by himself (at least over here in Europe), or he's there much stronger than anyone else. Bill Hohnstein, KØHA, swears by his vertical Yagis. His farm grows vertical parasitic arrays in all sorts and sizes (see **Fig 13-53**).

There is no need to use full-size elements for putting together effective and efficient arrays. Bill uses a shunt-fed 32-meter tower as the driven element for his 160-meter array, while his parasitic elements are approximately 26 meters high, and top loaded.

It is obvious that in a parasitic array, neither the feed method nor the exact electrical length affect the performance of the array. Shunt or series feeding can be used without preference. The elements should, however, not be much longer than ¼ λ.

I will take you on a little tour of some of the classic parasitic arrays, and point out what you should watch for if you want to build one. A modeling program seems to be essential, as you probably will be using existing towers as part of the antenna, and you will need to do some specific modeling. Watch out that you understand what the modeling program tells you, and be aware of what it does not tell you.

3.9.1. Turning Your Tower Guy Wires Into Parasitic Elements

Several good articles have been published on this subject (Refs 981, 982 and 983). I recommend reading those if you plan to try one of these antennas.

3.9.1.1. An Array with One Sloping Element

It seems logical to think of a sloping guy wire as a reflector or a director. But, you can also use the sloper as a driven element, and use the tower as a parasitic element! This last case may not be so practical, since in many cases it will probably not be possible to tune the tower to the exact required length. The tower could be tuned by changing its length, or by tuning it — eg, at its insulated bottom by inserting a coil or capacitor to ground.

I analyzed the case of a 40-meter high tower (25 cm

Fig 13-53 — Bird's eye view from the driven element of the 160-meter array at KØHA. The first director, on the left of the picture, is hiding from the second director. On the right a line of elements for the 80-meter array aims at Europe. It also appears that Bill enjoys some of the best ground conductivity around. No wonder he's loud!

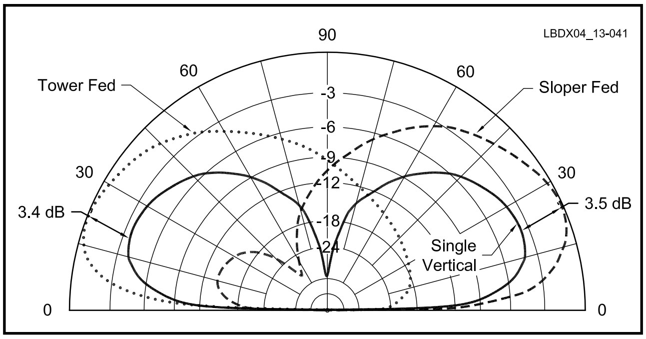


Fig 13-54 — Vertical radiation patterns for the tower and one sloping wire. See text for details.

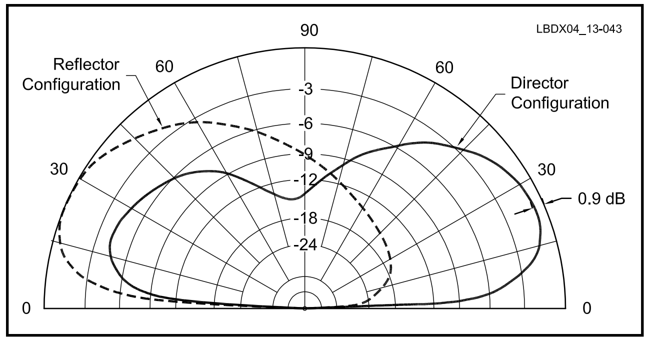


Fig 13-56 — The same sloping wire tuned as a reflector and as a director. The director configuration yields a poor F/B and mediocre gain.

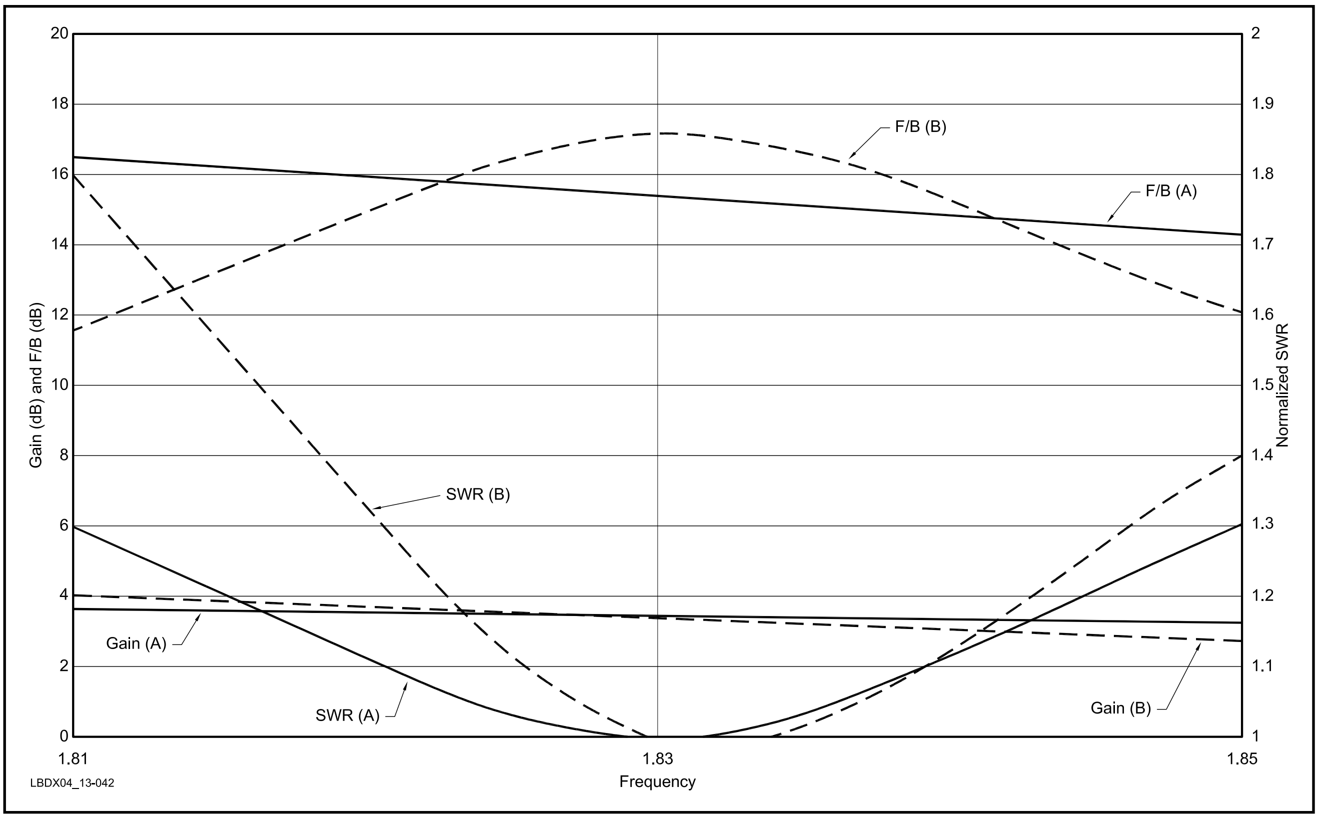


Fig 13-55 — Gain, F/B and normalized SWR for the vertical and sloping wire array. Case A is the array with the sloper being fed, and the tower acting as a reflector. Case B is the tower being fed with the sloper acting as a reflector. See text for details.

equivalent effective diameter) with a 2-mm-OD wire measuring 40.5 meters, sloping from the top of the tower with a length of insulated rope to the ground point, 27 meters away from the tower base. This is an appropriate distance for the guy-wire anchor points for a 40-meter high tower. The top of the sloping wire is 4.7 meters from the tower. These dimensions are valid only for conductors of the diameter indicated. The resonant frequency of the vertical conductor by itself is 1.78 MHz, and the sloping wire is resonant at 1.822 MHz. These data make it possible to duplicate the array with conductors of different diameters. All you have to do is to dip the wires to the listed frequencies.

With the tower fed, the wire acts as a perfect reflector, giving 3.4 dB gain over the tower by itself, and a useful 17 dB of F/B at 1.83 MHz. With the tower grounded, and feeding the sloping wire, the array now shoots in the opposite direction. The tower now acts as a director, and the gain is about the same (3.5 dB), with a F/B of 15 dB on 1.83 MHz. **Fig 13-54** shows the vertical patterns of these arrays, as compared to the tower by itself. Modeling was done over average ground, and a perfect radial system was assumed for both conductors.

Fig 13-55 shows the main performance data for both configurations. In most practical cases, however, you would probably try to use the sloping wire as either a director or as a reflector, in which case the sloping wire would need to be tuned with a reactive element (L or C). Using the same sloping wire (dimensions, placement) I now tuned it by a series capacitor ($X_C = -j\ 50\ \Omega$) for best performance. With less than 4 dB of

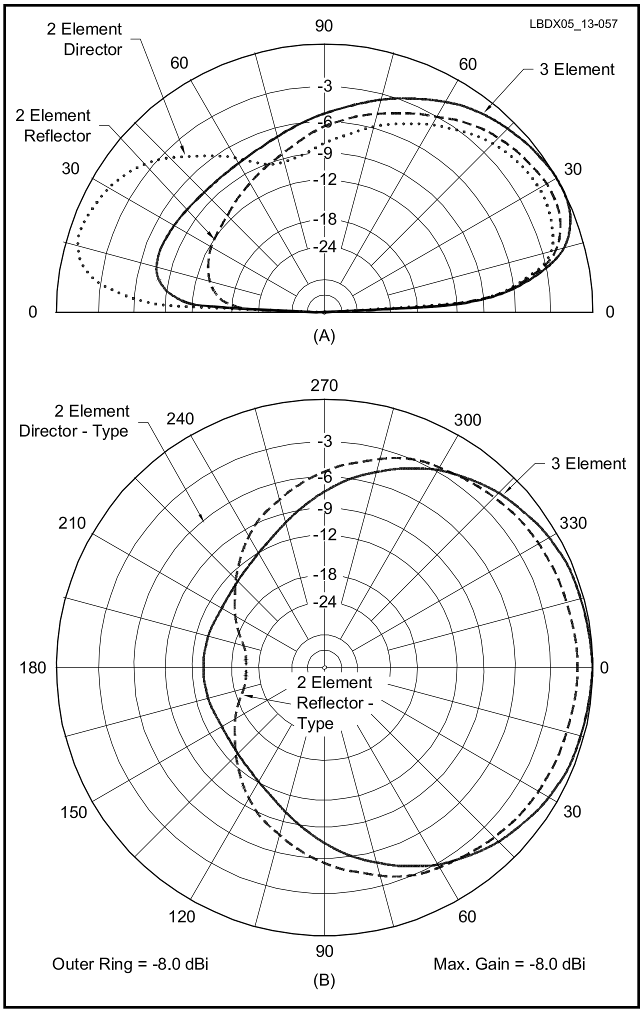


Fig 13-57 — Vertical and horizontal radiation patterns (over excellent ground) of three types of slant-wire parasitic arrays: reflector parasitic, director parasitic and reflector plus director parasitic. These patterns are valid for parasitic elements spaced 0.18 λ from the driven element (at their bases), which appears to be a typical situation for guy wires on a 40-meter guyed tower. The horizontal patterns are for an elevation angle of 20°.

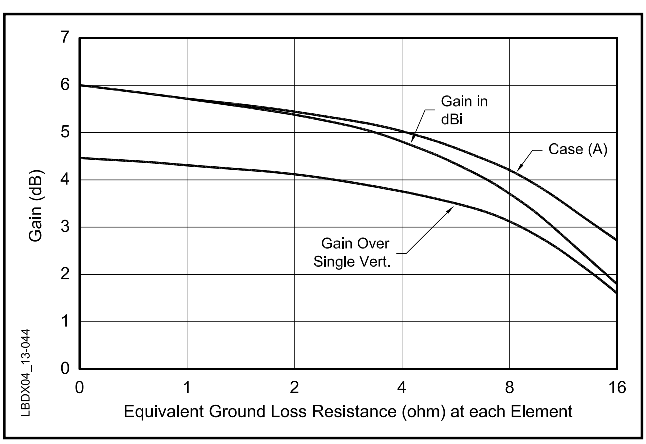


Fig 13-58 — Gain of the 3-element guy-wire array as a function of the equivalent ground-loss resistance. Case A is for a driven element with a fixed 1-Ω loss resistance.

F/B the gain was 2.9 dB over the vertical by itself. Not a very spectacular result. This was also reported by J. Stanley, K4ERO (Ref 982). The gain obtained is also almost 1 dB less than for the reflector case. Note that the results of this configuration are far inferior to those obtained when feeding the sloping wire and using the tower as a director (see **Fig 13-56**).

It is obvious that we must also keep the driven element-reflector configuration when switching directions. In this case, where we feed the tower or the sloping wire alternatively, it appears that perfect directivity and pattern reversal can be obtained without need of a capacitor or inductor. The feed point impedance, in the case of the fed tower, is 43 + j 62 Ω on 1.83 MHz. When the sloping wire is fed, its feed impedance on 1.83 MHz is also 42 + j 62 Ω. In both cases a small L-network (or just a series capacitor, if you can live with about 1.3:1 SWR at the design frequency) should be used to match the antenna to a 50 Ω feed line.

Most of the 40-meter tall towers used as 160-meter λ/4 verticals are probably guyed in four directions. That means that we probably can hang four sloping wires from the top. This is the next case I investigated.

3.9.1.2. A Guy Wire Array Using a Reflector and a Director

Continuing with the same physical configuration (the guy cables being anchored approximately 29 meters from the base of the 40-meter tower), the combination of using both a director as well as reflector was obvious. This combination can typically boost the gain another 1 dB, but has the disadvantage of reducing the F/B by about 6 dB.

Tuning the director with a series reactance of –j 70 Ω is a compromise solution that does not yield maximum gain, but still yields a more or less acceptable F/B ratio. This compromise was described in detail by Christman (Ref 389). The resonant frequencies of the parasitic elements, when fully decoupled from one another and from the driven element are: director, 1.952 MHz and reflector, 1.822 MHz.

Fig 13-57 shows the radiation patterns for three types of slant-wire parasitic arrays.

3.9.1.3. The Ground System

It is clear that for all of the above arrays it is important to have a good ground radial system, not only for the driven element but also for the parasitic elements. **Fig 13-58** shows the gain of the array as a function of the equivalent ground loss resistance. Case A is for a radiator with an almost-perfect ground radial system (= 1 Ω) but for varying ground loss resistances at the parasitic elements. Whereas 1-Ω ground systems yield a gain of 5.6 dBi for the array, the gain drops to 3.67 dBi if all elements have an equivalent ground loss of 8 Ω. If we have a 1-Ω loss resistance for the radiator, but a rather mediocre loss resistance of 8 Ω for the parasitic elements, the gain drops to 4.13 dBi. This shows that there really is very little room for a poor ground system under the parasitic elements too. It is obvious that it makes no difference whatsoever how the driven element is fed, series or parallel.

3.9.2. Three-Element Vertical Parasitic Array

The arrays I analyzed in the last section all showed rather substantial high-angle radiation, which is caused by the horizontal component of the sloping parasitic wires. To improve

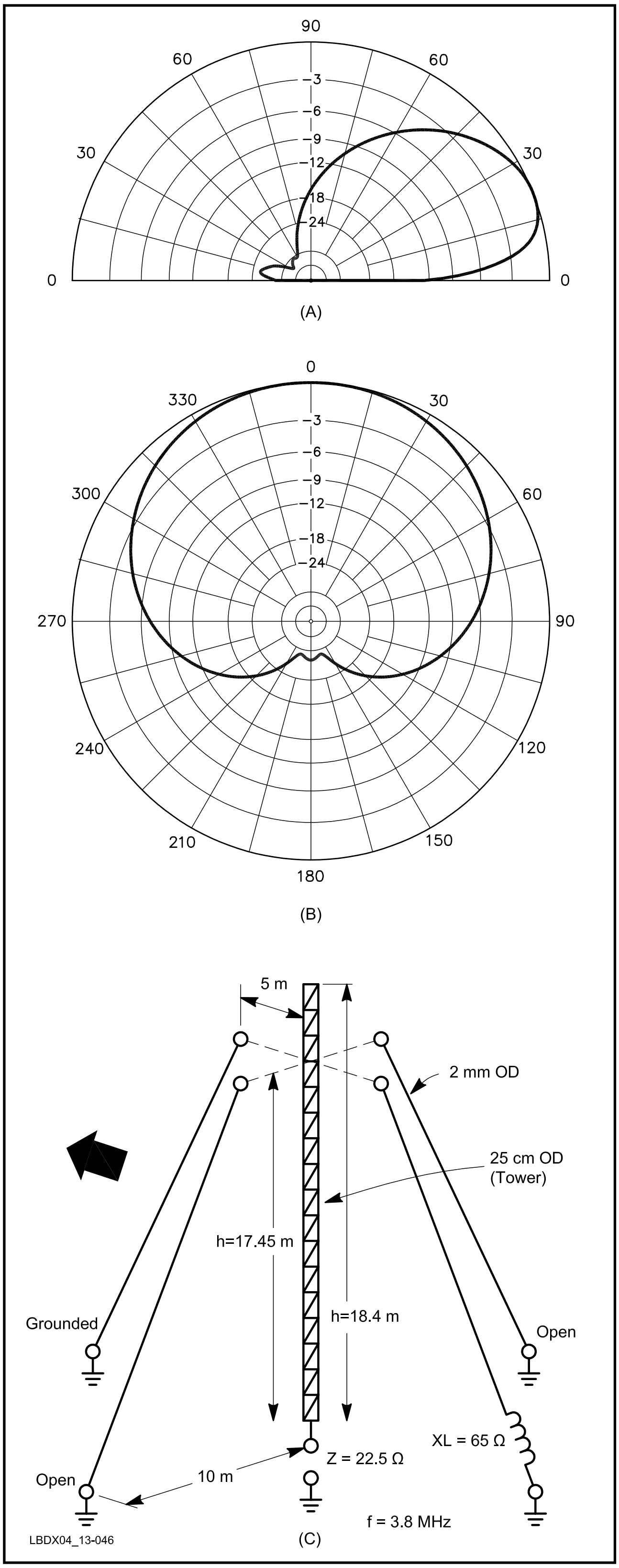


Fig 13-59 — Three-element parasitic array, consisting of a central support tower with two support cross-arms mounted at 90° near the top. Two of the sloping wires are left floating, a third one is grounded as a director, and the fourth one is loaded with a coil to act as a reflector. The azimuth pattern at B is taken for a takeoff angle of 22°. Radials are required on all five ground points but they have been omitted on this drawing for clarity.

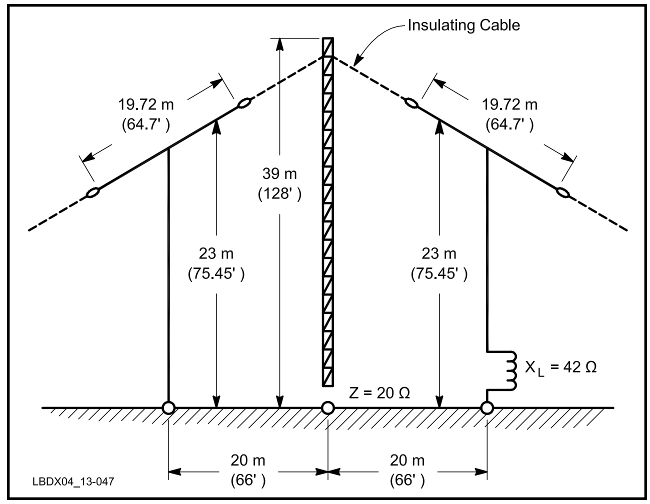


Fig 13-60 — This 160-meter 3-element parasitic array produces 4.8 dB gain over a single vertical, and better than 25 dB F/B over 30 kHz of the band centered around 1832 kHz. With such an array there is no need for Beverage receiving antennas! The drawing shows only two of the four parasitic elements. The two other elements are left floating.

on that situation we can try to bring the sloping elements as vertical as possible. If you do not use the parasitic-element wires as guy wires for the tower, you can consider the configuration shown in **Fig 13-59**.

The four cross arms mounted at the top of the tower support four sloping wires. The tower itself is a quarter-wave vertical. The bases of the parasitic elements are 0.125 λ away from the driven element. All four sloping wires are dimensioned to act as directors. When used as a reflector, a parasitic element is loaded with a coil at the bottom. The two sloping elements off the side are left floating. This array has a very respectable gain of 4.5 dB over a single vertical. At the main elevation angle the F/B ratio is an impressive 30 dB, as can be seen from the patterns in Fig 13-59.

You can "grid dip" the sloping wires to tune them. You can of course also use your antenna or network analyzer, which will give more accurate results. Make sure the driven element as well as the other three sloping wires are left floating when dipping a parasitic element.

The resonant frequency of the director should be 4.055 MHz (for f_{design} = 3.8 MHz). You must, of course, dip the reflector wires with the loading coil in place to find the resonant frequency. The resonant frequency for the reflector element is 3.745 MHz (for f_{design} = 3.8 MHz).

If you want to totally eliminate the horizontal high angle radiation component, you can hang top-loaded elements from catenary cables, as shown in **Fig 13-60**. In this 160-meter (f_{design} = 1.832 MHz) version the parasitic elements are top loaded with sloping T-shaped wires. These wires can be supported along the catenary support cable or may be an intrinsic part of the support structure.

Such slanted top-loading wires do not produce far-field horizontal radiation because they are symmetrical with respect to the vertical wire.

To make this an array that can be switched in four directions, you should slope four catenary cables at 90° increments

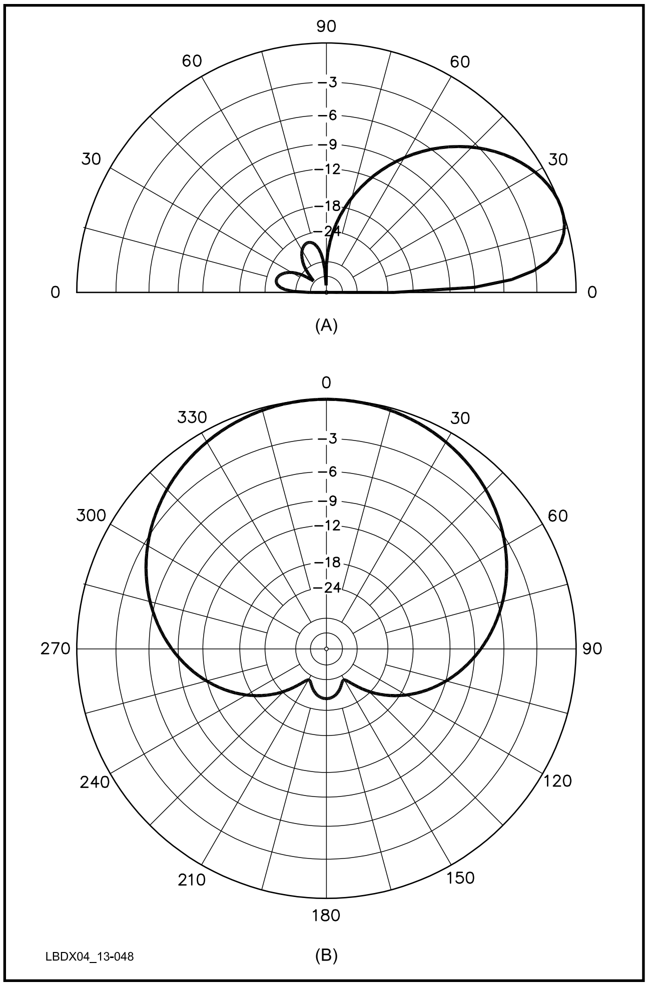


Fig 13-61 — Horizontal and vertical radiation patterns for a 3-element vertical parasitic array for 160 meters. The azimuth pattern at B is for an elevation angle of 20°. Note the excellent pattern and F/B for the array.

Fig 13-62 — Base insulated feed point of the 3-element K3LR vertical Yagi for 160 meters. The aluminum box contains an L network to transform the array impedance (about 25 Ω) to the 50 Ω Hardline impedance (1⅝ inch feed line!). The plastic box distributes the relay control wiring to the parasitic element relays. The coaxial cable coil to the right is a λ/4 shorted-stub that serves as a static drain and also provides attenuation of the 80-meter harmonic. This is very important at a multi-operator contest station! Note the 120 radials connected to the annular ring.

Fig 13-63 — A metal box contains the T-200-2 loading coil for the reflector mode and two small relays to switch the coil in and out of the circuit at each of the four parasitic element bases. The base of each parasitic element is also equipped with a λ/4 coaxial stub and complemented with no less than 120 λ/4 radials.

from the top of the driven-element tower. With the appropriate hardware you can connect the parasitic elements directly to ground to serve as a director, to ground via a loading coil to serve as a reflector. Or you can leave the element floating, with the unused elements off the side of the firing direction. It's a good idea to provide a position in your switching system to have all elements floating, in which case you will have an omnidirectional antenna. This may be of interest for testing the array, or for taking a quick listen around in all directions.

The example I analyzed uses 23-meter-long vertical parasitic elements. Each is top-loaded with a 19.72-meter long sloping top wire that is part of the support cable. As the length of the top-loading wire is the same on both sides of the loaded vertical member, there is no horizontally polarized radiation from the top-loading structure.

The four parasitic elements are dimensioned to be resonant (as directors) at 1.935 MHz. The same procedure as explained above for the 80-meter array can be used to tune the parasitic elements. When used as a reflector, the elements are tuned to be resonant at 1.778 MHz. This can be done by installing an inductance of 3.65 µH (reactance = 42 Ω at 1.832 MHz) between the bottom end of the parasitic element and ground. The radiation resistance of this array is around 20 Ω, and it has a gain of 4.8 dB over a single full-size vertical. **Fig 13-61** shows the radiation patterns of this array. The bandwidth behavior is excellent. The array shows a constant gain over more than 50 kHz and better than 25 dB F/B over more than 30 kHz. When tuned for a 1:1 SWR at 1.832 MHz, the SWR will be

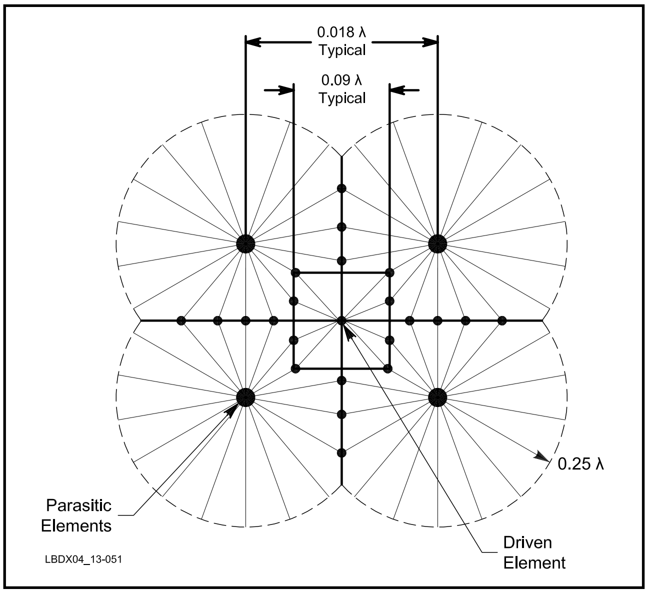


Fig 13-64 — Radial layout used at K3LR on his 3-element 160-meter parasitic array. A total of 15 km of wire is used for radials in this array!

less than 1.2:1 from 1.820 to 1.850 kHz. This really is a winner antenna, and it requires only one full-size quarter-wave element, plus a lot of real estate to run the sloping support wires and the necessary radials.

The same principle with the sloping support wires and the top-loaded parasitic elements could, of course, be used with the 80-meter version of the 3-element vertical parasitic array. Tim Duffy, K3LR, made an almost exact copy of this array, after initially having used inverted Ls for the parasitic elements. Tim recognized that these inverted-L elements introduced a fair amount of horizontally polarized high-angle radiation. For a single-element vertical, this may be of very little importance, but for a parasitic element of an array, this will greatly reduce the directivity of the array, especially at high angles This can be important if the array is also used for receiving. Tim reported changing from inverted-L shaped elements to the sloping-T shaped elements and reported that the T-shaped elements work much better.

At K3LR the parasitic directors were resonated at 1.903 kHz, and the loading coils were 4.0 μH with a vertical length of 19.58 meters and a sloping-T-shaped top hat of 17.78 meters (all made of #12 Copperweld wire). The array is matched to a 75 Ω coaxial feed line with an L network (see **Fig 13-62**). The measured SWR is 1.3:1 on 1.8 MHz, 1:1 on 1.83 MHz and 1.3:1 on 1.85 MHz. K3LR reports about 5 dB of gain and 30 dB of F/B at the design frequency (1.83 MHz). At 1.82 MHz the measured F/B is still 25 dB and at 1.84 MHz Tim measured 15 dB.

Tim has an omnidirectional mode, where he floats all parasitic elements. See **Fig 13-63**. Tim can't run many Beverage antennas at his location. He has just one 1200 footer on Europe. He appreciates the excellent directivity for his 160-meter array on receive. Don't forget that in order to make such an array work correctly, you need an impressive radial system under each of the elements. K3LR uses not less than 15 km of radials in this system!

A recommended radial layout, which K3LR uses, is shown in **Fig 13-64**. Note that radials are even more important for a parasitic array than for an all-driven phased array. With parasitic arrays the gain seems to suffer even more quickly from poor ground systems, so a good radial system is mandatory.

Incidentally, computer modeling indicates that elevated radials do not work well with parasitic arrays. No matter what kind of elevated radial system I modeled (different numbers of radials, varied lengths, and different orientations), the result was a badly distorted pattern. This is logical in view of the influence of the capacitive coupling of the raised resonant radials. Compare this situation with what was experienced with the top-loaded 80-meter Yagis designed by W7CY and K6UA. With phased arrays using current-forcing methods, the feed method itself is responsible for overcoming the effects of mutual coupling due to the proximity of the wires.

3.9.3. The N7JW – K7CA Array with Parasitic Elements

In Chapter 7, Section 1.32 I covered "parasitic receiving arrays." Since you cannot make parasitic receiving arrays with lossy elements, such arrays are equally good transmitting arrays! The various parasitic arrays that were built in the Utah desert really put N7JW and K7CA in the front row when the show is on to Europe on Top Band.

3.9.4. The K1VR/W1FV Spitfire Array

Fred Hopengarten, K1VR, and John Kaufmann, W1FV, designed a somewhat novel 3-element parasitic array, which they called the "Spitfire Array." **Fig 13-65** shows the layout of the antenna. John, W1FV, described the array as a 3-element parasitic array with a vertical tower as the driven element and two sloping-wire parasitics, one a director and one a reflector. He adds that the parasitic elements are not grounded and do not require a separate radial system of their own. Rather the parasitic wires are folded around to achieve the required λ/2 wave resonances. The tower driven element does have its own radial system.

The antenna was modeled with *EZNEC*, using the *MININEC* ground analysis method (the *NEC-2* ground analysis method cannot be used because the driven element is a grounded element). Over average ground, the model shows a gain of

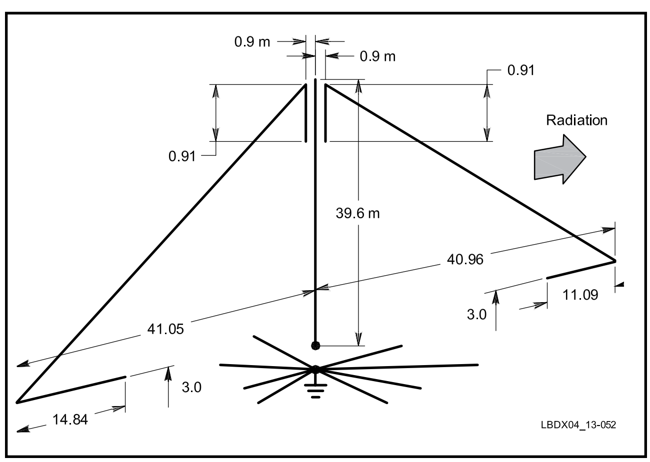


Fig 13-65 — Layout and dimensions for the 160-meter Spitfire array.

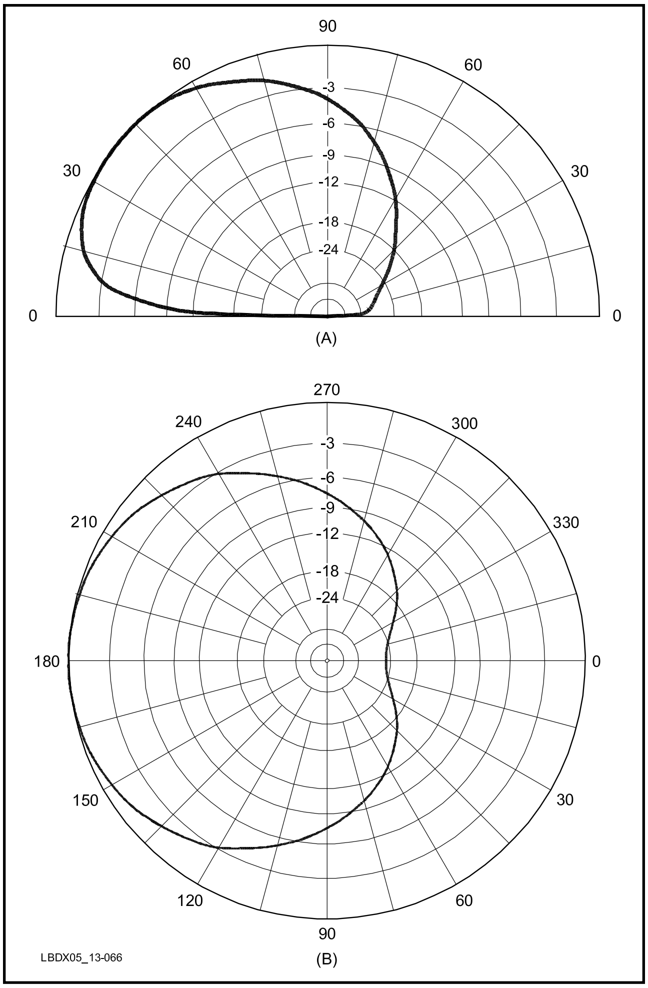


Fig 13-66 — Vertical and horizontal radiation patterns for the Spitfire array. The horizontal pattern was calculated for an elevation angle of 23°.

4.8 dB over a single vertical at a takeoff angle of 23°.

The antenna has a substantial amount of high-angle radiation and its pattern resembles that of a EWE antenna (see **Fig 13-66**). The high angle radiation is mainly caused by the radiation of the bottom half of the sloping half-wave parasitic elements. You can consider the bottom half of each parasitic element as a single radiating radial that is bent backward toward the driven element.

To change directions, you use a relay to switch in or out an additional wire segment on the lower horizontal portion of each parasitic to change from director to reflector operation. It is important to point out that this switching happens at high voltage points, and a well-insulated vacuum relay is certainly no luxury.

A model for four switching directions can be constructed by adding an identical set of parasitic wires (for a total of four wires) oriented at right angles to the original two. Only two wires at a time are active. The other two are detuned so they don't couple. Simply grounding them appears to accomplish this.

3.9.4.1. Critical Analysis

If you compare this array with the classic 3-element parasitic array as described in Section 3.7, you will notice that the main difference is that this Spitfire array claims not to require radials for the parasitic elements. We learned in Chapter 9 about ground losses and radial systems for vertical antennas. I also pointed out that the Spitfire array has been modeled with a *MININEC*-based modeling program, which means that a perfectly conducting ground is assumed in the near field of the antenna. This is certainly not true in real life.

The bottom half of the sloping parasitic elements are very close to ground, and undoubtedly will cause a great deal of near-field absorption losses in the lossy ground, unless the ground is hidden from these low wires by an effective ground screen. The model used to develop this antenna does not take any of this into account. What does that mean? It does not mean that the antenna will not be able to give good directivity. But it means that the quoted gain figures are probably several dB higher than what can be accomplished in real life, if no radial or ground screen system is used that effectively screens the lossy ground under the array from the antenna. This could be accomplished by using extra long radials on the driven element that extend at least λ/8 beyond the tips of the parasitic elements. This would mean radials that are at least 60 meters long, with their tips separated not more than 0.015 λ (see Chapter 11). This means that 157 radials, each 60 meters long, fulfill this requirement. Only under these circumstances will we achieve the same gain as with ground-mounted quarter-wave parasitic elements, each using an elaborate radial system (as with the K3LR antenna).

I modeled the same antenna, but using λ/4 grounded parasitic elements, and maintained the same average spacing from the tower. The sloping parasitic elements are both 38.7 meters long, spaced 9.2 meters from the tower at the top and 29 meters from the tower at the bottom. The tops of the parasitic elements are at 33 meters, which is 6 meters below the top of the 39 meter high tower. The directors can be tuned to become reflectors by loading them with a coil with an inductance of 2.85 µH. This classic antenna has 0.7 dB more gain than the Spitfire over a perfect ground, and has a F/B and bandwidth that is comparable to what's been calculated for the Spitfire antenna. Most important is that the antenna does not show the high-angle radiation associated with the Spitfire array (see **Figs 13-67** and **13-68**).

3.9.4.2. Conclusion

Modeling tools are fantastic, but they are tools. Each tool has its limitations. As users of these tools, we should be aware of their limitations and know how to handle them. Elevated radials, half-wave parasitic elements, voltage-fed verticals, etc, do not have any magic properties. They do not make real ground vanish. It's still there, and if it's close to any radiating wires, it will cause losses, what we call the near-field absorption losses.

Modeling programs based on *MININEC* use a perfect near-field ground, which means that the results from those models do not take into account these real-world losses. If you would use a *NEC-4* based modeling program, where you can ground the driven element, you would likely still arrive at gain figures that are too high. This is because of a widely recognized flaw in *NEC* that results in too-low near-field losses for wires that are close to ground (see Chapter 9).

All parasitic vertical arrays with grounded elements suffer from the drawback that the real gain rapidly diminishes when the resistive connection loss of an imperfect ground system is considered. This can easily be modeled on a *MININEC*-based program by inserting a small resistor in series with the elements

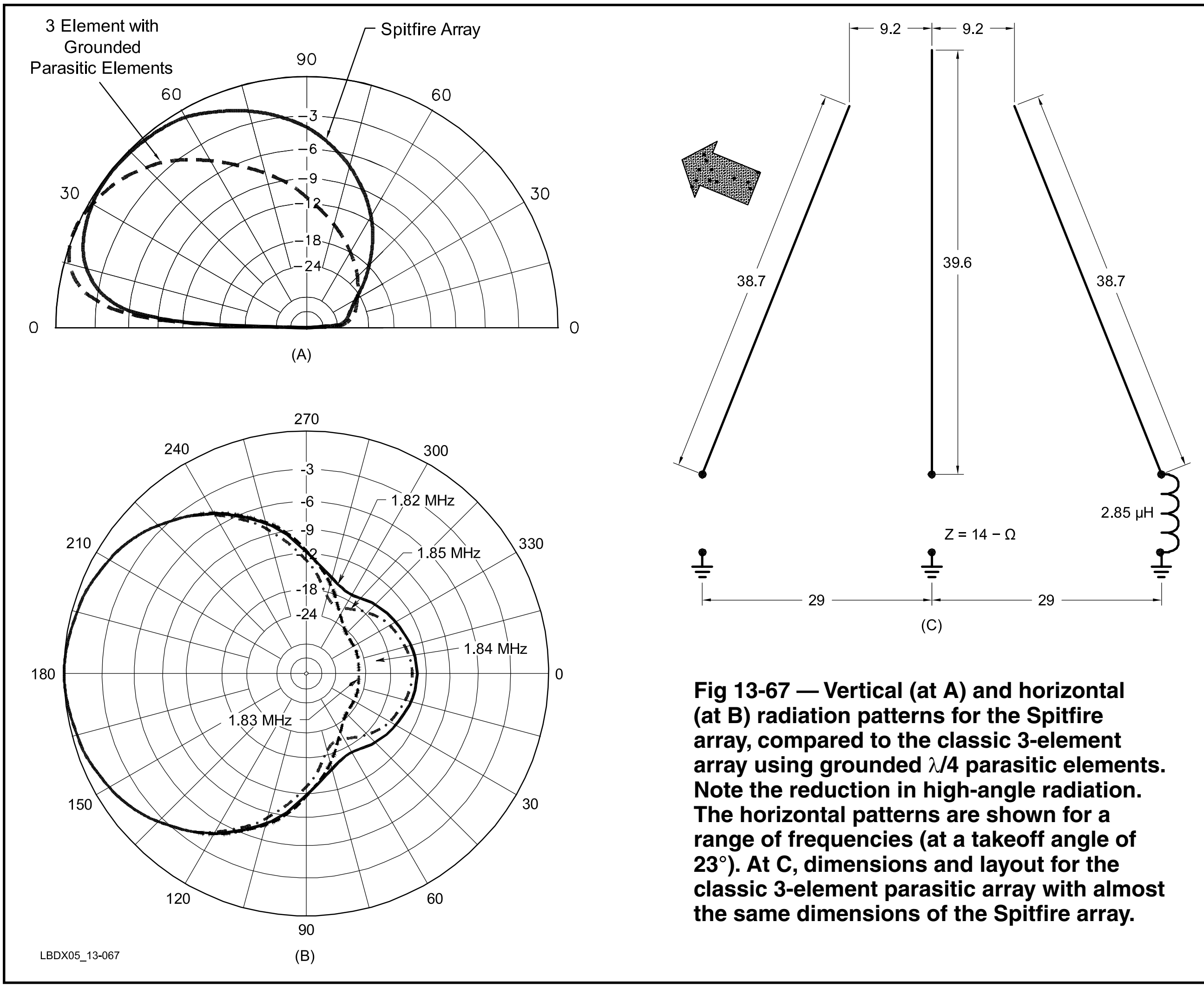


Fig 13-67 — Vertical (at A) and horizontal (at B) radiation patterns for the Spitfire array, compared to the classic 3-element array using grounded λ/4 parasitic elements. Note the reduction in high-angle radiation. The horizontal patterns are shown for a range of frequencies (at a takeoff angle of 23°). At C, dimensions and layout for the classic 3-element parasitic array with almost the same dimensions of the Spitfire array.

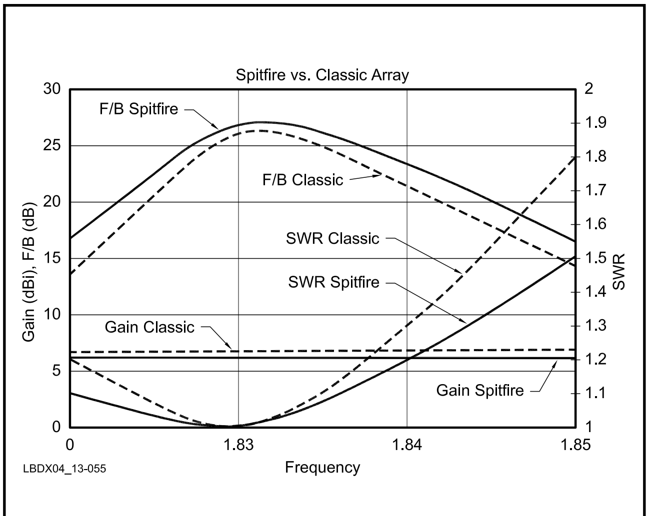


Fig 13-68 — Performance characteristics of the Spitfire array compared to a classic 3-element parasitic array of essentially the same dimensions (Fig 13-67). Note that the gain for the Spitfire can only be achieved with an extensive radial or ground-screen, which is mandatory to prevent several dB of near-field ground-absorption losses of the half-wave elements that are very close to ground.

at their connection to ground. Not having a direct ground connection for the parasitic elements (as in the Spitfire) does not mean that there are no ground-related losses. The losses here are the near-field absorption losses, associated with low-to-the-ground wires, and these cannot be properly modeled with today's modeling tools. This does not detract from the fact that they are there, and can account for several dB of signal loss!

The Spitfire is an array that has its merits. The extra high-angle radiation may be an asset under certain circumstances, such as in contesting where some extra local presence is welcome. Potential builders should know that a good ground screen is as essential with this antenna as it is for an array using grounded near-quarter-wave parasitic elements. Sorry, but again, there is simply no free lunch!

3.10. Yagi Matching Systems

The matching systems for Yagis I describe in this section are not only valid for the low bands. They work on higher frequencies just as well. I will cover the concept, design and realization of various popular matching systems used with Yagis, including:

- Gamma match
- Omega match
- Hairpin match
- Direct feed

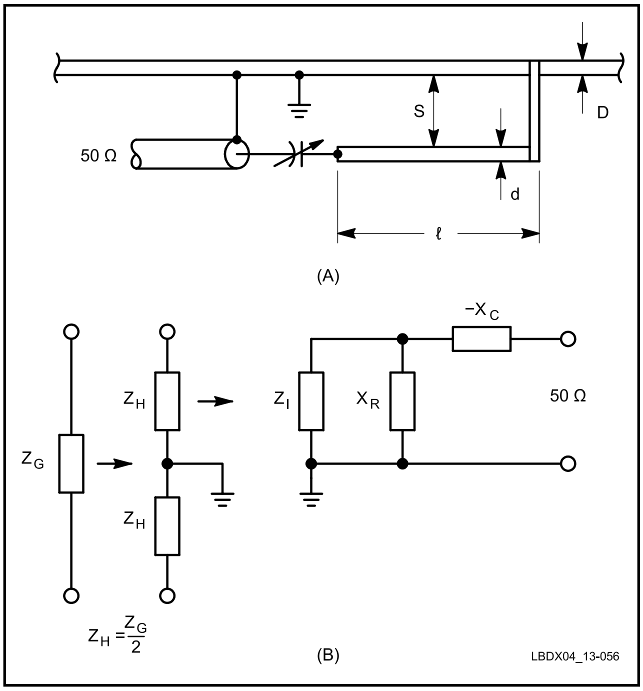


Fig 13-69 — Layout and electrical equivalent of the Gamma match.

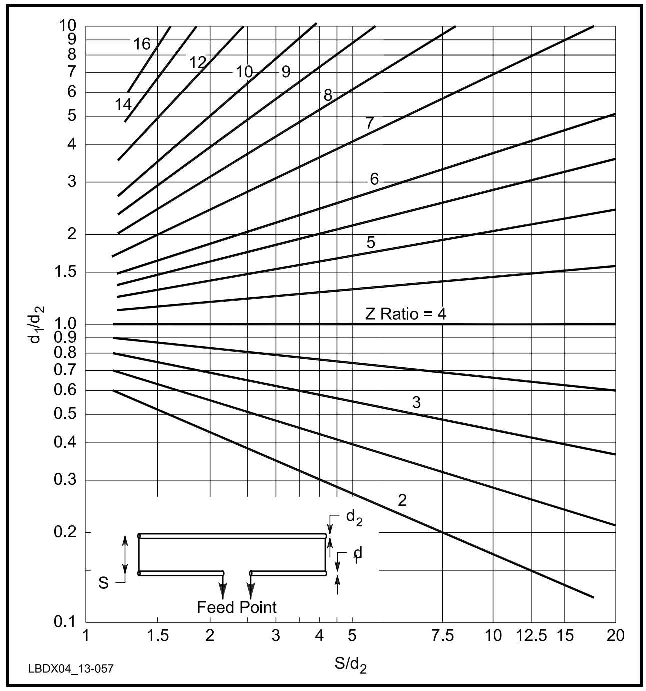


Fig 13-70 — Step-up ratio for the Gamma and Omega matches as a function of element diameter (d2), rod diameter (d1) and spacing (S). (After *The ARRL Antenna Book*.)

My *Yagi Design* software contains modules that make it possible to design these matching systems with no guesswork.

3.10.1. The Gamma Match

In the past, Gamma-match systems have often been described in an over-simplifying way. A number of home-builders must have gone half-crazy trying to match one of W2PV's 3-element Yagis with a Gamma match. The reason for that is the low radiation resistance in that design, coupled with the fact that the driven-element lengths were too long as published. The driven element of the 3-element 20-meter W2PV Yagi has a radiation resistance of only 13 Ω and an inductive reactance of +18 Ω at the design frequency for the published radiator dimensions of 0.489661 λ (Ref 957). Yagis with such low radiation resistance and a positive reactance cannot be matched with a Gamma (or an Omega) match. It is necessary to shorten the driven element to introduce the required capacitive reactance in the feed-point impedance!

Yagis with a relatively high radiation resistance, say 25 Ω, or with some amount of capacitive reactance, typically –10 Ω, can easily be matched with a whole range of Gamma-match element combinations.

Fig 13-69 shows the electrical equivalent of the gamma match. Z_g is the element impedance to be matched. The gamma match must match the element impedance to the feed-line impedance, usually 50 Ω. The step-up ratio of a Gamma match depends on the dimensions of the physical elements (element diameter, Gamma-rod diameter and spacing) making up the matching section. **Fig 13-70** shows the step-up ratio as a function of the driven-element diameter, the gamma rod diameter and spacing between the two.

The calculation involves a fair bit of complex mathematics, but software tools have been made available from different sources to solve the Gamma-match problem. The *Yagi Design* software addresses the problem in one of its modules (Matching Systems), as does *YW* (*Yagi for Windows*, supplied with late editions of *The ARRL Antenna Book*).

To illustrate the matching problems evoked above, I have listed the gamma-match element variables in **Table 13-14** for a Yagi with R_{rad} = 25 Ω, and in **Table 13-15** for a Yagi with R_{rad} = 15 Ω.

Table 13-14 shows that a Yagi with a radiation resistance of 25 Ω can easily be matched with a wide range of Gamma-match parameters, while the exact length of the driven element is not at all critical. It is clear that short elements (negative reactance) require a shorter Gamma rod and a slightly smaller value of Gamma capacitor.

Table 13-15 tells the story of a high-Q Yagi with a radiation resistance of 15 Ω, similar to the 3-element W2PV or W6SAI Yagis. If such a Yagi has a "long" driven element, a match cannot be achieved, not even with a step-up ratio of 15:1. With this type of Gamma (step-up = 15:1), the highest positive reactance that can be accommodated with a radiation resistance of 13 Ω is approximately +12 Ω. In other words, it is simply impossible to match the impedance (13 + *j* 18 Ω) of the W2PV 3-element 20-meter Yagi with a Gamma match without first reducing the length of the driven element.

The first thing to do when matching a Yagi with a relatively low radiation resistance is to decrease the element length to introduce capacitive reactance, perhaps –15 Ω in the driven-element feed-point impedance. How much shortening is needed (in terms of element length) can be derived from **Fig 13-71**. Table 13-15 shows that an impedance of 15 – *j* 15 Ω can be easily matched with step-up ratios ranging from 5 to 8:1.

Several Yagis have been built and matched with Gamma systems, calculated as explained above. When the reactance of the driven element at the design frequency was exactly known, the computed rod length was always right on. In some cases the

Table 13-14

Gamma-Match Element Data for a Yagi with a Radiation Resistance of 25 Ω

Rod *d*	*S*	*Step up Rat.*	*–20 Ω* *L*	*C*	*–15 Ω* *L*	*C*	*–10 Ω* *L*	*C*	*–5 Ω* *L*	*C*	*0 Ω* *L*	*C*	*+5 Ω* *L*	*C*
0.5	5.0	5.28	118	350	123	502	138	614	171	734	231	396	317	734
	4.0	5.42	131	342	135	488	151	592	184	700	255	376	331	700
	3.0	5.65	152	332	155	468	172	562	207	656	267	349	351	654
	2.5	5.83	169	324	172	452	189	540	224	634	285	332	369	624
0.38	5.0	5.87	119	322	121	450	133	536	158	618	203	328	269	618
	4.0	6.08	132	314	133	434	145	514	171	588	216	311	281	584
	3.0	6.43	153	302	154	412	165	482	192	548	238	288	302	558
	2.5	6.71	170	292	169	396	188	462	208	520	255	273	319	520
0.25	5.0	6.75	120	290	119	394	128	458	147	516	181	270	230	516
	4.0	7.07	133	282	131	374	140	430	158	482	192	251	239	482
	3.0	7.62	154	268	151	356	160	408	179	452	213	236	262	452
	2.5	8.06	172	258	167	340	175	398	195	428	230	223	278	428

Design parameters: D = 1.0; Z_{ant} = 25 Ω; Z_{cable} = 50 Ω. The element diameter is normalized as 1. Values are shown for a design frequency of 7.1 MHz. L is the length of the Gamma rod in cm, C is the value of the series capacitor in pF. The length of the Gamma rod can be converted to inches by dividing the values shown by 2.54.

Table 13-15

Gamma-Match Element Data for a Yagi with a Radiation Resistance of 15 Ω

Rod *d*	*S*	*Step up Rat.*	*–20 Ω* *L*	*C*	*–15 Ω* *L*	*C*	*–10 Ω* *L*	*C*	*–5 Ω* *L*	*C*	*0 Ω* *L*	*C*	*+5 Ω* *L*	*C*
0.5	5.0	5.28	93	410	92	586	116	1180	—	—	—	—	—	—
	4.0	5.42	103	400	102	566	121	1074	—	—	—	—	—	—
	3.0	5.65	120	386	117	538	136	948	—	—	—	—	—	—
	2.5	5.83	134	376	131	518	123	874	—	—	—	—	—	—
0.38	5.0	5.87	94	372	91	514	130	860	—	—	—	—	—	—
	4.0	6.08	104	362	101	494	113	996	206	3906	—	—	—	—
	3.0	6.43	121	346	117	466	128	716	208	1680	—	—	—	—
	2.5	6.71	136	334	130	446	140	666	210	1306	—	—	—	—
0.25	5.0	6.75	96	334	91	442	99	660	147	1268	—	—	—	—
	4.0	7.07	106	322	101	424	107	614	152	1060	376	1268	—	—
	3.0	7.62	123	304	117	396	122	556	161	864	309	1188	—	—
	2.5	8.06	138	292	131	376	135	518	172	766	295	982	—	—

Design parameters: D = 1.0; Z_{ant} = 15 Ω; Z_{cable} = 50 Ω. The element diameter is normalized as 1. Values are shown for a design frequency of 7.1 MHz. C is expressed in pF; L in cm (divide by 2.54 to obtain inches). Note there is a whole range where no match can be obtained. If sufficient negative reactance is provided in the driven-element impedance (with element shortening), there will be no problem in matching Yagis even with a low radiation resistance.

series capacitor value turned out to be smaller than calculated. This is caused by the stray inductance of the wire connecting the end of the gamma rod with the plastic box containing the gamma capacitor, and the wire between the series capacitor and the coaxial feed line connector. The inductance of such a wire is not at all negligible, especially on the higher frequencies. With a pure coaxial construction, this should not occur.

A coaxial gamma rod is made of two concentric tubes, where the inner tube is covered with a dielectric material, such as heat-shrink tubing. The length of the inner tube, as well as the material's dielectric and thickness, determine the capacitance of this coaxial capacitor. Make sure to properly seal both ends of the coaxial gamma rod to prevent moisture penetration.

Feeding a symmetric element with an asymmetric feed system has a slight impact on the radiation pattern of the Yagi. The forward pattern is skewed slightly toward the side where the gamma match is attached, but only a few degrees, which is of no practical concern. The more elements the Yagi has, the less the effect is noticeable.

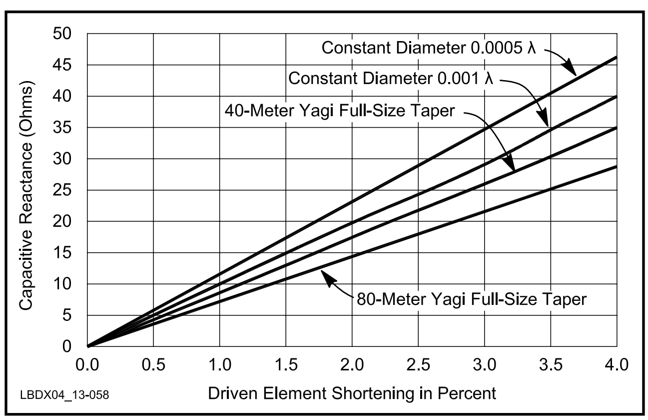


Fig 13-71 — Capacitive reactance obtained by various percentages of driven-element shortening. The 40-meter full-size taper is the taper described in Table 13-1. The 80-meter taper is that for a gigantic Yagi using latticed-tower elements, varying from 42 cm at the boom down to 5 cm at the tips.

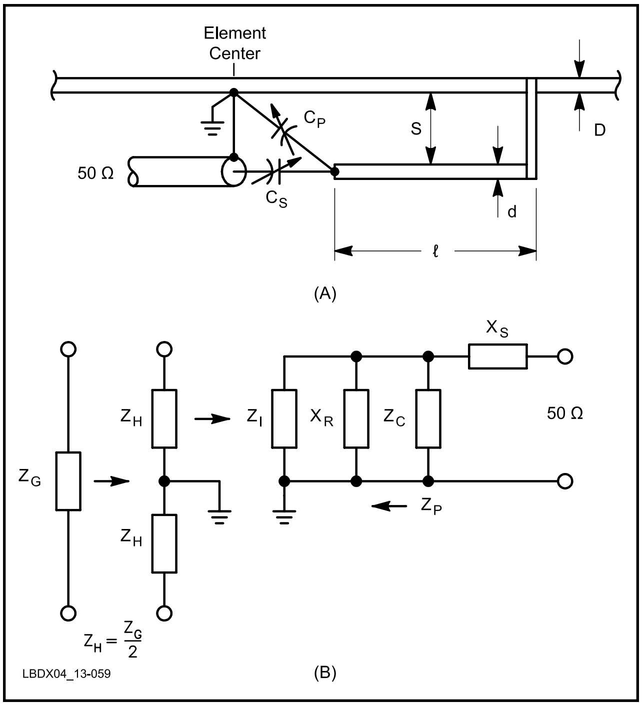


Fig 13-72 — Layout and electrical equivalent of the Omega match.

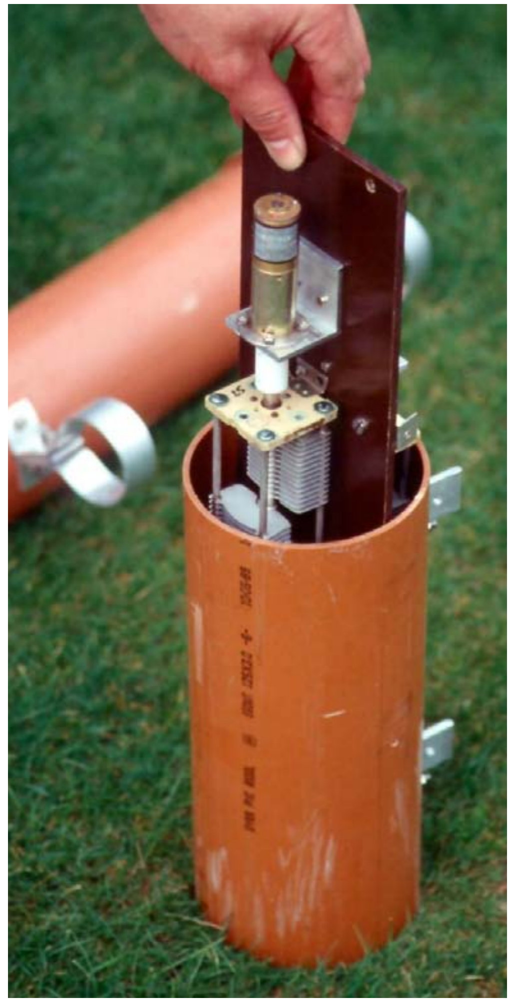


Fig 13-73 — Motor-driven Omega matching unit. The two capacitors with their dc motors and gear boxes are mounted in-line on a piece of insulating substrate material. This can then be slid inside the housing, which is made of stock lengths of PVC water drainage pipe. The PVC pipe is easily cut to the desired length. The round shape of the housing also has an advantage so far as wind loading is concerned.

The voltage across the series capacitor is quite small even with high power, but the current rating must be sufficient to carry the current in the feed line without warming up. For a power of 1500 W, the current through the series capacitor is 5.5 A (in a 50 Ω system) The voltage will vary between 200 and 400 V in most cases. This means that moderate-spacing air-variable capacitors can be used, although it is advisable to over-rate the capacitors, since slight corrosion of the capacitor plates normally caused by the humidity in the enclosure will derate the voltage handling of the capacitor.

3.10.2. The Omega Match

The Omega match is a sophisticated Gamma match that uses two capacitors. Tuning of the matching system can be done by adjusting the two capacitors, without having to adjust the rod length. **Fig 13-72** shows the Omega match and its electrical equivalent. Comparing it with the Gamma electrical equivalent of Fig 13-69 reveals that the extra parallel capacitor, together with the series capacitor, now is part of an L network that follows the original Gamma match.

Again, the mathematics involved are complex, but the Matching section of the *Yagi Design* software will do the job in a second. From a practical point of view the Omega match is really unbeatable. The ultimate setup consists of a box containing the two capacitors, together with dc motors and gear-reductions. **Fig 13-73** shows the interior of such a unit using surplus capacitors and dc motors from a flea market. This system makes the adjustment very easy from the ground, and is the only practical solution when the driven element is located away from the center of the antenna.

The remarks given for the Gamma capacitors as to the required current and voltage rating are valid for the Omega match as well. The voltage across the Omega capacitor is of the same magnitude as the voltage across the Gamma capacitor, usually between 300 and 400 V, with a current of 2 to 4 A for a power of 1500 W.

3.10.3. The Hairpin or Beta Match

If we split the driven element (insulate it from the boom) we can use different types of direct feed systems. Probably the most popular one (especially with commercial manufacturers) is the Hairpin or Beta match system.

Let me first point out that the hairpin shaped inductor which is part of a so-called hairpin matching system on Yagis for the higher bands, is often replaced by a coil. A coil is much more compact and also has the advantage of having a much higher Q (if well designed and constructed) than a hairpin inductor. In what follows we will use coils as "hairpin" match inductors.

The feed-point impedance of a Yagi driven element that is about λ/2 long consists of a resistive part (the radiation re-

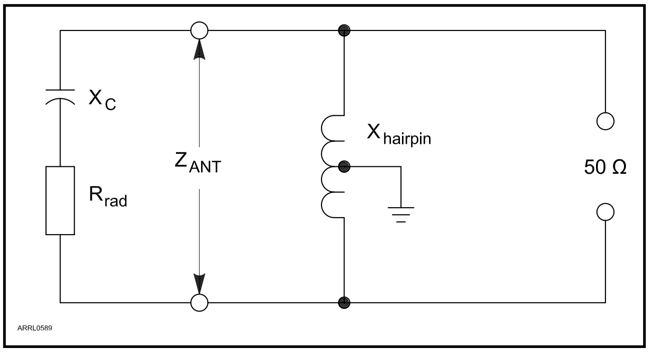


Fig 13-74 — Electrical equivalent of the hairpin match.

Table 13-16

Required Capacitive Reactance in Driven Element and in Hairpin Inductance

R_{rad} Ω	*Antenna Reactance* Ω	*Inductance Hairpin Coil* Ω
10.0	–20.0	25.0
12.5	–21.6	28.9
15.0	–22.9	32.7
17.5	–23.8	36.7
20.0	–24.5	40.8
22.5	–24.9	45.2
25.0	–25.0	50.0
27.5	–24.9	55.3
30.0	–24.5	61.2

Note: The feed-point impedance is 50 Ω.

Table 13-17

Values of Transformed Impedance and SWR for a Range of Driven-Element Impedances

Driven Element Impedance	*Hairpin Reactance, Ω*	*Resulting Impedance*	*SWR (vs 50 Ω)*
20 – *j* 20	40.8	40 – *j* 0.81	1.25:1
20 – *j* 24.5	40.8	50 – *j* 0	1.00:1
20 – *j* 25	40.8	51.2 + *j* 0.26	1.02:1
20 – *j* 30	40.8	64.5 – *j* 5.92	1.32:1

sistance) in series with a reactance. The reactance is positive if the element is longer and negative if the element is shorter than the resonant length. In practice, resonance never occurs at a physical length of exactly 0.5 λ, but always at a shorter length. With a hairpin matching system we deliberately make the element short, meaning that the feed-point impedance will be capacitive. **Fig 13-74** shows the electrical equivalent of the hairpin matching system.

If we connect an inductor across the terminals of a short driven element, we can now consider the series capacitor and the parallel inductor to be the two arms of an L network. This L network can be dimensioned to give a 50-Ω output impedance. The parallel inductor is commonly replaced with a short length of short-circuited open-wire feed line having the shape of a hairpin, and hence the matching system's name.

The radiation resistance of the feed-point impedance changes only slightly as a function of length if the element length is varied plus or minus 5% around the resonant length. The change in the reactance, however, is quite significant. The rate of change will be greatest with elements having smaller diameters (see Fig 13-71). The required hairpin coil reactance is given by:

$$X = 50 \times \sqrt{\frac{R_{rad}}{50 - R_{rad}}} \qquad \text{(Eq 13-5)}$$

The required driven element reactance is given by:

$$X_C = -\frac{4_{rad} \times 50}{X_{hairpin}} \qquad \text{(Eq 13-6)}$$

Table 13-16 shows the required values of capacitive reactance in the driven element, as well as the required reactance of the hairpin inductor for a range of radiation resistances.

The question now is how long the driven element must be to represent the required amount of negative reactance ($-X_C$). Fig 13-71 lists the reactance values obtained with several degrees of element shortening. Although the exact reactance differs for each one of the listed element diameter configurations, you can derive the following formula from the data in Fig 13-71.

$$X_C = -Sh \times A \qquad \text{(Eq 13-7)}$$

where

X_C= reactance of the element in Ω
Sh = shortening in % versus the resonant length
A = 8.75 (for a 40-meter full-size Yagi) or 7.35 (for an 80-meter full-size Yagi)

This formula is valid for shortening factors of up to 5%. The figures are typical and depend on the effective diameter of the element.

3.10.3.1. Design Guidelines for a Hairpin System

Most HF Yagis have a radiation resistance between 20 Ω and 30 Ω. For these Yagis the following rule-of-thumb applies: The required element reactance to obtain a 50-Ω match with a hairpin is approximately 25 Ω (Table 13-16). This almost constant reactance value can be translated to an element shortening of approximately 2.8% compared to the resonant element length for a 40-meter Yagi, and 3.5% for an 80-meter Yagi.

Table 13-17 shows the values of the transformed impedance and the resulting minimum SWR if the reactance of the driven element was off +5 Ω and –5 Ω versus the theoretically required value for an R_{rad} of 20 Ω. An error in reactance of 5 Ω either way is equivalent to an error length of 0.5% (see **Table 13-18**). In other words, an inaccuracy of 0.5% in element length will deteriorate the minimum SWR value from 1:1 to 1.25 or 1.3:1.

The mounting hardware for a split element will always introduce a certain amount of shunt capacitance at the driven-element feed point. This must be taken into account when designing a hairpin- or beta-match system (see example in Section 3.8.3.3).

3.10.3.2. Hairpin System for a 3-Element 80-Meter Yagi

The 3-element Yagi we developed in Section 3.5.6. has a R_{rad} of approximately 30 Ω in the phone band and approximately

Table 13-18

Capacitive Reactance Obtained by Various Percentages of Driven Element Shortening

Shorten Element	*Diameter in wavelengths 0.0010527*	*0.0004736*	*Light Taper*	*Heavy Taper*
0%	*0 Ω*	*0 Ω*	*0 Ω*	*0 Ω*
0.5	–4.8	–5.5	–4.6	–4.8
1.0	–9.6	–11.1	–9.1	–9.7
1.5	–14.3	–16.5	–13.6	–14.3
2.0	–19.1	–22.2	–18.2	–19.2
2.5	–23.8	–27.5	–22.7	–23.7
3.0	–28.6	–32.8	–27.2	–28.6
3.5	–33.5	–38.2	–31.7	–33.5
4.0	–38.5	–43.5	–36.2	–38.3

25 Ω in the CW band. Refer now to Table 13-16. For R_{rad} = 30 Ω we need –*j* 24.5 Ω feed point reactance, and 61.2 Ω of hairpin inductance across the feed terminals to obtain a 1:1 SWR in a 50-Ω system.

Adjust the antenna driven element length till you get –*j* 24.5 Ω reactance. Make sure the reflector and director are switched for operation at 3.8 MHz. Put a coil with a reactance of 61.2 Ω (L = approximately 2.56 µH) across the driven element terminals. Adjust the coil for minimum SWR (should be 1.2:1 or better).

Next, switch the antenna to the low end, remove the Beta coil, and measure the driven element impedance. It should be approximately 25 Ω with a high amount of negative reactance, of which the exact value depends on the Q-factor of the element (or the element diameter). Let us assume it is –125 Ω. From Table 13-16 we learn that we need approximately –25 Ω of negative reactance, so we will have to put a coil having a reactance of approximately +100 Ω (4.53 µH at 3.51 MHz) in series with the element to get our – *j* 24.5 Ω reactance for the element impedance . We will split this coil in two identical parts (both 2.26 µH), as shown in **Figs 13-75** and **13-76**. The required hairpin coil inductance is +50 Ω (see Table 13-16), which is (also, by coincidence) 2.26 µH on 3.51 MHz.

Fig 13-75 — K7ZV's driven element feed system. Relays add in loading inductors to switch from phone to CW band segments. The center coil is the hairpin matching inductor.

This means that the same hairpin coil (take 2.4 µH) can be used on both ends of the band, and that on the CW-end of the band we need to insert two loading coils, as shown in Fig 13-76.

The Yagi can also be fed via a direct feed, if the driven element is split and insulated from the boom. Section 3.10.4 describes two very attractive alternatives in detail.

3.10.3.3. Designing a Hairpin Match with the Yagi Design Software

You can use the Matching Systems module in the *Yagi Design* software to design a hairpin. From the prompt line you can change any of the input data, which will be immediately reflected in the dimensions of the matching system. The value

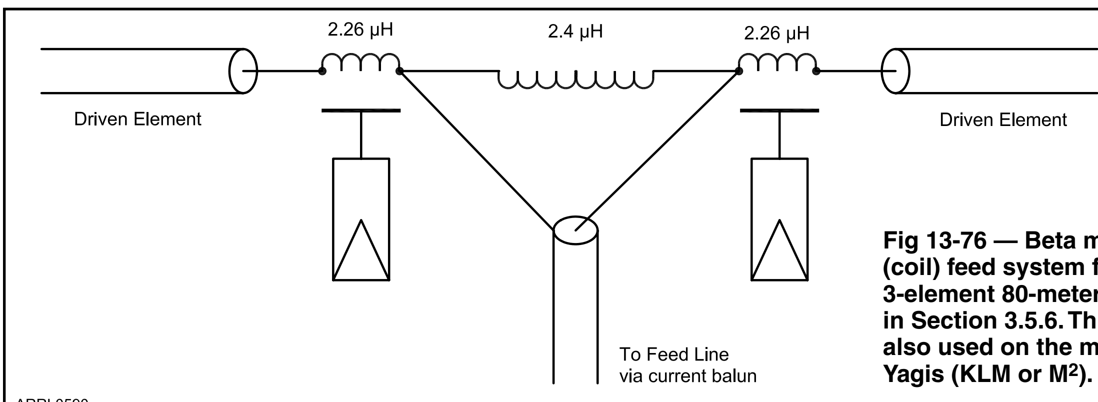


Fig 13-76 — Beta match or hairpin (coil) feed system for a the 3-element 80-meter Yagi developed in Section 3.5.6. This system is also used on the modified loaded Yagis (KLM or M²).

of the "parasitic" parallel capacitance can be specified, and it is accounted for during the calculation of the matching system. The program calculated both the required inductance as well as the dimensions of the hairpin, if you want to use a hairpin instead of a coil.

3.10.3.4. Using a Parallel Capacitor to Fine-Tune a Hairpin Matching System

A parallel capacitance of reasonable value across the split element only slightly lowers the resistive part of the impedance, while it introduces an appreciable amount of negative reactance.

Example: At a frequency of 3.79 MHz a capacitor of 150 pF (Z = – *j* 210 Ω) in parallel with an impedance of 30 + *j* 0 Ω lowers the impedance to 29.4 – *j* 4.2 Ω. (500 pF would yield 26.6 – *j* 9.5 Ω.)

This means that instead of fine-tuning the matching system by accurately shortening the driven element to obtain the required negative reactance (Table 13-16), you can use a variable capacitor across the driven element to electrically shorten the element. This is a very elegant way of tuning the hairpin matching system "on the nose." The only drawback is that it requires another vulnerable component.

3.10.4. Other Direct Feed Systems

Most 3-element Yagis do not present a 50-Ω feed-point impedance, unless you design an antenna with a low Q and trade matching ease for some forward gain. Three-element Yagis will typically show feed-point impedances varying between 18 Ω and 30 Ω. This means we really do have to use some kind of system to match the Yagi impedance to the feed-line impedance. In addition, if we want to use a feed system for an 80-meter Yagi, which has to cover both the CW and the SSB end of the band, we also will have to deal with the reactances involved.

3.10.4.1. Split Element Direct Feed System with Series Compensation

You can often use a quarter-wave transformer to achieve a reasonable match between the Yagi impedance and a 50-Ω feed line. There are two solutions. For a feed-point impedance lower than 25 Ω you can use a λ/4 length of line with an impedance of 30 Ω. This can be made by paralleling a 50-Ω, λ/4 cable with a 75-Ω, λ/4 cable. The cable can be rolled up into a coil measuring about 30 cm in diameter, which will serve as common-mode choke balun.

For impedances between 25 and 30 Ω the required impedance for a quarter-wave transformer is 37.5 Ω, made by paralleling two 75 Ω, λ/4 cables. An example of such a matching system is given in **Fig 13-77**.

Let us analyze the case of a direct feed for the 80-meter Yagi described in Section 3.3. The real part of the impedance of the Yagi is around 28 Ω, which is easy to match to a 50-Ω feed line through a 37.5-Ω, λ/4 transformer made with two parallel 75-Ω cables.

There are different approaches to handling the inductive part of the impedance at the opposite end of the band. You could dimension the driven element to be resonant on 3.8 MHz and tune out the capacitive reactance (approximately 75 Ω) by using a coil in series with the coaxial feed line. The other alternative is to dimension the driven element for resonance in the CW band and then tune out the inductive reactance on 3.75 MHz using a series capacitor. In an example I dimensioned the driven element for resonance on 3.55 MHz, and calculated the value of the series capacitor to achieve resonance on 3.75 MHz. The capacitor value is 900 pF. This matching system is extremely simple, and will guarantee maximum bandwidth as well. Again, the λ/4, 37.5-Ω transformer can be coiled up and serve as a choke balun.

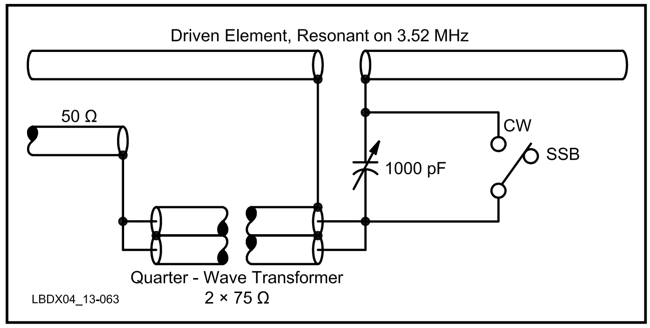


Fig 13-77 — Split-element matching system for the 80-meter Yagi. The driven element is tuned to resonance in the CW end of the band. On phone (3.8 MHz) the inductive reactance is tuned out by a simple series capacitor of 560 pF. A relay can short out the capacitor on CW. A λ/4, 37.5-Ω transformer made of two parallel 75 Ω coaxes may be coiled up to serve as a choke balun, representing a 50 Ω impedance at its end.

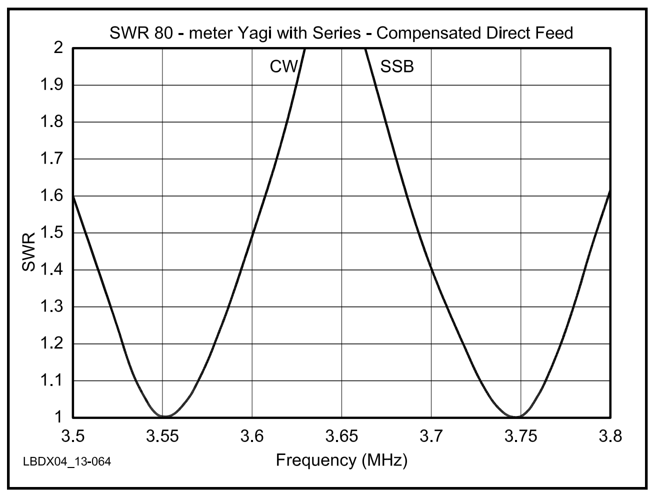


Fig 13-78 — SWR curves for the split-element feed method with series compensation and a 37.5 Ω, λ/4 transformer.

Fig 13-78 shows the SWR curves for this feed arrangement. Note that with such an arrangement it is impossible to have a good SWR in the middle of the band.

3.10.4.2. Split Element Direct Feed System with Parallel Compensation

The direct-feed system with parallel compensation does not require a λ/4 transformer as it provides a good match directly to a 50-Ω impedance. In the case of the driven element of an 80-meter Yagi, you could dimension the driven element for resonance in the middle of the band at 3.65 MHz. In that case the reactance of a typical 3-element Yagi such as the antenna developed in Sections 3.3 and 3.5, exhibits about – *j* 35 Ω at 3.5 MHz and + *j* 35 Ω at 3.8 MHz. The reac-

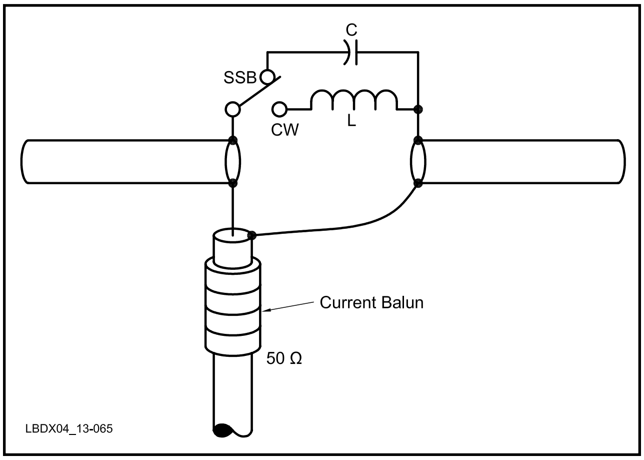


Fig 13-79 — Direct-feed system for the driven element of an 80-meter array with parallel compensation, which makes it possible to obtain a good SWR in both the CW and SSB sections of the band. See text for details.

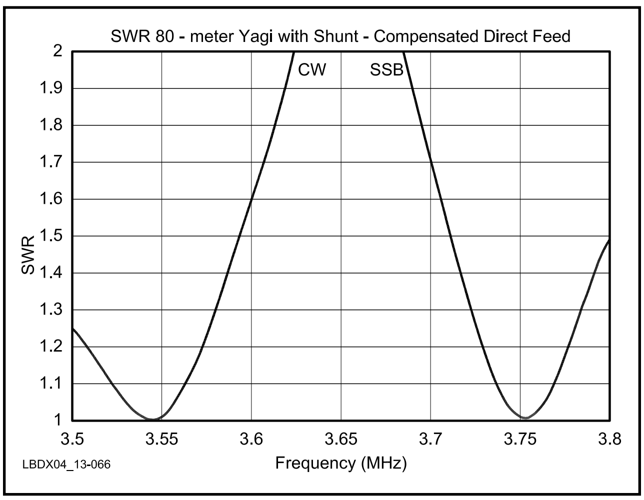


Fig 13-80 — SWR curves for the 80-meter 3-element Yagi using parallel compensation. The driven element, initially tuned to resonance on 3.65 MHz, was tuned to resonance on 3.55 MHz using a parallel inductor of 2.2 µH. Likewise, the same element was tuned to resonance on 3.75 MHz using a parallel capacitor of 700 pF. The resulting SWR curves show an outstanding bandwidth.

tance can be tuned out by a parallel coil or capacitor (see **Fig 13-79**). A coil with an inductance of 2.2 µH will tune the driven element to resonance on 3.55 MHz and yield a resistance of 50 Ω (a lucky coincidence!). Likewise, a capacitor with a value of 700 pF will tune the element to resonance on 3.8 MHz, also with a resistance of 50 Ω.

The value of these components can easily be calculated using the Shunt Impedance Network module of the *New Low Band Software*. With a simple relay you can switch either the coil or the capacitor in parallel with the feed point, and obtain a fine matching system for either the CW or the SSB end of the band. **Fig 13-80** shows the SWR curves obtained with such an arrangement.

3.11. The OH8X Mammoth Antenna System

Up until the 4th edition of this book, I wrote that a rotary Yagi for 160 meters had never so far been built. Well, I guess, some of those Finns must have read my book.

In this book I have always tried to describe antenna systems that could be built by a whole range of motivated Low Banders. This "monsterissimo" 160- and 80-meter Yagi antenna was designed and built by a small group of hams in Finland who really belong right at the top of the aforementioned range.

No, I am not going to give you the step by step instruction on how to build your own Radio Arcala system. I honestly do not think very many hams really want to build such an antenna system. I think this fantastic realization is just there to marvel all of us. But it is real, it is not a dream.

Those Finns are something else. Back in the 70s I made a trip to Oulu near the Arctic Circle in Finland to marvel at OH8OS's 6 × 6 element stack Yagi system. I thought this was the biggest Amateur Radio antenna system I would ever see.

I was wrong. But you have to be a Finn to do better, I guess. I must say that the area around Oulu appears to be fertile ground for impressive accomplishments. Arcala, where the OH8X antenna system was built, is only approximately 40 km from Oulu, where I met Simon, OH8OS more than 30 years ago to see his 6 × 6 element Yagi stack. Juha, OH8NC, was there as well, as a very young ham.

This time the Finns built a 100-meter tall huge, heavy tower carrying a couple of full-size Yagis for 160 (yes, you read it correctly) and for 80 meters. The system consists of a 3-element full size 160-meter Yagi, designed by OH1TV, using 59-meter long elements on a 71-meter long boom at a height of 80 meters. The 160-meter antenna is topped by a 5-element full-sized 80-meter Yagi, designed by OH5BR, using 46 meter long elements on a 60-meter long boom Yagi at 90 meters. The tower height is 100 meters, and the whole system weighs nearly 40,000 kg. And that is without the weight of the two 40-meter 4-element Yagis that were added later. See **Figs 13-81** and **13-82**. Radiation patterns are shown in **Fig 13-83**.

The 160-meter beam can switch directions instantaneously (you don't rotate this monster in just a few seconds!). The Yagi has a gain of 12.9 dBi (at 26° take off angle) and a F/B varying between 20 and 30 dB over its entire operating range going from 1810 to 1890 kHz (in two ranges). The SWR is maximum 1.5:1 over a span of 50 kHz.

The 80-meter beam has 15.7 dBi gain, with 20 dB F/B (at 12° take off angle), and that is on two operating ranges: 3500-3560 and 3700-3800 kHz. The SWR is less than 1.3:1, in both the CW section and the SSB section.

The system was designed to withstand a maximum wind speed of not less than 250 km/h. The size of the 160-meter boom is a startling 2.2 meters wide, large enough to drive a small car inside the triangular shaped boom. While we're at it, let me continue marveling you: the rotator weighs 2000 kg and uses a little electrical motor having a power of 11 kW. **Fig 13-84** gives you an idea of the size of the tower.

Some of the tower anchoring points are not less than 120 meters from the base of the tower. The top rotating ring installed just below the 160-meter Yagi weighs 3300 kg, and

Fig 13-81 — Here it is, one of the seven amateur world wonders.

has a diameter of 3.8 meters. The top guys wires are Phillystran and a total of 3.5 km of guy wires are used. Final detail to impress you: about 600 liters of paint were used on the tower. And no, I do not have an idea of the budget involved.

3.11.1. Some Design Topics

All elements of the 160-meter Yagi are identical in size and insulated from the boom but dc grounded for lightning protection. This makes it possible to fine tune the elements for different sections of the band.

In the 80-meter Yagi the element lengths were adjusted for the SSB band. The reflector also uses a small center loading coil for fine tuning. For operation in the CW band, coil loading is done on all elements.

Needless to say, the engineering challenges in such a project were immense, both from a mechanical as well as an electrical point of view. The lattice structure of elements has so many details that it was impossible to enter all of it in the model, as that would imply too many segments. Therefore simplified models were used, which were calibrated with the results of real life measurements.

It appears that heavily tapered sections behave very differently from what we are used to seeing with wire or tubular sections with moderate tapering. Even the size and the diameter of the cross tubes from the lattice structure appears to have a profound influence on the impedance of the element.

The initial 160-meter elements did not use the T-shaped top loading, but rather aluminum tubing tapering further down to a very small diameter. Amazingly these elements did not show resonance at all. Replacing the in-line tips with the T-shaped top load and adding some center inductive loading cured that problem. On the 80-meter beam the designers were confronted with a similar problem on the reflector element. Adding some extra inductive loading solved that problem.

Another problem the designers ran into was too high a parasitic capacitance between the insulated elements and the boom (see Sections 3.8.3.3. and 3.8.3.5). In Edition 3 of this book I described the design of a 3-element full-size Yagi (Section 3.5.3, page 13-27) where the problem of the element-to-boom capacitance was also discussed. If you try to tune an element to resonance (no reactive part in the impedance), and if that element has high capacitance to the boom (which is inevitable), resonance will be obtained at a higher frequency than if there was no capacitance. This is the frequency at which some inductive reactance together with the parasitic capacitance will form an L-network, which in turn will raise the measured element impedance. Eventually the coupling between element and boom was redesigned, ensuring a much lower capacitance, which helped solve the problem.

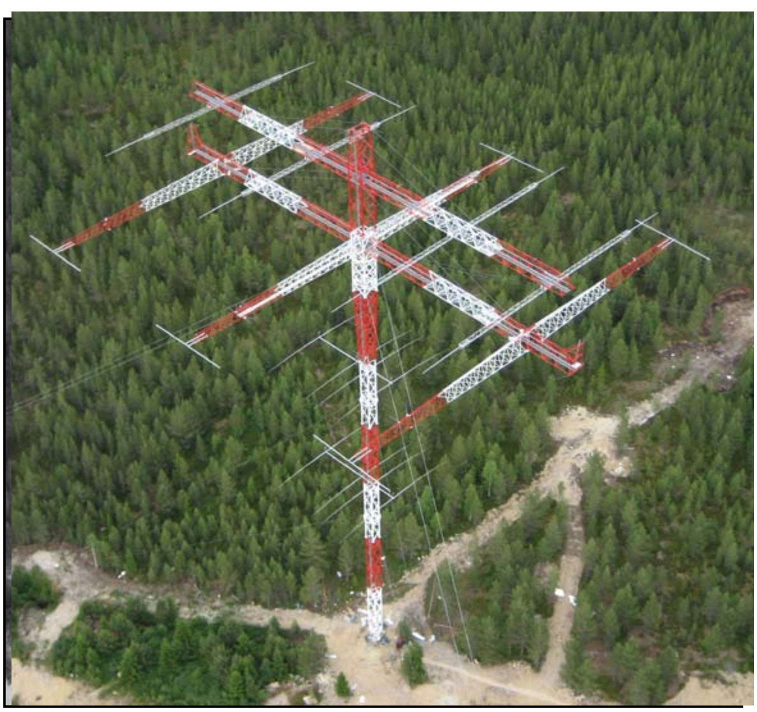


Fig 13-82 — Seen from above it's just as impressive, if not more. Note the stacked 40 meter 4-element Yagis, which can serve as dimensional references.

Both antennas cannot be used simultaneously, as, to avoid interaction, the antenna that is not used must be heavily detuned. When operating the 80-meter antenna, it appeared to be impossible to reduce parasitic currents in the 160-meter elements to a very low value, but the phase of these currents could be controlled, which was done to make these parasitic currents cooperate with the 80-meter an-

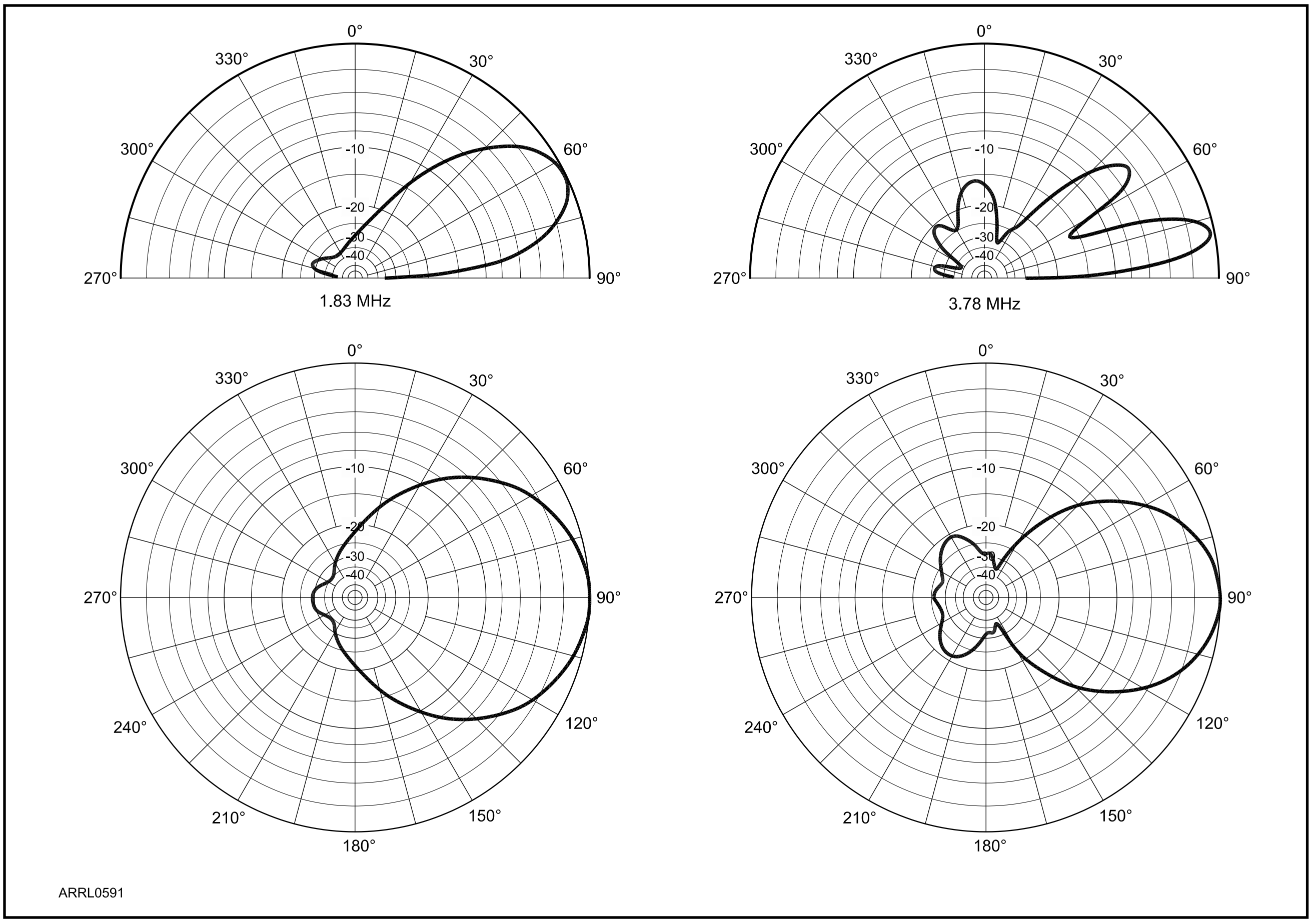


Fig 13-83 — Radiation patterns of the 160 and the 80 meter Yagis at OH8X.

Fig 13-84 — Juha, OH8NC, who owns the biggest Amateur Radio antennas in the world, at the base of the bottom section of the tower.

tenna, rather than to upset the pattern. As to be expected, the higher frequency antenna (80 meters) has less influence on the lower frequency antenna (160 meters).

3.11.2. The Future

Will this antenna be good enough to win all low band contests? A challenge for the poor soldiers, fighting with lighter weapons, is to prove that although antennas are very important, location (read propagation, and aurora) may be at least as important. That is my humble opinion. Only the future will tell us what the outcome will be.

For more details, visit **www.radioarcala.com**.

4. QUADS

4.1. Modeling Quad Antennas

4.1.1. *MININEC*-Based Programs

Modeling quad antennas with *MININEC*-based programs requires very special attention. To obtain proper results the number of wire segments should be carefully chosen (see **Table 13-19**). Near the corners of the loop, the segments must

Table 13-19

Influence of the Number of Sections on the Impedance of a Quad Loop

Taper Arrangement	*Calculated Impedance, Ω*
Nontapered, 4 × 5 segments	123 – *j* 20
Nontapered, 4 × 10 segments	130 + *j* 44
Nontapered, 4 × 20 segments	133 + *j* 78
Nontapered, 4 × 40 segments	135 + *j* 95
Nontapered, 4 × 50 segments	135 + *j* 97
Nontapered, 4 × 60 segments	135 + *j* 98
32 sections, tapering from 1.0 to 5.0 m	131 + *j* 85
56 sections, tapering from 0.4 to 2.0 m	134 + *j* 102
64 sections, tapering from 0.2 to 2.0 m*	135 + *j* 104
104 sections, tapering from 0.2 to 1.0 m	135 + *j* 104

Note: See Fig 13-85 regarding the taper procedure.
*Taper arrangement illustrated in Fig 13-85.

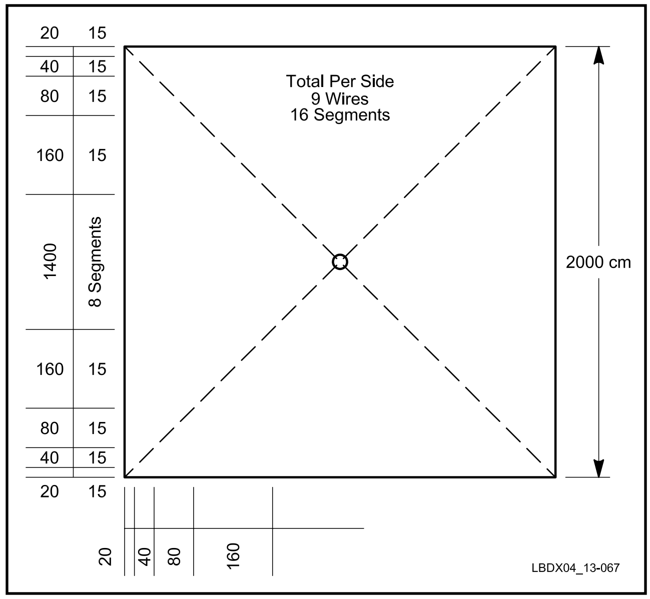


Fig 13-85 — Tapering of segment lengths for *MININEC* analysis. See Table 13-19 for the results with different tapering arrangements. With the segment-length-taper procedure shown here, the result with a total of just 56 tapered segments is as good as for 240 segments of identical length.

be short enough not to introduce a significant error in the results. Segments as short as 20 cm must be used on an 80-meter quad to obtain reliable impedance results on multi-element loop antennas. Years ago, when most modeling programs still used a *MININEC* core, W7EL developed a taper technique in his *ELNEC* software, where segments automatically got progressively shorter when coming to a corner. See **Fig 13-85**.

4.1.2. *NEC*-Based Programs

NEC-based programs do not exhibit the above problem, and no special precautions have to be taken to obtain correct results.

4.2. Two-Element Full-Size 80-Meter Quad with a Parasitic Reflector

Fig 13-86 shows the configuration of a 2-element 75-meter quad on a 12-meter boom, and **Fig 13-87** shows the radiation patterns. The optimum antenna height is 35 meters for the center of the quad. Whether you use the square or the diamond shape does not make any difference. The dimensions remain the same, as well as the results. I will describe a diamond-shaped quad, which has the advantage of making it possible to route the feed line and the loading wires along the fiberglass arms.

I designed this quad with two quad loops of identical length. The total circumference for the quad loop is 1.0033 λ (for a 2-mm-OD conductor or #12 wire). The parasitic element is loaded with a coil or a stub having an inductive reactance of + *j* 150 Ω. The gain is 3.7 dB over a single loop at the same height over the same ground.

In the model I used 3.775 MHz as a central design frequency. This is because the SWR curve rises more sharply on the low side of the design frequency than it does on the high side. You can optimize the quad by changing the reactance of the loading stub as you change the operating frequency.

Figs 13-88 and **13-89** show the gain, F/B and SWR for the 2-element quad with a fixed loading stub or coil (+150 Ω) as well as for a design where the loading stub reactance is varied. To make the antenna instantly reversible in direction, you can run two λ/4, 75-Ω lines, one to each element. Using the Coax Transformer/Smith Chart module from the *New Low Band Software* we see that a + *j* 160 Ω impedance at the end of a λ/4 long 75-Ω transmission line (at 3.775 MHz) looks like a – *j* 35 Ω impedance. This means that a λ/4, 75-Ω (RG-11) line terminated in a capacitor having a reactance of –35 Ω is all that we need to tune the parasitic element into a reflector. A switch box mounted at the center of the boom houses the necessary relay switching harness and the required variable

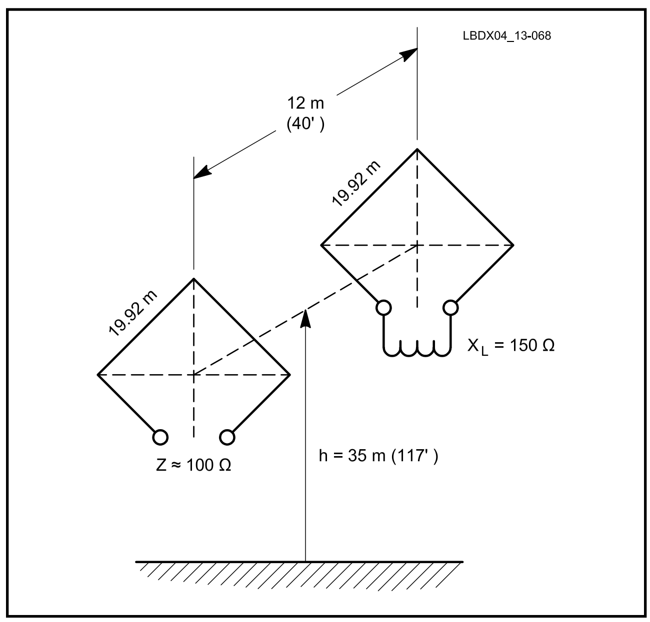


Fig 13-86 — Configuration of a 2-element cubical quad antenna designed for 75-meter SSB. Radiation patterns are shown in Fig 13-87. By using a remote tuning system for adjusting the loading of the reflector, the quad can be made to exhibit an F/B of better than 22 dB over the entire operating range. See text for details.

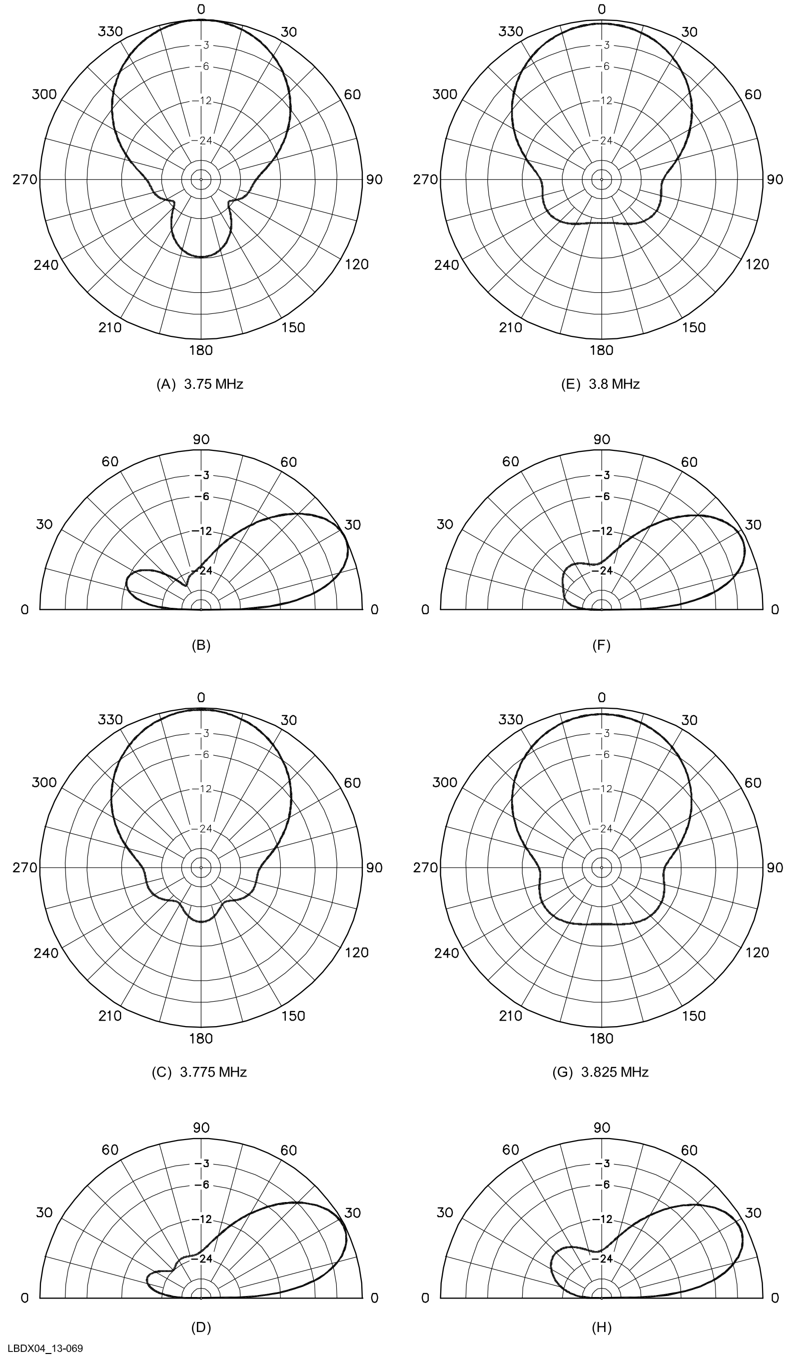


Fig 13-87 — Radiation patterns of the 75-meter SSB 2-element quad at various frequencies. The antenna was optimized in the 3.775 to 3.8-MHz range. In that range an F/B of better than 20 dB is obtained. All patterns are plotted to the same scale, and azimuth patterns are taken at a takeoff angle of 28°.

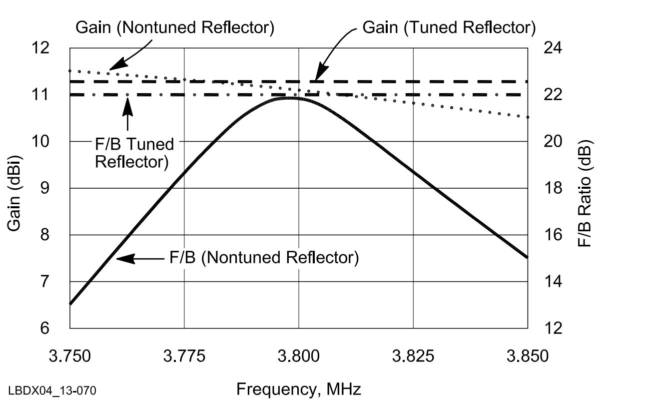


Fig 13-88 — Gain and F/B for the 2-element 75-meter quad with fixed reflector tuning, and with adjustable reflector tuning. The antenna is modeled at a height of 35 meters above good ground. With fixed tuning the F/B is 20 dB over 30 kHz, and the gain drops almost 0.5 dB from the low end to the high end of the operating passband (100 kHz). When the reflector tuning is made variable, the gain as well as the F/B remain constant over the operating band.

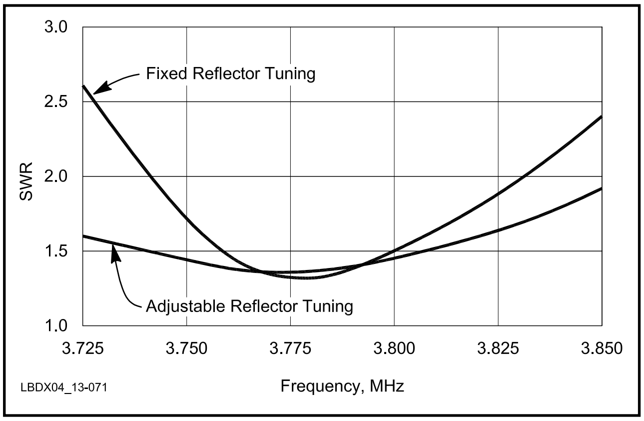


Fig 13-89 — SWR curves for the 2-element 75-meter quad. The SWR is plotted versus a nominal input impedance of 100 Ω, which is then matched to a 50-Ω impedance using a λ/4, 75-Ω line. Note that the variable reflector-tuning extends the operating bandwidth considerably for the lower frequencies.

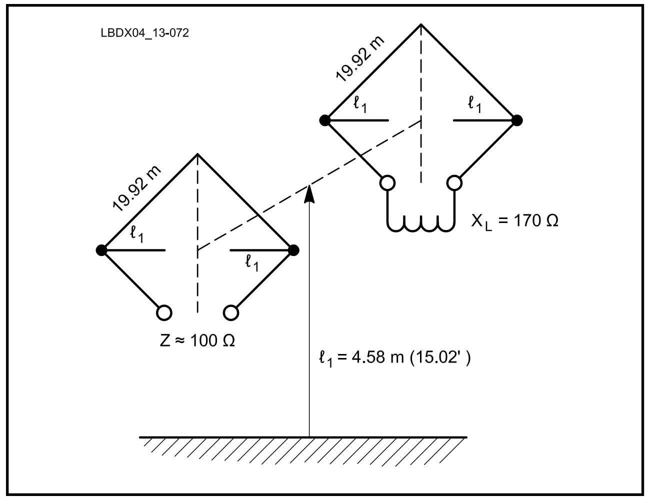


Fig 13-90 — The configuration the 2-element quad tuned for the CW end of the band.

capacitor to do the job.

The required optimal loading impedances can be obtained as follows:

3.750 MHz: X_L = 180 Ω, C = 1322 pF
3.775 MHz: X_L = 160 Ω, C = 1205 pF
3.800 MHz: X_L = 150 Ω, C = 1148 pF
3.825 MHz: X_L = 120 Ω, C = 931 pF
3.850 MHz: X_L = 100 Ω, C = 785 pF

If you tune the reflector for optimum value you will obtain better than 22-dB F/B ratio at all frequencies from 3.75 to 3.85 MHz, and the SWR curve will be much flatter than without the tuned reflector (see Figs 13-88 and 13-89).

The quad can also be made switchable from the SSB to the CW end of 80 meters. There are two methods of loading the elements, inductive loading and capacitive loading (see also the chapter on Large Loops). The capacitive method, which I will describe here, is the most simple.

4.2.1. Capacitive Loading

A small single-pole high-voltage relay at the tip of the horizontal fiberglass arms can switch the loading wires in and out of the circuit. The calculated length for the loading wires to switch the quad from the SSB end of the band (3.775 MHz) to the CW end (3.525 MHz) is 4.58 meters. Note that you are switching at a high-voltage point, which means that a high-voltage relay is essential.

As you will have to run a dc feed line to the relay on the tip of the spreader, it is likely that the loading wire will capacitively couple to the feed wire. Use small chokes or ferrite beads on the feed wire to decouple it from the loading wires.

Fig 13-90 shows the configuration and **Fig 13-91** shows the radiation patterns for the 2-element quad tuned for the CW end of the band. The patterns are for a fixed reflector-loading reactance of 170 Ω.

As described above, you can optimize the performance by tuning the loading system as we change frequency. Using the same λ/4, 75-Ω line (cut for 3.775 MHz), you can obtain a constant 22-dB F/B (measured at the peak elevation angle of 29°) on all frequencies from 3.5 to 3.6 MHz with the following capacitor values at the end of the 75-Ω line:

3.500 MHz: X_L = 180 Ω, C = 1083 pF
3.525 MHz: X_L = 160 Ω, C = 999 pF
3.550 MHz: X_L = 140 Ω, C = 897 pF
3.575 MHz: X_L = 120 Ω, C = 795 pF
3.600 MHz: X_L = 100 Ω, C = 660 pF

The optimized quad has a gain at the low end of 80 meters that is 0.3 dB less than at the high end of the band. The gain is 10.8 dBi at 35 meters over good ground.

Fig 13-92 shows the SWR curve of the quad at the CW end of the band, with both a fixed reflector loading (X_L = 170 Ω) and a variable setup as explained above. The switching harness for the 2-element quad is shown in **Fig 13-93**. The tuning capacitor at the end of the 75-Ω line going to the reflector can be made of a 500-pF fixed capacitor in parallel with a 100 to 1000-pF variable capacitor. Note that you need a choke balun at both 75-Ω feed lines reaching the loops. This can be in the form of a stack of ferrite beads or as coiled-up coax.

It is also possible to design a 2-element quad array with both elements fed. With the dimensions used in the above design, a phase delay of 135° with identical feed-current magnitudes yields a gain that is very similar to what is obtained with the parasitic reflector. The F/B may be a little better than with the parasitic array. As the array is not fed in quadrature, the feed arrangement is certainly not simpler than for the parasitic array, however. The parasitic array is simpler to tune, since the reflector stub (the capacitor value) can be simply adjusted for best F/B.

4.3. Two-Element Reduced-Size Quad

D. Courtier-Dutton, G3FPQ, built a reasonably sized 2-element 40-meter rotatable quad that performs extremely well. The quad side dimensions are 15 meters, and the elements are loaded as shown in **Fig 13-94**. The single loop showed a radiation resistance of 50 Ω. Adding a reflector 12 meters away from the driven element (0.14 λ spacing), dropped the radiation resistance to approximately 30 Ω. The loading wires are spaced 110 cm from the vertical loop wires, and are almost as long as the vertical loop wires. The loading wires are trimmed to

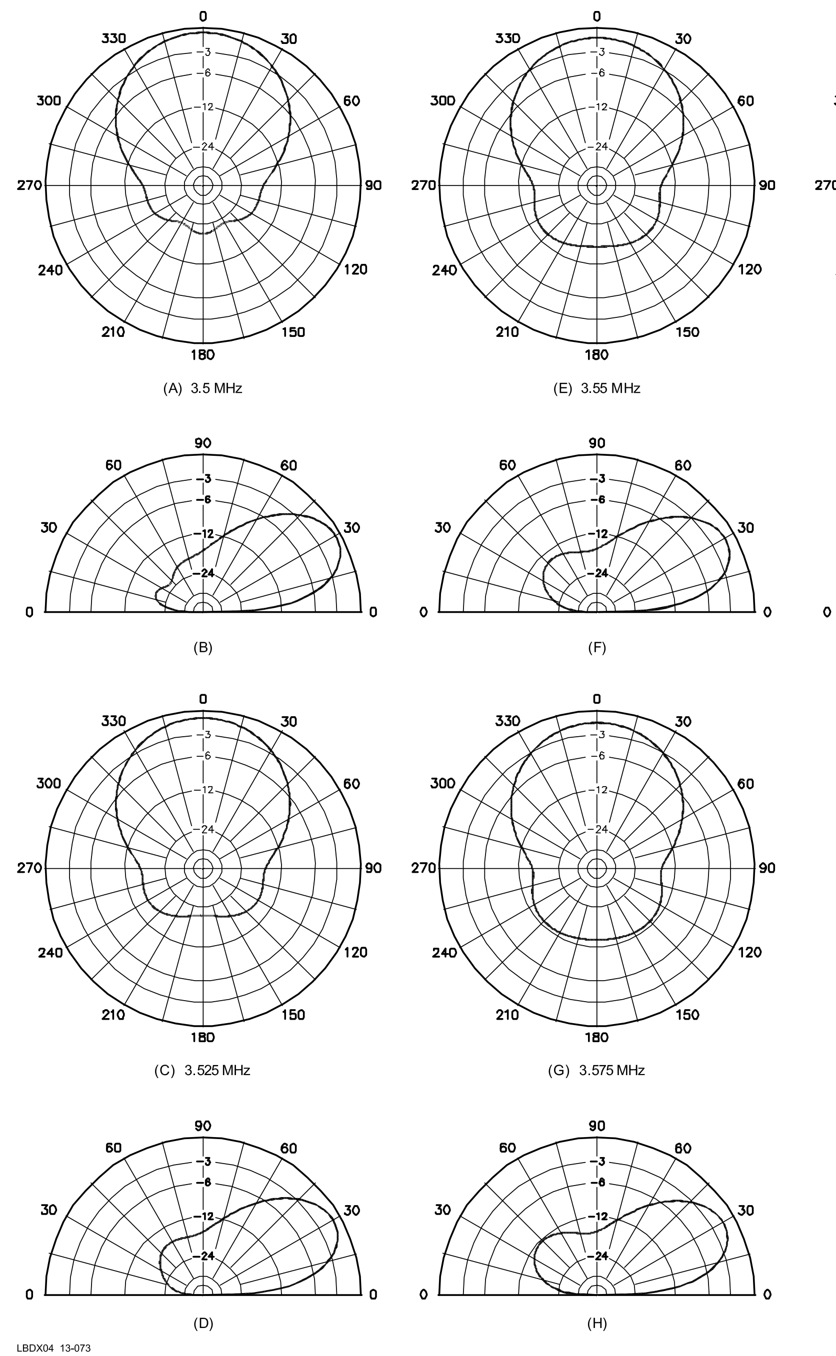


Fig 13-91 — Radiation patterns of the 2-element 80-meter quad when capacitively loaded to operate in the CW portion of the band. Azimuth patterns are taken at an elevation angle of 29°. All patterns are plotted to the same scale as the SSB patterns in Fig 13-87. The gain is a fraction of a dB less than at the high end of the band, but the directional properties are identical. The loading was optimized to yield the best F/B between 3.5 and 3.525 MHz.

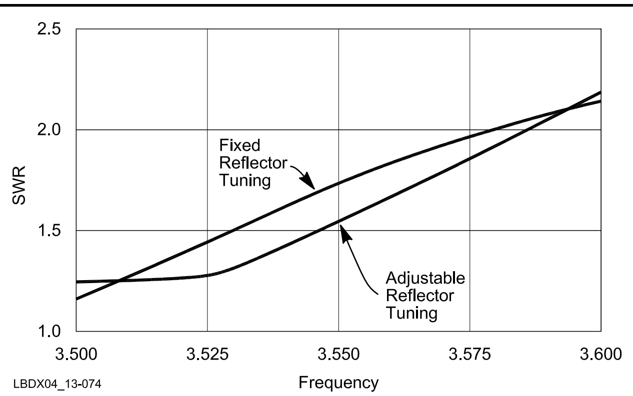


Fig 13-92 — SWR curves for the 2-element 80-meter quad referred to the nominal 100-Ω feed-point impedance. The tuned reflector does not significantly improve the SWR on the high-frequency end. The design was adjusted for the best SWR in the 3.5 to 3.525-MHz region.

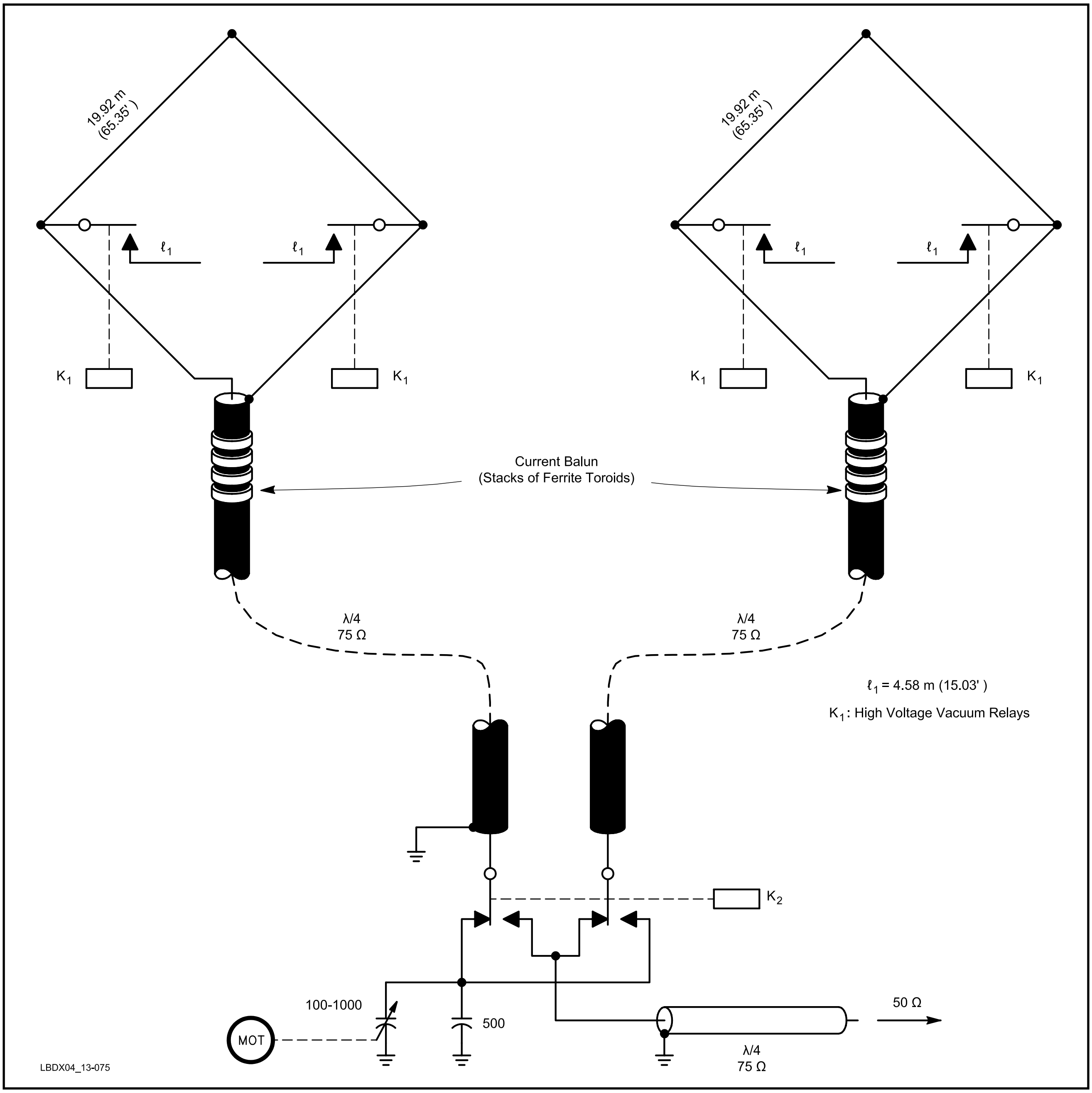


Fig 13-93 — Feeding and switching method for the 2-element 80-meter quad. The four high-voltage vacuum relays connect the loading wires to the high-voltage points of the quad, to load the elements to resonance in the CW band. Relay K2 switches directions. The motor-driven variable capacitor (50-1000 pF) is used to tune the reflector for maximum F/B at any part in the CW or phone band.

adjust the resonant frequency of the element. G3FPQ reports a 90-kHz bandwidth from the 2-element quad with the apex at 40 meters. The middle 7 meters of the spreaders are made of aluminum tubing, and 3.6-meter long tips are made of fiberglass. A front-to-back ratio of up to 30 dB has been reported.

G3FPQ indicates that the length of the reflector element is exactly the same as the length of the driven element for the best F/B ratio. This may seem odd, and is certainly not the case for a full-size quad. The same effect has been found with some 2-element Yagi arrays.

4.4. Three-Element 80-Meter Quad

Fig 13-95 shows the 3-element full-size 80-meter quad at DJ4PT. The boom is 26 meters long, and the boom height is 30 meters. Interlaced on the same boom are five elements for a 40-meter quad.

The greatest challenge in building a quad antenna of such proportions is mechanical in nature. The mechanical design was done by H. Lumpe, DJ6JC, now a Silent Key, who was a well-known professional tower manufacturer in Germany. The center parts of the quad spreaders were made of aluminum lattice sections that are insulated from the boom and broken up at given intervals as well. The tubular sections are made of fiberglass. The driven element is mounted less than 1 meter from the center. This makes it possible to reach the feed point from the tower. To be able to reach the lower tips of the two

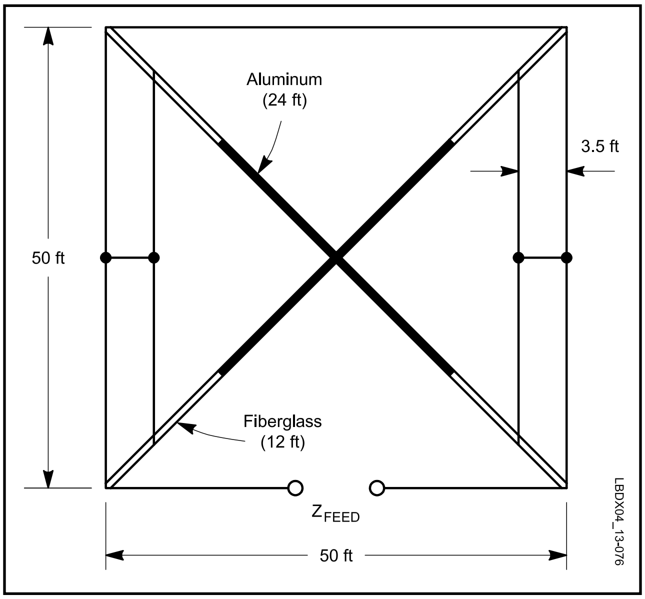


Fig 13-94 — Reduced-size 2-element 80-meter quad by D. Courtier-Dutton, G3FPQ. The elements are capacitively loaded, as explained in detail in the chapter on large loop antennas.

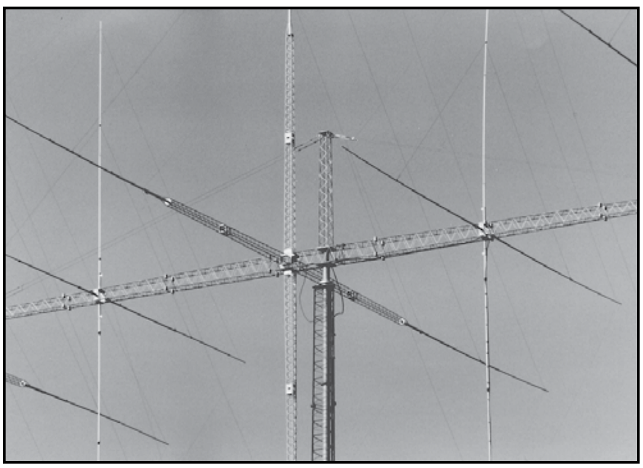


Fig 13-95 — This impressive 3-element full-size 80-meter quad, with an interlaced 5-element 40-meter quad, on a 26-meter (87-foot) long boom, sits on top of a self-supporting 30-meter (100-foot) tower at DJ4PT. The antenna was built by DJ6JC (SK).

parasitic elements for tuning, a 26-meter tower was installed exactly 13 meters from the main tower. On top of this smaller tower a special platform was installed from which one can easily tune the parasitic elements.

The weight of the quad is approximately 2000 kg (4400 lb). The monster quad is mounted on top of a 30-meter self-supporting steel tower, also built by DJ6JC. The rotator was placed at the bottom of the tower, and a 20-cm OD rotating pipe with a 10-mm thick wall takes care of the rotating job.

4.5. The W6YA 40-Meter Quad

Jim McCook, W6YA, lives in a fairly typical suburban QTH, and has his neighbors and family accustomed to one crank-up tower (Tri-EX LM-470), on which he must put all of his antennas. Jim has 4-element monoband Yagis for 10, 15 and 20 meters and a 30/17/12 meter triband dipole. Jim set out to make it work on 9 bands. See **Fig 13-96**.

For 40 meters, Jim has extended the 12-meter boom of

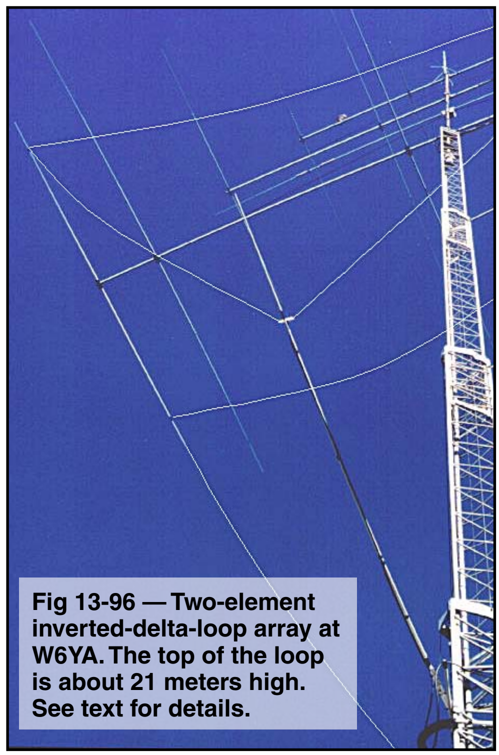


Fig 13-96 — Two-element inverted-delta-loop array at W6YA. The top of the loop is about 21 meters high. See text for details.

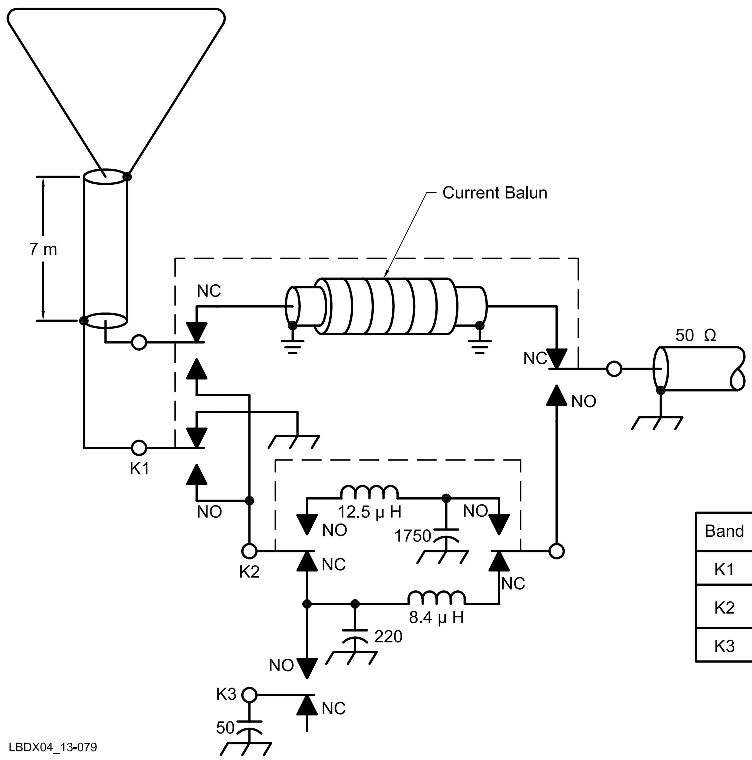


Fig 13-97 — Arrangements for feeding the driven loop of the 40-meter inverted-delta quad, as a vertical on both 80 and 160 meters. Two LC networks are used. The values indicated are approximate.

Band	40	75	80	160
K1	O	X	X	X
K2	n.a.	O	O	X
K3	n.a.	O	X	n.a.

O = Open
X = Closed
n.a. = Doesn't Matter

his 20-meter Yagi to 14.5 meters. At the ends of the boom he mounted fiberglass quad poles, which support two inverted delta loops, separated 6 meters from each other. One loop is tuned as a reflector (3% longer). The driven element is fed through a λ/4 section of RG-11 75-Ω cable. The inverted delta loops are kept taut by supporting two more abutted quad poles at the bottom. This quad-pole assembly hangs freely, supported only by the loop wires. The assembly pivots around the tower during rotation. The top horizontal sections are allowed to sag slightly (about 2 meters) to minimize interaction with the 20-meter Yagi. This arrangement has been up for 18 years, and has helped Jim to work all but three countries on 40 meters! Jim reports a 2:1 SWR bandwidth of 200 kHz and a F/B of 15-20 dB.

The driven element loop has a 14.78-meter "flat top," 14.32-meter sloping length on one side and 4.63 meters on the other side. This offset is to keep the bottom fiberglass-pole assembly free from the tower. The reflector measures 14.78 meters, 15.04 meters and 15.34 meters respectively. The above lengths are for peak performance on 7.020 MHz.

This quad arrangement has low wind load. Jim also uses this arrangement on 80 and even on 160 meters. On 80 and 160, Jim straps the feed point of the driven loop, and feeds the loops with its feed line, at ground level via appropriate matching networks. If you feel tempted to try this combination, I would advise you to use an antenna analyzer to measure the feed-point impedance on both 80 and 160, and design an appropriate network. The feed-point impedance on 80 meters is approximately $90 + j\,366\ \Omega$, and on 3.8 MHz, $120 + j\,460\ \Omega$. On 1.83 MHz the impedance, including an estimated series-equivalent ground loss resistance of 10 Ω is $25 - j\,72\ \Omega$. The appropriate matching networks for the different frequencies are shown in **Fig 13-97**.

It goes without saying that a good ground-radial system is essential for this antenna. Jim complements his 9-bands-on-one-tower antenna system with a modified 30-meter rotary dipole, which he center loads for 80 meters (see Fig 14-19 in Chapter 14). He anticipates adding a second set of loading coils to use the same short loaded dipole on 160 as well.

4.6. Quad or Yagi

I must admit I have very little first-hand experience with quad antennas. But I can think of a few disadvantages of quad antennas as compared to Yagi antennas:

- Much better F/B can be obtained with Yagis.
- Quads are three-dimensional; you can't easily assemble a quad on the ground, and then pick it up with a crane and put it on the tower. You must do a lot of assembly work with the boom in the air.
- Wires break.
- It is a non-efficient material user. All the metalwork you put up is not part of the antenna; it is just a support structure.

So far as electrical performance is concerned, a well-tuned quad antenna should marginally outperform a Yagi with the same boom length, at least as far as gain is concerned. The difference, on the order of a fraction of a dB to a maximum of 1 dB, is more of an academic than of a practical nature.

To prevent ice build up on the quad wire, you can feed a current (ac or dc) through the loops. The voltage should be adjusted so as to raise the temperature in the wire just enough to prevent ice loading.

The fact is that the great majority of rotatable arrays on the low bands are Yagis. This seems to indicate that the mechanical issues are probably harder to solve with quads than with Yagis.

5. BUYING A COMMERCIAL LOW-BAND YAGI ANTENNA

5.1. 80-Meter Yagis

5.1.1. The Linear-Loading Approach

To my knowledge, there is no manufacturer who is currently offering a full-size 80-meter rotatable Yagi antenna. The 2 and 3-element shortened Yagis with linear-loaded elements, originally developed by Mike Staal, K6MYC, for KLM more than 20 years ago, have held up over the years. Later M^2, Mike Staal's new company, and Force-12 sell 80-meter Yagis using linear-loaded Yagis based on his original design. The merits of the linear-loaded design have long been established. The only inherent design compromise seems to be the sacrifice in top-notch directivity, which is caused by some radiation from the slant loading wires (see Section 3.7), as well as the intrinsic lower Q of the loading devices as compared to well designed and well manufactured loading coils (see section 3.5.1).

5.1.2. Mechanical Issues

When deciding to buy one of these antennas, check the mechanical issues closely. This is what makes an 80-meter Yagi last or not. I must admit I am really scared when I see how some of these antennas are made. I see 80-meter Yagis using booms with a wall thickness that is less than half of the wall on my 40-meter Yagi boom! I see how a simple aluminum plate of a few mm thickness is connected by a few simple rivets to the boom, and this is supposed to hold the 25-meter long element in a lasting way. I have doubts these antennas can ever stay up in windy areas. In my QTH they would not last one winter!

Mechanical issues are the real issues for a long-lasting 80-meter antenna. So, if you decide to spend a lot of money, take a very close look at the mechanics. An antenna built to withstand high winds and lots of ice loading will inevitably use more aluminum than a flimsy antenna that won't withstand a 90-km/h breeze. And aluminum costs money. There is a price for a good mechanical design. There are no two ways about it.

I know it takes approximately 45 kg of 6061-T6 aluminum to make a full-size 40-meter element that will withstand 160 km/h winds (+30% gusts). I have my doubts that an 80-meter 3-element Yagi, with elements that are 20% longer than for a 40-meter reflector, can be built for a total weight of only 120 kg.

As an example, I modeled the elements of the old KLM 80M-3 Yagi to assess its wind-survival speed. The element mechanical data were taken from the assembly manual of the 80M-3 antenna. The safe wind survival speed turned out to be 90 km/h, without a 30% higher gust factor. An element stress-analysis shows a very unbalanced design: While sections 2 (2-inch OD) and 3 (1.75-inch OD) are loaded to the limit, the three next sections are only loaded to about 60% of their capabilities.

The tip section is only loaded 25%. This does not necessarily mean that the element will disintegrate at 90 km/h, since this assumes that the wind blows at a right angle with respect to the elements. Putting the boom into the wind (perpendicular to the wind direction) will take all the stress off the elements. Provided that the side bracing of the boom is well done, it is likely that the Yagi boom will survive wind speeds above 90 km/h. Using the guidelines explained by Leeson in his book (Ref 964), sections 2, 3 and 4 can be reinforced by doubling the wall thickness to increase the wind survival speed to 123 km/h. In any case, you should add side guying of the central 3-inch section of the elements. Short boom extensions will be required to do this.

Fig 13-98 — TheOptibeam OB3-80, a 3-element high-Q coil loaded Yagi.

5.1.3 Low Losses

A very interesting point was brought up by W6ANR. Often the weak point of a design is the lack of long-lasting, low-resistance electrical contacts. Lossy contacts ruin the gain and the pattern of any array. Invest in some good contact grease, Parker screws and heat-shrink tube for assembling the Yagi. Make sure all is done to prevent corrosion in the linear-loading wires. Better still, stay away from these wires and invest in high-Q loading coils.

5.1.4 High-Q Coil-Loaded Yagis

Over the years it has become obvious that high-Q loading coils (see Section 3.6.1.) have replaced the linear loading de-icers from yesteryear.

Optibeam (sold in the US by Array Solutions) makes very nice 2 and 3-element high-Q coil loaded Yagis that are switchable from the CW band to the SSB band. The elements are "somewhat short" (only 23 meters, barely longer than a 40 meter full-size element) but judging by the reports written by the customers, the antenna works very well. Of course we all know that the company has taken care of its excellent reputation in the market. **Fig 13-98** shows one of their antennas.

5.1.5. Gain Figures

Be very careful when comparing published gain figures. The only thing that really makes sense are free-space dBi gain figures, but the sales and marketing guys like to inflate these low figures and add ground reflection gain, which could be anything up to 6 dB, depending on ground quality and antenna height. Don't let these guys fool you!

I have withheld from publishing a list of commercial low-band antennas as I fear that I might not list all of them. And I worry that doing so might indicate some kind of endorsement on my part. If you plan to buy a commercial low-band antenna, I suggest you get a reference list from the manufacturer, and contact some of the customers. Or better yet, ask around on the Internet.

Contents

CHAPTER 14

Low Band DXing From a Small Garden

The story to follow is undoubtedly the story of many, and it could be the story of even more people, provided they tried. If you don't have a large backyard or a farm, read it. It's the story of how a very good friend of mine, George Oliva, K2UO, started on 160 meters.

"Having been an avid DXer for many years and having achieved 'Number One' Honor Roll status on CW, SSB and Mixed, 5BDXCC, etc, I was in search of a new challenge. Some of the locals had started on 160 meters but I assumed that I didn't have the space for the antennas needed to work Top Band on a half-acre lot. My amplifier didn't cover 160 and my tower was a crank-up type so a shunt feed wouldn't work very well. In 1985, I grew bored of the WARC bands and took on the 160 challenge! I put up what has since become known as my 'stealth' dipole, a full quarter wave on 160, not in a straight line and not very high the air. I worked 75 countries over the next 36 months with 100 watts and no special receiving antennas. Although most were relatively non-exciting, I did manage to snag 3B8CF, D44BC and even VK7BC.

"I next picked up a linear which did cover 160 meters. Now I began to see the need for special receiving antennas because I could now work everything I could hear but knew from the locals and packet clusters that I was not hearing a lot. I asked my 'friendly' neighbor if I could run a wire up the back end of his property line and I was now in business with 550+ foot single wire terminated Beverage antennas pointed to about 65 and 245 degrees. This antenna is truly amazing. I could now hear stations that I couldn't even imagine hearing on the 'stealth' dipole.

"Although I am not the first to get through, I usually make it in the pileups. I have worked Bouvet, Peter I, Heard Island, South Sandwich, and now have 247 countries worked on 160 meters, almost all on CW of course.

"When other hams visit my station and look at my Top Band antennas, they are amazed at the results I have achieved. The bottom line of all this is that you do not need a super station to work a lot of DX on Top Band. What you do need is a little imagination, ingenuity and perseverance to succeed and have a lot of fun."

What better introduction could I have than the above testimony of a dedicated Top Band DXer, who's not frustrated living in a beautiful (I must say) but fairly typical suburban house on a 1⁄2-acre lot. George did not use his QTH handicap as an excuse. No, for him it was just another challenge, another hurdle to overcome.

So don't lament if you don't have a multi-million dollar QTH.

Fig 14-1 — Showing a stealth antenna is easy — you show the sky. Rather than just the sky, here's the view at K2UO's QTH showing his low-profile 10/15/20-meter quad and his beautiful home in a wooded residential area in New Jersey. With his invisible 160-meter stealth dipole, George has worked nearly 250 countries on Top Band.

You can work DX on the low bands as well. Maybe you won't be the first in the pileups, but you will get even more satisfaction from succeeding, since you did have to take the extra hurdle!

My good friend George Oliva, K2UO, holds BSEE and MSEE degrees. He is retired from the US Army's Communications and Electronics Command Research, Development and Engineering Center at Fort Monmouth, New Jersey, but continues to work full time as a consulting engineer He got his first amateur license in 1961 and has operated from a few exotic locations such as Lord Howe Island, Guernsey, Turkey and even Belgium. He is a Senior Member of the IEEE and holds several patents.

George not only volunteered the above striking testimony, he also volunteered to godfather this section of the book, for which I am very grateful.

Another encouraging (for the city dwellers) testimony comes from Kumar, VU2BGS, who writes "I have had the pleasure of working on Top Band. Here comes the surprise: I run an IC-738 (70 watts) into a 10 meter vertical wire supported by bamboo and a fiberglass pole on top, using about 5 meters of top loading wire in two directions. This "monster" antenna is mounted on the roof of the two storied house. The roof space is 9 by 12 meters. I have six radials lying on the floor… This probably is the smallest antenna in the world for Top Band, but I must have the biggest diehard spirit to work Top Band. The noise level is always S9 +, and I find it a real challenge to even hear anything on the band! I was also happy to work your friend W4ZV with this on 160 meters. I hope some day I have some place to put one decent vertical of 20 meters at least."

1. THE PROBLEM

If you have decided to read on, this is not going to be news for you. But let me nevertheless describe the typical suburban antenna syndrome.

You have this wonderful house, in this wonderful-looking neighborhood, at the right driving distance from your work. A dream, however, may not be a ham's dream. There really isn't enough space for the three towers and the Four Square vertical array you would like to put up, and the neighbors would rather see trees growing than antennas. And your spouse won't really tell it to your face, but thinks one multiband vertical is more than enough. At the very best, one tower is what you can obtain your spouse's permission for.

If you really want to compete with the big guns on the HF bands, you need Yagis. Not a simple tribander, but mono-band Yagis. On the low bands though, you can be relatively competitive with rather simple antennas. Remember the above testimonies. This is good news! Read on.

2. SET YOURSELF A GOAL

Maybe you should set yourself a realistic goal for your circumstances. You can achieve satisfaction that way as well. Compete with your equals.

But there are nevertheless "fantastic" stories from average suburban QTHs. Here is another testimony of perseverance (or maybe addiction): "I was a young engineer working for IBM, just emigrated from Europe and lived until 1986 in a Toronto suburb, on a 46 by120-foot city lot surrounded by houses, TVI, power line noise and nasty neighbors. First I had a home-brewed 65-foot TV tilt-over tower with a used TH6 and 402BA and inverted Vs ($350). Later I thought I struck gold when I found a second-hand Telrex Big Bertha monopole with the antennas for $1200. I designed and built my own antennas (about $200 in material from junkyards). The rig was a used Drake B-line + R4C (about $500). All the rest of the station, the amplifier and the gadgets were home-brewed. I realized that I had a hard time beating the multi-multi stations in the contests, so I specialized in single-band operation. This netted me about 16 world records and all Canadian monoband records from 160 through 10 meters in the CQWW and WPX contests..."

All that from a 14 by 36 meter city lot (⅓ acre)! This was Yuri Blanarovich, VE3BMV, ex-OK3BU, now K3BU. But you are not that addicted? Keep on reading.

This book has explained propagation and focused on various types of antenna configurations for both receiving and transmitting. We have covered factors such as gain, polarization, radiation angle, incoming signal direction and angle, soil conductivity and the many other factors affecting receive and transmit performance. It is up to you as an individual to assess your own situation, set your own goals and use the information in this book in conjunction with basic engineering judgment to experiment in the true amateur spirit.

Every QTH has its own limitations, and you must apply your own skills to optimize your station based on your individual goals. Let's have a look at some simple but very effective antennas that might help overcome some of the limitations.

3. THE FLAGPOLE VERTICAL ANTENNA

A $\lambda/4$ vertical for 7 MHz measures 10 meters high, about the size of a really good patriot's flagpole. There you have a wonderful full-size 40-meter vertical. If the pole is a metal pole, make sure there is a good electrical contact between the different sections. If you are using a wooden flagpole, you will have to run a wire along the pole. It is best to use small stand-off insulators, so that the wire does not make contact with the wood. If your neighbor is curious about the wire, tell him it's part of a lightning protection system. Being a vertical antenna, the flagpole requires radials, but you can hide these in the ground so nobody should object. You should of course insulate the flagpole from the ground. If the flagpole is exactly resonant on 40 meters, you can probably feed it directly with a 50 Ω feed line. Chances are the flagpole may be a little shorter, so you can load it at the bottom with a coil.

An L network, as shown in **Fig 14-2**, will load and match the antenna at the same time. For a flagpole measuring 8 meters, typical component values (assuming a 5 Ω equivalent ground loss resistance) are: C = 500 pF and L = 2.8 µH. With a 10-meter long flagpole, no matching network will be required on 40 meters.

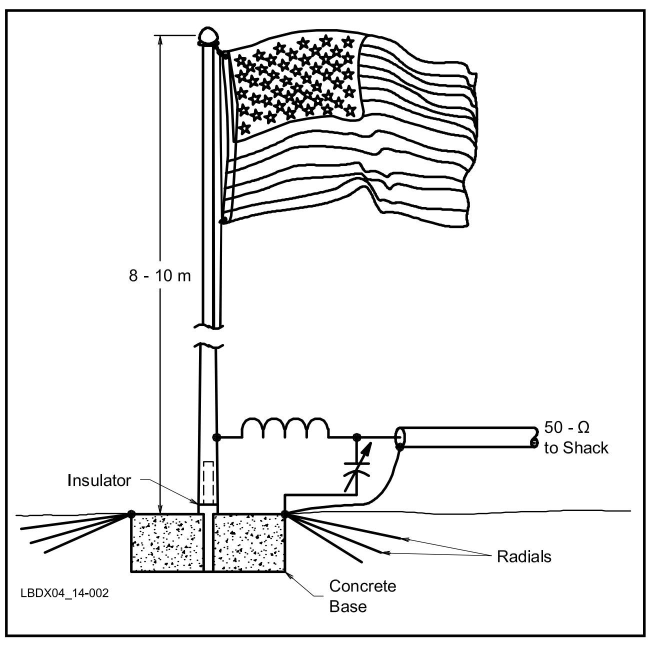


Fig 14-2 — Forty-meter flagpole antenna. Any metal flagpole between 8 and 10 meters high will do. Use an L-network to match to the 50 Ω feed line.

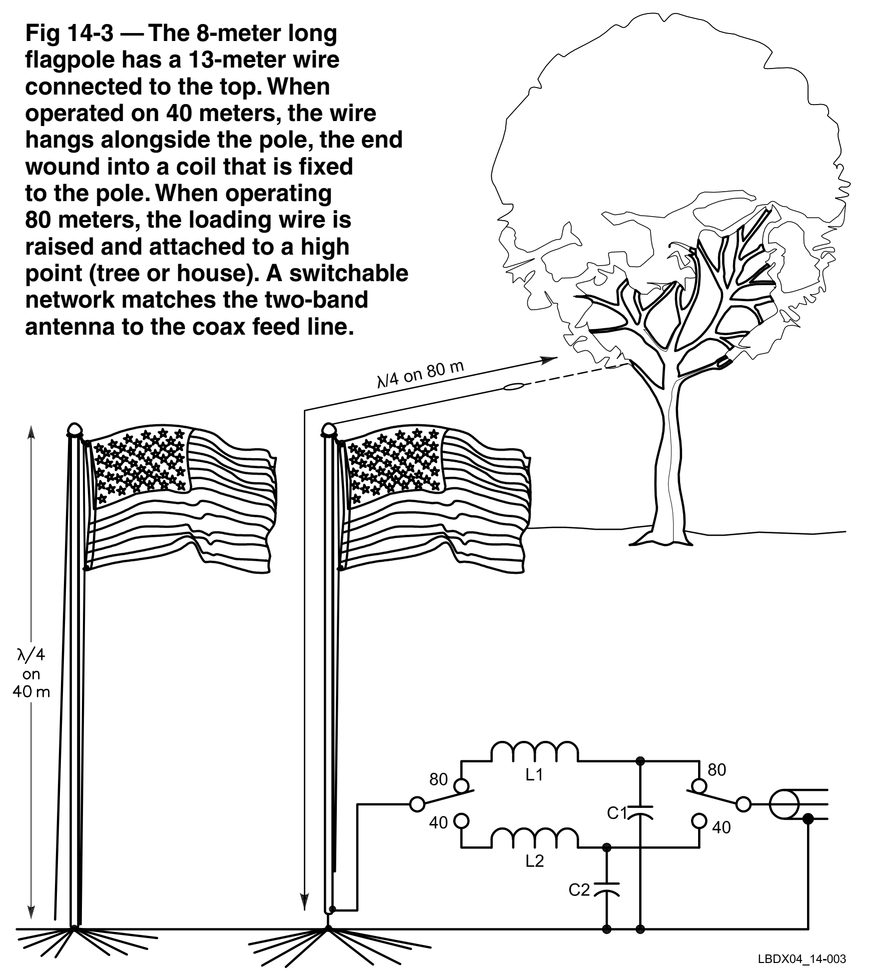


Fig 14-3 — The 8-meter long flagpole has a 13-meter wire connected to the top. When operated on 40 meters, the wire hangs alongside the pole, the end wound into a coil that is fixed to the pole. When operating 80 meters, the loading wire is raised and attached to a high point (tree or house). A switchable network matches the two-band antenna to the coax feed line.

How about 80 meters? You can transform the 8- to 10-meter tall 40 meter vertical into an efficient inverted L at night, if it has to be a super stealth antenna. See **Fig 14-3**. Connect the top loading wire to the top of the metal flagpole. When you operate 40 meters, or during daytime, hang the top wire along the flagpole (coil up the bottom end so that it does not touch the ground). When you want to operate 80, raise the wire with an invisible nylon fishing line and stretch it toward the house or a tree. The top loading wire can be any thin wire, since it hardly carries any current (all the current is at the base of the flagpole). Now you're all set on 80 meters. For this 40/80 meter flagpole antenna (using an 8-meter long flagpole) the typical L-network component value for 80 meters is: L1 = 1.1 μH and C1 = 1100 pF.

If your spouse or the neighbors won't object to a permanent tiny wire running from the top of the flagpole (maybe they haven't even seen it), install an 40 meter trap at the top of the flagpole. Disguise it by using your imagination. See **Fig 14-4**. Appropriate traps are described in Chapter 9.

For an efficient 160 meter vertical antenna you need at least 15 meters of vertical conductor. Have you looked at the trees in that corner of your lot? They should do as supports. Maybe you need to practice a bit with a bow and arrow, but if you can shoot a nylon wire over the trees, you're probably set for a good 160 meter antenna. If you use an inverted L or T antenna, you can use the tree-supported vertical on 40, 80 and 160 meters. And your neighbors will hardly see it! Don't forget that this antenna requires a good radial system. But you can put those down during the night.

Don't forget that the open ends of an antenna are always at very high voltage. If you run the outer ends of these wire antennas through the foliage toward a tree, it's a good idea to use Teflon-insulated wire. This will help prevent setting your trees on fire. And, by the way, all these wires don't have to be perfectly horizontal or perfectly vertical. Slopes of up to 20° will not noticeably upset the antenna performance.

You could of course also buy a commercial antenna, and spend lots of money for lots of loss. Use your imagination instead, and put your brains to work instead of your wallet!

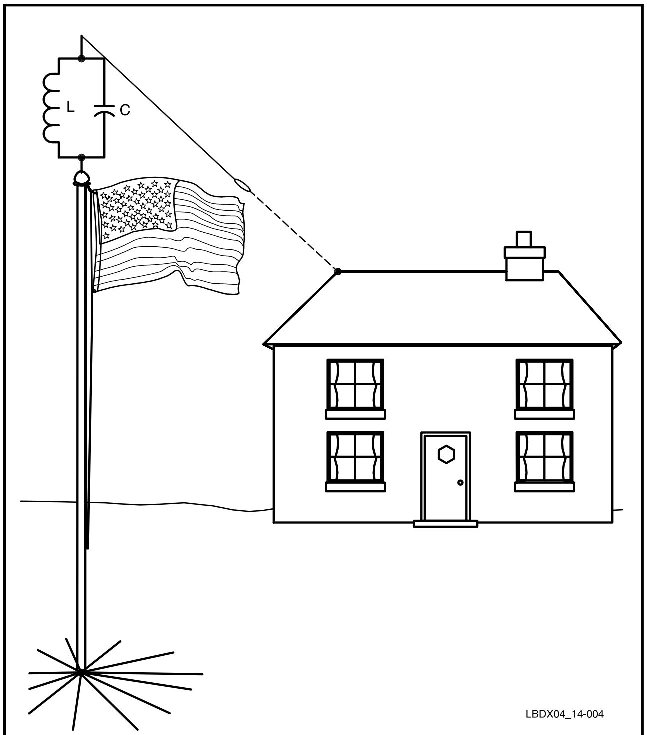


Fig 14-4 — With a 40-meter trap installed at the top of the flagpole you can get the 80-meter top-loading wire permanently connected. It can be directed to a tree, the house or any other available support.

4. LOADING YOUR EXISTING TOWER WITH THE HF ANTENNAS ON 80 OR 160 METERS

If you have a tower with one or more HF or VHF Yagis, you can probably turn it into an efficient vertical on 80 or 160 meters. A tower of about 15 meters with a simple tribander will give you the right amount of loading to turn it into an excellent 80 meter vertical.

For 160 meters you will need a little higher tower, but starting at about 18 meters with a reasonably sized tribander antenna will get you about 70° electrical length on 160. See Chapter 9 for details on how to shunt-feed these antennas.

If the tower is guyed, make sure the guy wires are broken up in short insulated sections. Short means about $\lambda/4$. Better still, use dielectric guy rope, such as Phillystran (aramid fibers) or glass-epoxy rods (fiberglass reinforced plastic or FRP).

If you use a crank-up tower, you will do better running a solid copper cable along the sections (an old coaxial cable will be fine), as the electrical contact between sections may not be all that good. If in doubt, measure the resistance.

It is imperative that you run the feed line and control cables inside the tower all the way down to ground level, and run them underground to the house. Otherwise it will be extremely difficult to decouple these cables. It is a good idea to coil all the cables at ground level, to provide enough inductance to form a good common-mode RF choke.

Don't forget that shunt-fed towers do require a good ground system. Run as many radials as you can in all possible directions. Don't be overly concerned if the tower is next to the house — you may lose a couple of dB in that direction but that's all.

5. HALF SLOPERS

Half slopers are covered in Chapter 10, Section 6. These antennas are popular with those who have a tower with a rotary antenna, and who want to get it working on 80 meters. A minimum height of about 13 meters (depending on the loading on top of the tower) is required to make a good vertical radiator on 80 meters. For a 160 meter sloper to work well, you would need a tower about twice that high. Don't forget that it is not the sloping wire that does most of the radiating, it is the vertical tower. The sloping wire merely serves as a kind of resonating counterpoise for the feed line to push against. As with all vertical antennas, the efficiency of a half sloper will depend primarily on the radial system used.

Don't feel tempted to use sloping wires in various (switchable) directions. As the sloping wire only radiates a small part of the total field, this effort would be in vain. As with shunt-fed towers, all cables that run to the top of the tower should run inside the tower, and run underground to the shack to maximize RF decoupling.

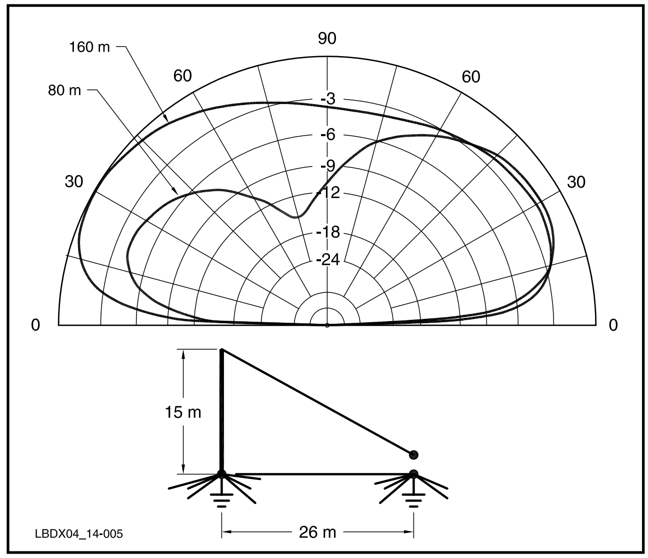


Fig 14-5 — Vertical radiation pattern for the half-sloper on 80 and 160 meters. See text for details.

6. HALF LOOPS

Half loops are covered in Chapter 10, Section 5. Fed at the bottom of the sloping wire, this antenna is attractive where space is limited. A 15-meter high tree could support the vertical wire, and from the top a slant wire can run to the shack or any other convenient place. If you use a 26-meter long sloping wire, the antenna will be resonant around 3.5 MHz and have a feed-point impedance of 60 to 75 Ω, good for a direct feed to the transmitter. To make it work on 3.8 MHz, shorten the total length of the antenna by approximately 3 meters, or simply feed it through an antenna tuner or L network.

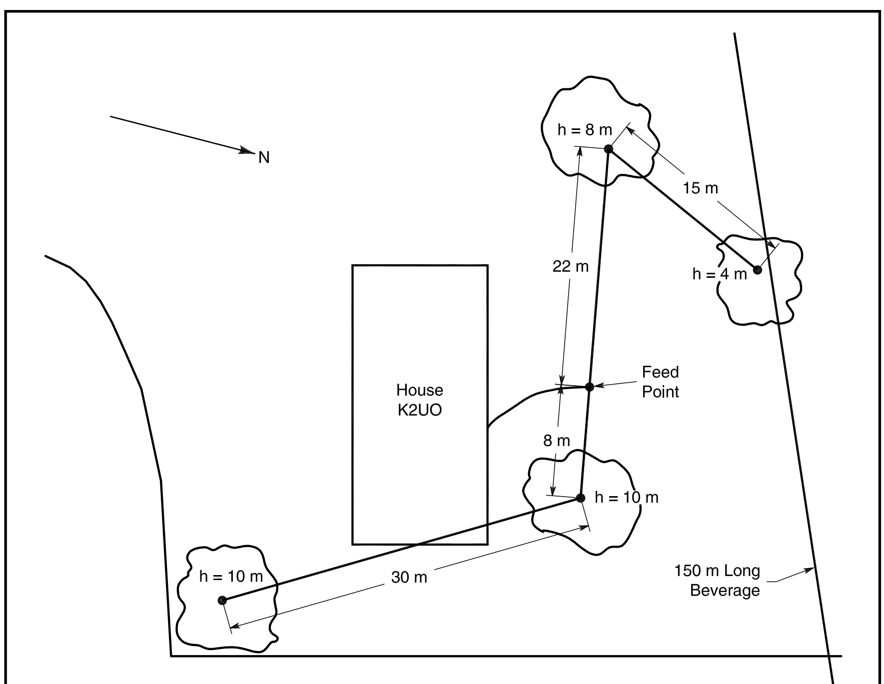


Fig 14-6 — Top view of K2UO's 160-meter stealth dipole, which is supported by trees and which is at no point higher than 10 meters!

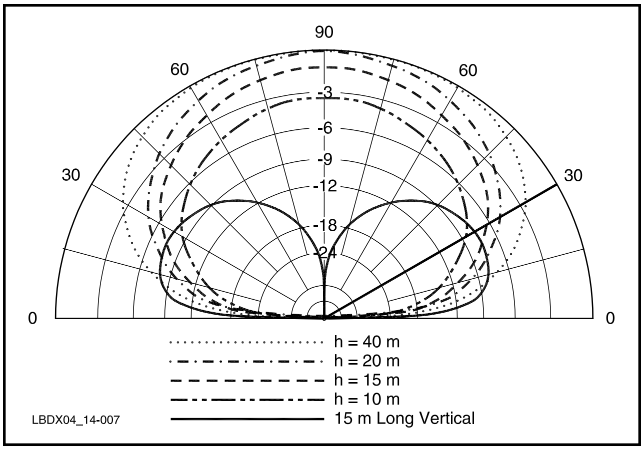


Fig 14-7 — Vertical radiation pattern of dipoles at various heights, compared to a short 15-meter long vertical with 5 Ω equivalent ground-loss resistance.

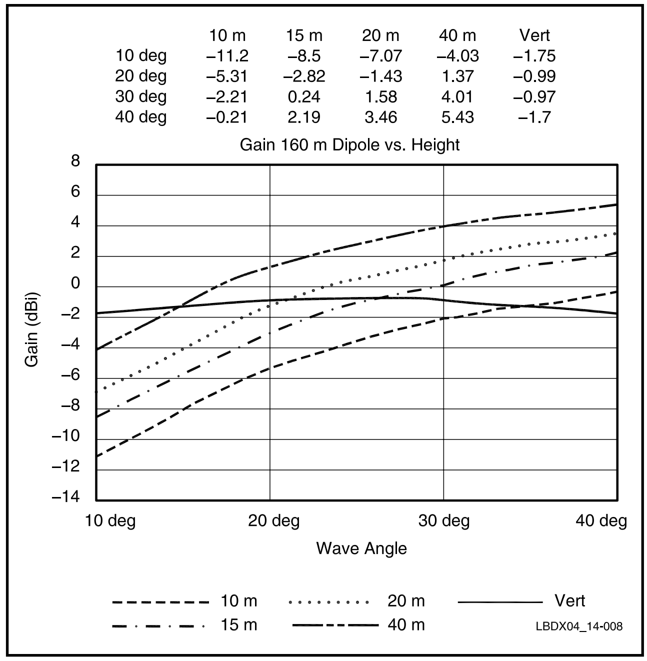


Fig 14-8 — Gain of low dipoles compared to a reference 15-meter long vertical.

This antenna will also work quite well on 160. Its feed impedance will be very high, however. The best feed system is to use a parallel-tuned circuit as shown in Chapter 12, Fig 12-20. Needless to say, the feed point is at very high RF voltage, and the necessary precautions should be taken to prevent accidental touching of the antenna at this high voltage point.

Fig 14-5 shows the radiation patterns for this half-loop for 80 and 160 meters. The antenna shows some directivity on 80 meters, about 4 dB in favor of the direction of the sloping wire. Again, a good ground system is required for this antenna, at both ground connection points.

7. VERY LOW TOP BAND DIPOLES

The saying is that very low dipoles (10 meters high) are only good as receiving antennas. Is that so? **Fig 14-6** shows the layout of K2UO's *Zig-Zag dipole* for 160. When you walk around his lot, you can hardly see the wire. It really is a stealth antenna, but it has given George almost 250 countries on Top Band. And that's not only "heard" countries, but those worked and confirmed!

In this book I have described high dipoles as efficient low-angle radiators. In order to be competitive with vertical antennas at really low angles, a dipole must be at least λ/2 high. I think we will hardly ever find such high antennas on a typical suburban lot, though! But low dipoles can still function quite well on the low bands. The antenna at K2UO is an outstanding testimony for such low dipoles.

Fig 14-7 shows the vertical radiation patterns of low λ/2 dipoles, compared to a 15-meter long vertical (R_{rad} = 17 Ω) using a fairly decent radial system (R_{loss} = 5 Ω). A 160-meter dipole between 10 and 15 meters high produces the same signal as our reference vertical (±1 dB) at a wave angle of 30°, which may come as a surprise to some. At very low angles (10°), the vertical will be 13 dB better than the 10-meter high dipole. **Fig 14-8** shows the gain of the various antennas for wave angles of 10°, 20°, 30° and 40°.

Looking at the patterns in Fig 14-7 we see that the big difference is in the high angles. The low dipole will be much better than the vertical for local coverage, but that means also that the signals from local stations will be much stronger than they would be on a vertical. Although the dipole may have the big advantage of reducing man-made noise (which is generally vertically polarized), it has the disadvantage of producing very strong signals received at high elevation angles.

What may come as an even bigger surprise is that we have learned that not all (though most) of the DX on Top Band comes in at very low angles. Especially on 160 meters, however, we know that gray-line enhancement at sunrise or sunset often coincides with an optimum angle of radiation that is rather high, and that definitely gives the advantage to the low dipole. So, you might even beat the big gun with his super low-angle antenna, using a K2UO-style dipole!

As a rule I'd like to stress that it is important that you keep the center of the antenna as clear and as high as possible. The ends are just "capacitance hats" — they don't really radiate a lot, so you can bend and hide them as appropriate without hurting the antenna's performance a lot. If you don't have room for a straight 80-meter long dipole (who has?), rather than loading it with coils, or using a W3DZZ-type dipole, just bend the ends. That's much better, and will introduce less loss than the usual lossy coils. What holds for 160 meters is of course applicable to 80 as well.

K2UO is certainly not the only one who's been successful with low dipoles. Recently I read a similar testimony from Ivo, 5B4ADA (ex-HH2AW): "My 160 meter antenna is 1/10-λ high (apex of the inverted V is 16.5 meters above ground, wire ends are 1.5 meters above ground). Theoretically, it radiates up most of the RF, but I still have fun working USA, JA, VK, breaking the XW3Ø pileup, etc. I had a 57 meter long wire in Haiti on a bamboo pole 10 meters above around. Worked many USA and EU stations on 160. Don't be scared with too much theory, get on the air..."

I would not necessarily agree with the "theory" part of Ivo's statement, since the theory does predict that low dipoles are a viable alternative...to nothing at all.

8. WHY NOT A GAIN ANTENNA FROM YOUR SMALL LOT?

8.1. An Almost Invisible 40-Meter Half-Square Array

I am convinced that on 40 meters you can put up this almost invisible gain antenna. You need to be able to run a horizontal wire about 10 meters up, and 20 meters long. Perhaps from the chimney of the house to a tree in the corner of the lot. **Fig 14-9** shows a 40-meter half-square array that can be squeezed into many small lots. Gain is approx 3.4 dB over a single full-size $\lambda/4$ vertical. The ends of the vertical wires are also at very high RF potential, and precautions should be taken to prevent accidental contact. The half-square is fed via a parallel-tuned circuit as shown in Chapter 12, Fig 12-20.

You can also feed the half-square in one of the top corners. This may be a good idea if one element is close to the house as shown in Fig 14-9B. When fed in the corner, the feed impedance is about 52 Ω, a perfect match for a 50-Ω feed line. Do not forget to install a current balun on the coaxial feed line. **Fig 14-10** shows the radiation pattern for the 40-meter half-square.

8.2. Using the 40-Meter Half-Square on 80 Meters

What about using the 40 meter half-square on 80 meters? A bit of magic turns the antenna into two close-spaced in-phase fed end-fire arrays with top loading. The only thing you need is to short the base of the second element to ground and feed the array at the other element at ground level (**Fig 14-11**).

This 2-element array has a gain of 1.6 dB over a single full-size (20 meter high) vertical and provides excellent low-angle radiation. The antenna has about 4 dB front to side ratio. Its feed-point impedance is about 70 Ω excluding ground-loss resistance at each vertical element. This antenna requires a good ground radial system at the base of both elements.

With some ingenuity you could homebrew a switching system that grounds or ungrounds one element, and either feed the other element directly with coax on 80 meters, or feed it via a parallel-tuned network on 40 meters.

8.3. 40-Meter Wire-Type End-Fire Array

Maybe the half-square doesn't suit your most wanted direction. You can also turn this into a 2-element parasitic array as shown in **Fig 14-12**.

I worked out the example of an array where a maximum height of 8 meters was available as the catenary wire. The elements have a top loading wire which really makes them inverted-L verticals (see Fig 14-10 and 14-11). The array has a very good F/B and gain, and a feed-point impedance of about 25 Ω. The gain is approximately 3 dB over a single vertical, and the F/B is an impressive 25 to 40 dB. See **Fig 14-13.**

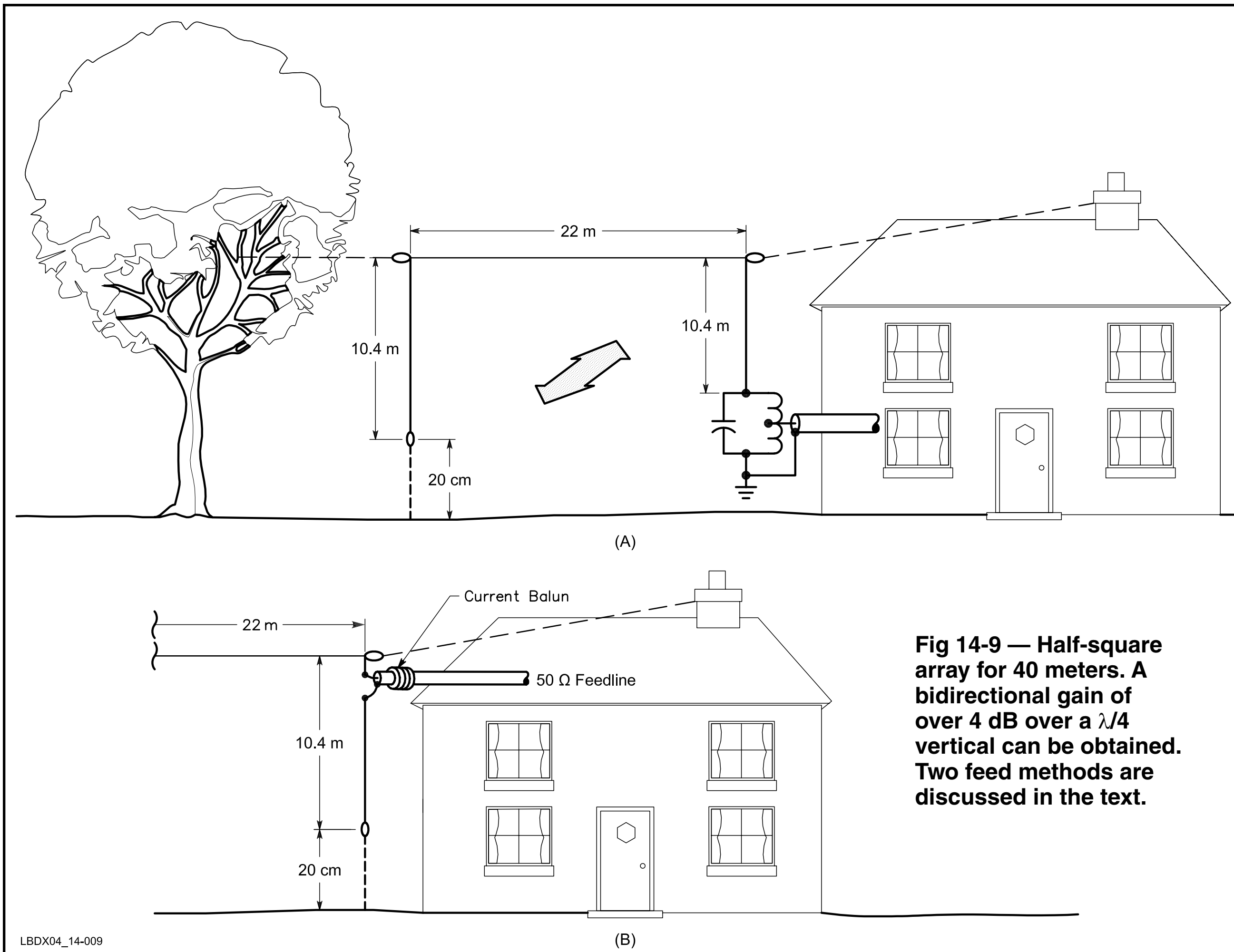


Fig 14-9 — Half-square array for 40 meters. A bidirectional gain of over 4 dB over a $\lambda/4$ vertical can be obtained. Two feed methods are discussed in the text.

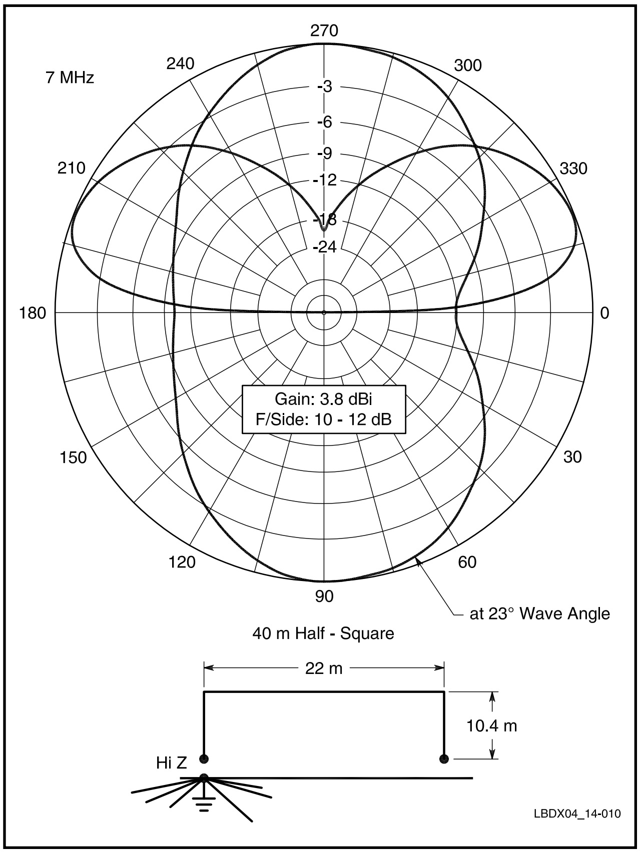


Fig 14-10 — Radiation pattern for the 40 meter half-square.

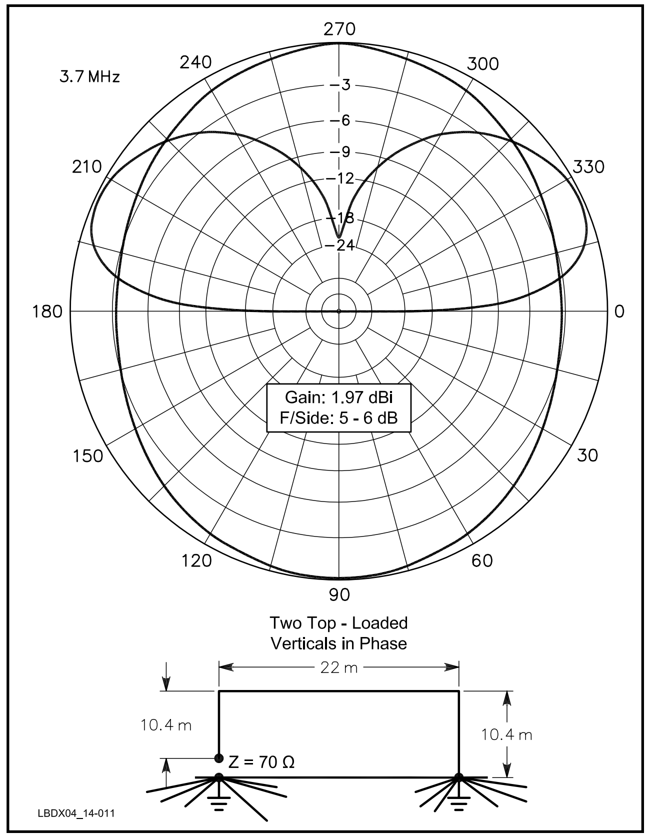


Fig 14-11 — The 40-meter half-square can be turned into a 2-element close-spaced top-loaded array for 80 meters, where both elements are also fed in phase. Both vertical and horizontal patterns (at 30° elevation) are shown.

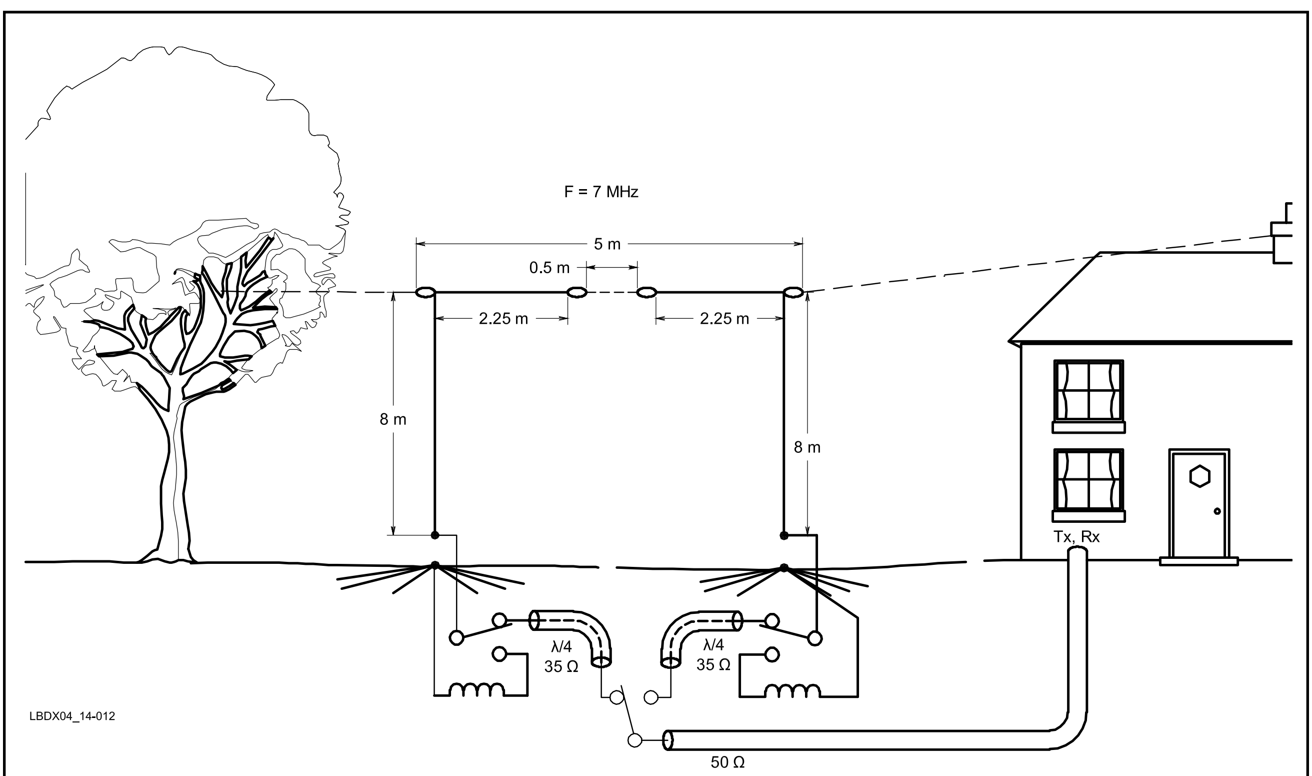


Fig 14-12 — Switchable 2-element parasitic array for 40 meters.

Matching can be done through a λ/4, 35 Ω line, consisting of two parallel 75 Ω coaxial cables (each measuring 7.03 meters for RG-11 or RG-59, both having VF = 0.66).

It is important to install as good a radial system as possible on this array. Where radials from the two elements meet they can be connected to a bus wire, as shown in Chapter 11.

8.4. An 80-Meter End-Fire Array

Maybe you have two high trees in the back that could help you support a 2-element array for 80 meters. I would recommend a minimum height of the elements of approximately 13 meters; the remainder can be top loaded if necessary. **Fig 14-14** shows two T-loaded 13-meter high verticals, suspended from a single catenary rope, for example between two tall trees.

This array has an excellent F/B ratio and gain and will certainly put you in the front seat in a pileup if you take care to install a good ground system. When properly adjusted, the array impedance, assuming about 5 Ω equivalent ground loss resistance, is about 20 Ω. The array can be fed via a 37.5-Ω, λ/4 transformer (two parallel 75-Ω cables) as shown in Fig 14-12, or via an unun transformer (20 to 50 Ω).

It is important that the top-loading wires are as shown (points facing one another). If you are forced to try another configuration, I would advise you to model the array exactly as in reality. Needless to say, if you have some really tall trees on your property, this antenna can be scaled up for 160 meters.

8.5. The Half-Diamond Array

Maybe you don't have the two supports required to put up the box-shaped arrays I described above (Section 8.1). With just one support, a few good arrays can be created as well. You will require one high support (15 meters), a tree for example. In **Fig 14-15** I've reshaped the half-square to become a half-diamond. You lose about 1 dB gain, but the pattern remains unchanged.

8.6. The Same Half-Diamond Shaped Array on 80 Meters

The same inverted-V shaped half-diamond 40-meter array can be used on 80 meters as well. As with the half-square (see Section 8.2) all you must do is ground the bottom end of the array at the side opposite to the feed point (see **Fig 14-16**).

This array is really like a delta loop with its bottom wire lying on the ground. Needless to say, this is again proof that the delta loop is the equivalent of two sloping verticals, fed in phase (see Chapter 10, Section 2).

The array has an impedance of about 75 Ω. The exact resonant frequency can be tuned by simply changing the total length of the antenna. For use on 40 meters the exact length

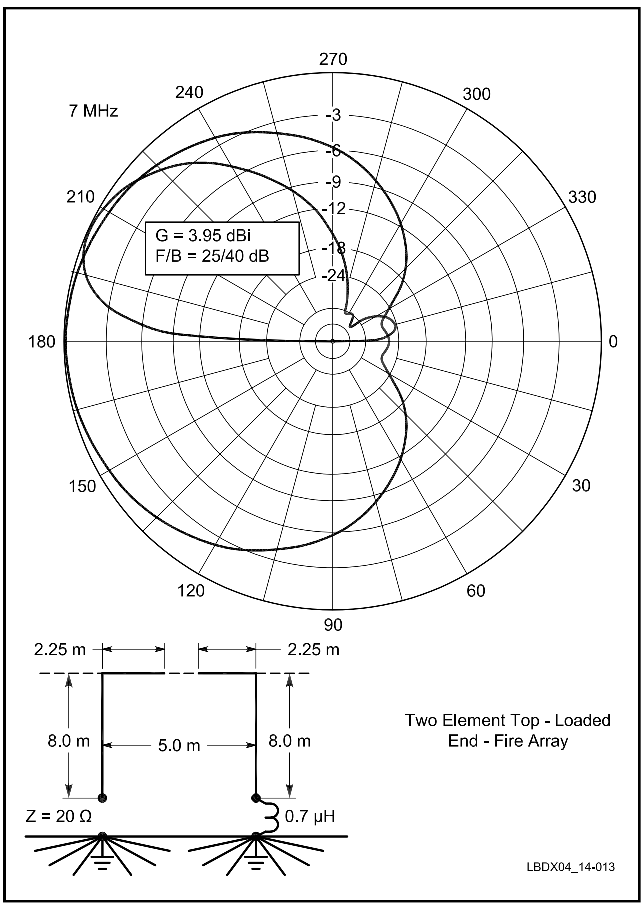


Fig 14-13 — Horizontal radiation patterns for the 2-element parasitic array of Fig 14-12.

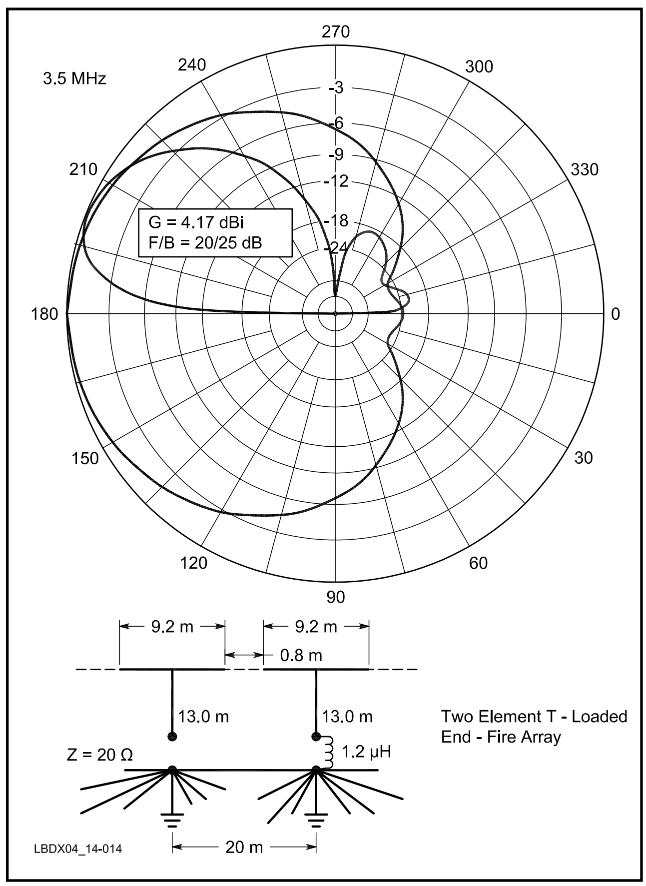


Fig 14-14 — Horizontal radiation patterns for the 2-element T-loaded parasitic array for 80 meters. The gain is over 4 dBi and the F/B is 20 to 25 dB.

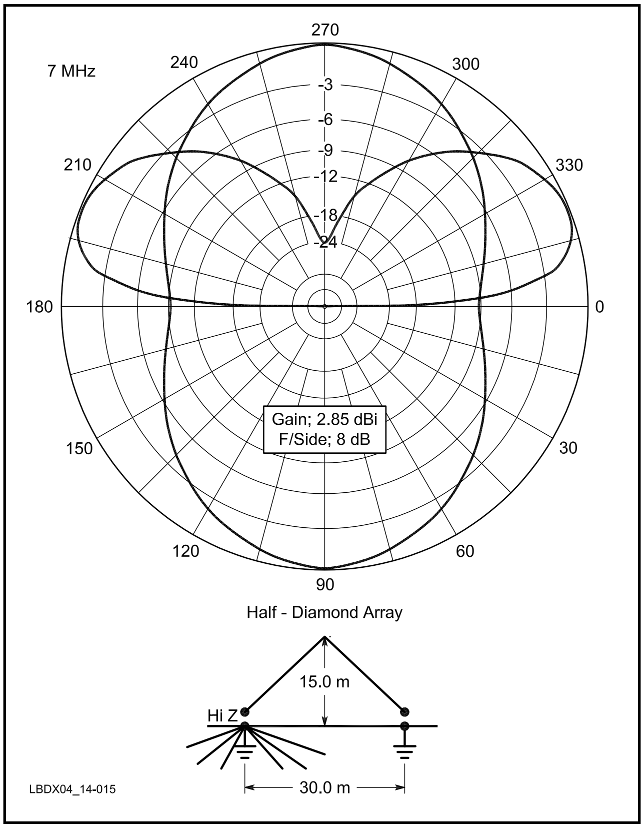


Fig 14-15 — Using a single support, you can turn the half-square array into a half-diamond array, at the sacrifice of about 1 dB of gain.

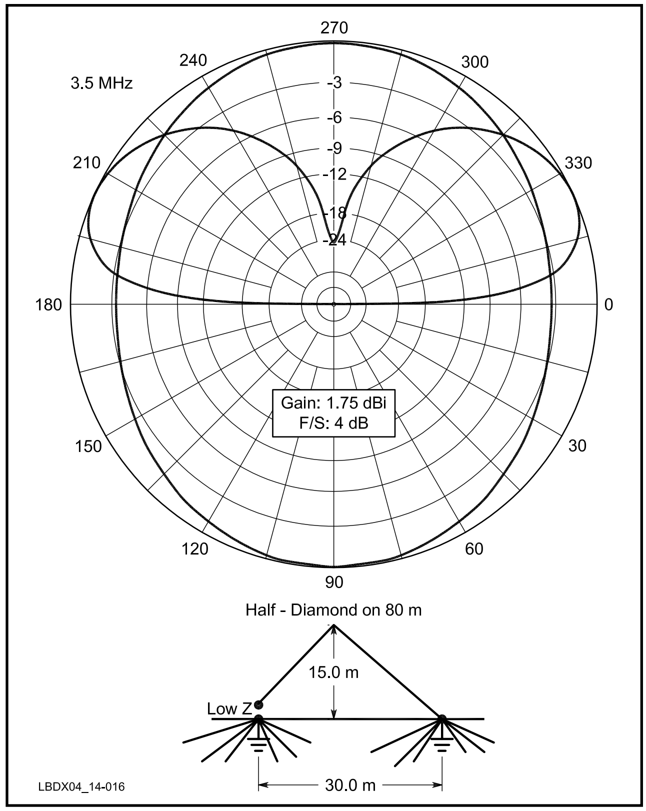


Fig 14-16 — This 80-meter capacitively loaded midget delta-loop shaped array requires only a 15-meter high support. With a 14-meter high support, and sloping the antenna about 30°, the radiation pattern will not be overly affected.

is not so critical, since the array can easily be tuned for low SWR anywhere in the band using the resonant tuning circuit.

8.7. Capacitively Loaded Diamond Array for 80 Meters

Maybe you can't quite get a height of 15 meters. One solution is to top load the two sloping verticals with a common capacitive wire, hanging right down as shown in **Fig 14-17**. Two wires are connected to the apex of the V. Their length and the angle between the wires is varied to tune the antenna to the required frequency.

The gain of this antenna is still about 1.5 dBi, which is more than 1 dB better than a single full-size (20 meter high) vertical over typical ground. The feed-point impedance, including about 10 Ω loss resistance, is about 50 Ω!

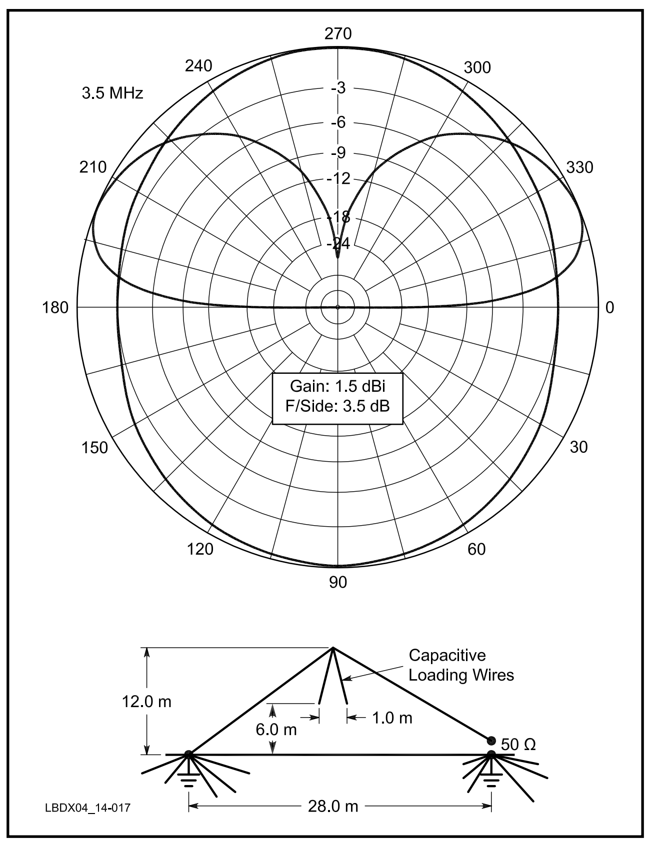


Fig 14-17 — This 80 meter array requires only 12 meters of height. The array is capacitively loaded at the top, using two wires in a V shape.

8.8. A Midget Capacitively Loaded Delta Loop for 80 Meters

The half-diamond antennas described above look very much like a delta loop with its bottom wire laying on the ground. Raising the wire turns it into a real delta loop.

The model shown in **Fig 14-18** has similar dimensions to the antenna in Fig 14-17, and yields the same gain, the same front-to-side ratio, and even the same feed-point impedance. Needless to say, this is additional proof that the delta loop is the equivalent of two sloping verticals, fed in phase (see Chapter 10, Section 2).

In this example, I used capacitive loading in a little different way. This delta loop can be tuned anywhere from 3.5 to 3.8 MHz by just changing the length of the bottom capacitance wire. L1 is 11 meters, and L2 is 6 meters for f = 3.5 MHz. For f = 3.8 MHz the bottom loading wire can be eliminated altogether.

There is nothing magical about these dimensions. Just keep in mind that the capacitive loading wires should be attached at the high-voltage points, and that they carry very high voltage. As a safety precaution, the horizontal wire, at only 1 meter above ground, should be kept out of reach of children and animals, as the center can carry very high voltages. The capacitance loading wires should be kept at a distance of at least 20 cm from each other.

Do not fool yourself into thinking that this antenna requires no radials. The radiation is affected just as much by near-field absorption losses under the antenna as is the case of the grounded verticals. In other words, delta loops require a ground screen, just as is the case with all antennas that have radiating elements close to ground!

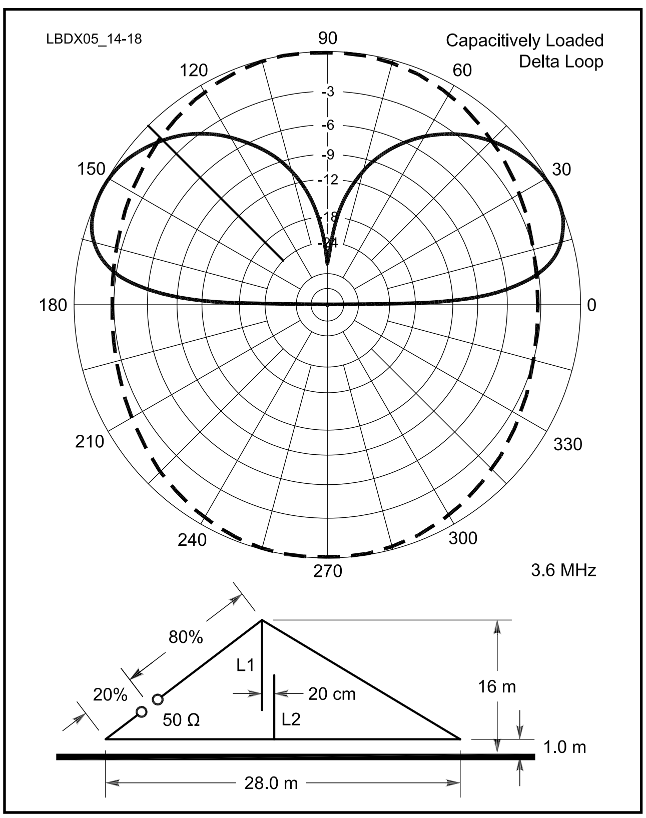


Fig 14-18 — This 80-meter capacitively loaded midget delta loop antenna requires only a 16-meter high support. With a 14-meter high support and sloping the antenna about 30°, the radiation pattern will not be overly affected.

9. ALTERNATIVE RECEIVING ANTENNAS

I would consider Beverages, flags and pennants as "special" and this as "alternative."

Typical suburban QTHs mean rather dense housing, which in turn means a lot of man-made noise. Now that you have used your imagination, and squeezed an efficient vertical — or even a couple — onto your lot, you're faced with a very high noise level. There are basically four ways to tackle this problem:

- Use a horizontally polarized receiving antenna to reduce man-made noise generated nearby (coming in on ground wave).
- Use special receiving antenna(s) that have directive properties (small arrays, snake antenna).
- Use a noise canceller device.
- Locate the offending noise sources and kill them (the noise sources, that is!).

Jim McCook, W6YA, swears by his very short rotatable dipole on top of his tower. Such a small dipole would fit almost every lot. You can actually rotate it as it has excellent rejection when its ends are turned toward the directions of the noise source. However, any low horizontal wire will probably be better than your vertical for receiving.

How about a Beverage? I know your property is not quite like a Texas ranch, but maybe you can run one or even two short ones along the property line. Maybe you can talk your good neighbor into a concession. George, K2UO, has room for only one (150 meter long) Beverage, which partly runs along the property lines of his neighbors. But he says this antenna really was an eye-opener. Even $\lambda/2$ long Beverages are better than nothing!

On-the-ground Beverages (BOGs, see Chapter 7, Section 2.12) have a much lower velocity factor, which means that a "short" BOG works as well (directivity wise) as an "elevated" wire that is 50% longer! And, you can easily hide them, even (barely) bury them in your neighbor's lawn (with his permission of course…).

How about EWEs, flags, K9AYs and the like? They are quite popular where space is limited. Don't forget that these antennas should be clear from any transmitting antennas. I would recommend $\lambda/4$ as the minimum distance between a special receiving antenna and your transmit antenna. I know you can hardly ever meet this rule from a suburban lot, but there is a solution: "detune" your transmit antenna on reception (see Chapter 7, Section 2.11.1 and Section 3.11).

I have been extremely successful in eliminating man-made noise (coming in on ground wave from a chemical plant 10 km away) by using a so-called *noise eliminator* or *noise canceling device*. This is nothing but a circuit in which you combine the inputs from two antennas, the main receiving antenna and a noise-pickup antenna, in such a way that they are added out-of-phase, resulting in cancellation of the noise (see Chapter 7, Section 1.36). The MFJ-1206 unit and DX Engineering model DXE-AAPS-1P can perform very well in such circumstances.

Steve WB6RSE lives on a 1/7 acre (approx 20 by 30 meter) lot on the west side of Los Angeles. That is a small lot by any Top Band standard. Despite this limitation Steve has 238 DXCC countries on 80 meters and 162 on Top Band!

Let him explain how he does it:

"Although working Europe on 160 is very difficult because of polar absorption, I have managed 29 countries in Europe in 9 seasons of activity on that band.... The only place for dedicated RX antennas is on the roof of the house. The 14 × 29 foot flag design of W7IUV proved to be a very good receive antenna, but my mechanical design had problems and I had to take it down. It was replaced by a mechanically simpler diamond loop which worked well but had a much smaller capture area. Before I had tried EWEs and K9AYs but these need a ground connection and are not the right solution for roof mounting. You need a ground-independent loop such as a flag. I now rebuilt the 14 × 29 foot (4.3 × 8.7 meter) flag utilizing a squat quad element design. This is simpler and probably lighter than W7IUV's configuration and yields a nicely shaped rectangle. You need to use the right matching transformer. After having tried many inferior transformer designs, the 8t/2t transformer on the binocular core (see Chapter 7) proved superior, making these loops actually sound like real antennas."

Wow, 162 DXCC countries on 160 meters in less than 10 years, from a ⅐ acre city lot. You now have no excuse!

What you need to be successful under less than ideal circumstances is: patience, perseverance, imagination, some technical know-how and the drive to succeed.

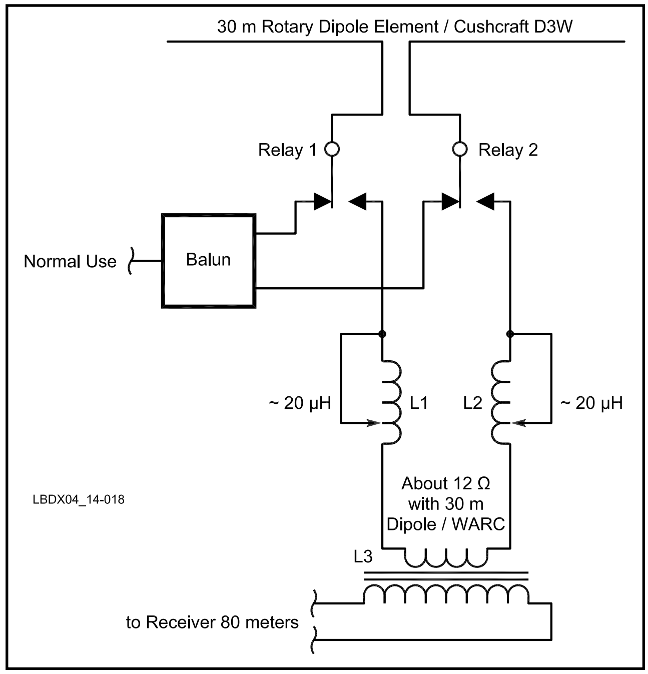


Fig 14-19 — W6YA uses a 30 meter rotary dipole, which he resonates on 80 meters with loading coils. The same could be done for 160 meters. As losses are quite irrelevant in a receiving antenna, the Q of the loading coil is not so important.

10. POWER AND MODE

Power and mode both factor into the results you will receive in terms of working DX as well as "working" your neighbor's TV/telephone/radio, etc! The more power you radiate, the more signal will be available at the DX station's antenna. However, look at the goals you have set for yourself. Do you really need to be the first one to get through the pileup? If the answer is yes, you need power; otherwise, you won't. Again, the question is not how much power is coming out of your amplifier; it's how much power are you radiating and are you radiating it in the right direction at the right angle? You must also remember that it is much easier to work DX on CW than on SSB, especially on the low bands. With few exceptions, DX on Top Band is on CW.

It is obvious that the more power you run the easier it will be for you to work DX. That does not mean you will get the most satisfaction from your results, though. Also, the more power you run, the more chances there are of creating some kind of TVI, broadcast interference or telephone interference problem in the neighborhood. The first and easiest way to avoid similar problems is not to run high power. But, if you are already handicapped with your pocket-size lot, some power may be one of the few available means to get that elusive DX on the low bands after all. CW as a mode also creates fewer audible interference problems than SSB (unless your neighbors copy CW).

11. ACHIEVEMENTS

You have read about K2UO working almost 250 countries on 160 meters with a Zig-Zag Stealth dipole that's nowhere higher than 10 meters. And about WB6RSE from the "difficult" West Coast on his way to 200 countries on Top Band from a ⅐ acre city lot. In Chapter 13 (Section 4.5) I described W6YA's 21-meter tower, on a typical suburban lot, which carries antennas for all nine bands! (**Fig 14-19** shows how W6YA resonates his 30-meter dipole on 80 meters.) Hams all over are successful in DXing on the low bands from locations that are far from ideal.

Don't let space restrictions scare you away from the low bands. If you work the elusive 3B7 on Top Band from your small lot QTH, you will get triple the satisfaction of the big gun who got through a couple minutes earlier.

Some time ago I read a very applicable statement: "I was always complaining about my shoes, until I saw a man without legs." Let's have fun with what we have and do our best under the circumstances. Be convinced you're not the only one who's not living in ham's paradise. There are many others in the same boat.

12. THE ULTIMATE ALTERNATIVE

Don't put up any antenna, don't get on the air from your pocket-sized QTH, but save your money and energy and go on a DXpedition once or even twice a year, and provide us hard-core Low Banders with the new countries we need to be able to prove we're the best!

Of course, there's another solution, if you can afford it. Set up a remote station. I have seen wonders of ingenuity and engineering when visiting some of the top remote-controlled stations on the low band in the US: K9DX, W6RJ and N7JW/K7CA come to mind. (See Chapter 11.) While that solution is out of reach for all but a very few, it shows what can be done. The sky is the limit!

13. SUMMARY

This chapter has provided many potential solutions for being successful on the low bands when you do not have the optimum space for both receive and transmit antennas. By using the techniques detailed in the chapter, you should be able to achieve successful results on the low bands. You will need to select and then refine the transmit and receive antenna designs that best fit your individual circumstances. *See you on Top Band!*

Contents

CHAPTER 15

From Low Band DXing to Contesting

When I wrote the First Edition of *Low Band DXing* almost 25 years ago, I was a very active, omnipresent low-band DXer. A few years later we finally got access to Top Band and I really lived and slept on the low bands. Some 15 years ago, I shifted from being a DXer to becoming an avid contester. How did that come about, and why?

My neighbor and friend Peter Casier, ON6TT (of Peter I and Heard Island fame, and others) stirred up my interest in *contesting*. Peter achieved "mission almost impossible" by helping us in Belgium obtain permission from the authorities to operate with high power during international contests. At last we could compete equally with contesters in other countries and at last we did not have to fib about our power levels! While our authorities had been frightened about high power, the experiment with 2 kW licenses for international contests in 2001 led to high-power (1 kW) licenses for all full-license hams in Belgium. Over a period of nearly 20 years when 2 kW was allowed only during these contests, not a single complaint about interference was received by the monitoring authorities. Why? Good contest stations are usually built by good engineers and technicians who know what they are doing!

Frank Grossman, DL2CC, was my advisor on this chapter.

For me, building a competitive contest station has been a unique experience: Getting help from friends and club members and building a team of excellent operators. It not only widened my technical horizons but also my social horizons in terms of working with people, enjoying Amateur Radio as a group and making new friends from all over the world. One of the highlights of nearly half a century being a ham was being part of three WRTC (World Radio Team Contesting) competitions, either as a competitor, a referee or just a visitor, and finally of being inducted to the CQ Contest Hall of Fame in 1997.

Besides Mark Demeuleneere,ON4WW, and Peter (ON6TT), both local neighbors and avid contesters, there are two more friends I met through contesting that I would like to mention in particular. They both played an important role in my becoming an avid contester.

I first met Harry Booklan, RA3AUU, at the Clipperton Club convention in Bordeaux, France, in 1993. Harry was then 23 years old and won both the CW and phone pile-up tests. Harry now has his own telecom company in Russia and has become a very successful businessman. He quickly became my friend as well as one of the fixed assets of our OTxT contest station (now OT7T). As ON9CIB, Harry and I represented Belgium at WRTC in San Francisco in 1997. Harry ended second in WRTC 2000 in Slovenia, and second again in WRTC 2002 in Helsinki. He is "big" in contesting. He's a good friend as well.

Frank Grossmann, DL2CC, is another highly esteemed operator and friend of the house. Frank is a young computer-programming professional who runs his own business in Stuttgart. In the 1990s he used to make frequent 700 km trips (one-way) a year to operate contests from here. Frank was third at WRTC 2002 in Finland.

Unfortunately I have had to cut back on my contesting activities and multi-single operations are no longer possible. I am approaching the blessed age of 70, and unfortunately my XYL Frida's health is not like it was 20 years ago. But I still greatly enjoy the 160 and 80-meter contests, every bit as much as in the past, and occasionally welcome foreign visitors for doing a "limited scale" effort, such as George Oliva, K2UO, setting a new all time European record on 160 meters during the CQ World Wide Phone contest in 2008 (2nd place worldwide).

Both Harry and Frank are superb CW operators. Frank used to be Germany's high-speed champion some years ago. They both turned me into a CW addict, a wonderful addiction that I am proud to admit.

For the third time Frank, DL2CC, graciously agreed to help as counselor and critic for this chapter. Thank you Frank!

1. IN SEARCH OF EXCELLENCE

Amateur Radio is all about satisfaction and self-fulfillment.

Gaston, ON4GV (SK) was my Elmer. He also was my uncle. At his home I saw my very first Amateur Radio station (**Fig 15-2**). I was not quite 10 years old. That was around 1950. I am eternally indebted to him for giving me the virus.

Fifty years ago Amateur Radio in my eyes was adventure-land and wonderland, all in one. To a young boy, telecommunication (the word probably was not invented yet) was Amateur Radio. You must realize that in the early years after WW II, out in the countryside where I lived with my parents, we still had hand-cranked telephones with manually operated telephone exchanges. These exchanges closed down at 10 PM — we had no chance to telephone anyone during the night. If you wanted to communicate with someone across the ocean you wrote a letter. If you wanted to travel across the water, you took a boat.

But for all I knew, if you wanted to talk to someone anywhere in the world, you needed to be a radio amateur. Being a ham made you an explorer, a discoverer. You could expose yourself to worlds others had hardly ever heard about.

It was this magic, this thrill of radio that lured me into this hobby.

I will never forget the oh-so-typical smell of Bakelite, wax and tar-filled capacitors and transformers that was very typical for the early-day radios. And the white filament glow of early tubes. Some of the early-day triodes were so “brilliant” that you could literally read a book by them at night. My very first hands-on experiences with electronics (not that it was called that, in those days; we called it simply “radio”) were in building small audio amplifiers, using directly heated triodes, such as the E, A416. My father’s wooden cigar boxes served as chassis for building three-stage audio amplifiers using heavy 3:1 interstage transformers. This was all about discovering an amazing and intriguing world, the world of radio.

It was technology (a modern word for these early-day sensations) that hooked me to Amateur Radio.

For a while my discovery trips were somewhat curtailed and it was not until I was 20 years of age, in high school, that I finally got my license (1961). The challenge then was to prepare for the license, and get it on the first try. The sense of fulfillment, once you got it, was enormous, as was the first antenna you built, the first QSO you made. I was doing something not everyone else could do! Now I was part of them.

Fig 15-2 — A 1965 picture showing ON4GV (SK) — the author’s Elmer — with ON4UN and his wife-to-be. In the background we recognize a Drake 2B, an SB-1000 and an NCL-2000.

In the early 1960s I stumbled across some guys working DX on 80 meters. I remember a few calls: GI6TK, GI3CDF, GW3AX and G3FPQ. Only David, G3FPQ, is still there and still a very active low-band DXer. This really seemed like something else — working across the pond and into New Zealand on frequencies where the others would work stations in a 500-km range. What a challenge! This really put you in a separate class among hams.

Working the elusive DX on the low bands was my next challenge, and it became my passion. This time I found out that, in order to be part of those low-band DXers, you needed to have a good signal. That meant you needed to have the know-how to do it. Amateur Radio was no longer a communicating hobby for me, but became an experimenter’s hobby. That meant building new, better and bigger antennas, experimenting with and learning about propagation, becoming a better technician and becoming a better operator.

DXing on the low bands is all about overcoming difficult hurdles. Everybody can work DX on 10, 15 and 20 meters. For me, there is not much sense of satisfaction involved. In 1987 my last iron curtain was lifted. We finally got 160 meters in Belgium. The last frontier. A vast terrain for chasing difficult game. And yes, Top Band certainly is where the DXer can get the ultimate sense of satisfaction. Technical knowledge and technical achievements are undoubtedly great assets in achieving success in low-band DXing. But, even with a modest station, if you have dedication, patience and operating experience you can be a successful DXer.

If so, what can provide you with the ultimate sense of achievement, of technical excellence, in Amateur Radio? Throughout history, competition has been one of the important motivations for progress in many fields. So is contesting to Amateur Radio. To be a very successful low-band DXer you need to be a very good operator, know propagation, be patient, be persevering and have a “decent” antenna system and station. You can determine yourself whether or not you are successful. You can set your goals as a function of your possibilities, and if you have worked 300 countries on 80 and 200 countries on 160 from an average urban-lot QTH, then you are, by all standards, a very successful DXer.

To be successful in “big game” contesting, you cannot compromise with yourself. You need to have the best antennas, the best station, the best operators, nothing but the best if you want to score high in world ranking. The best multioperator contest stations are all built, improved, maintained and run by engineers. This is no coincidence.

Undoubtedly international contesting is the ultimate challenge. It can provide the truth by excellence.

Contesting truly is the Formula 1 competition in Amateur Radio. This is what attracted me to this radio-sport.

Jim Reid, KH7M, was an operator at Stanford University, W6YX, in the years when SSB techniques were worked out there by Art Collins (yes, later from Collins Radio) in the 1950s. He was a witness to how Amateur Radio contributed to important advancements in communication technology, now almost half a century ago. He made an interesting comparison between the world of contesting and the world of car racing:

"Today, about the entire globe, the most sophisticated, and elaborate HF band stations are owned by contesters such as CT1BOH, WB9Z, IR4T, PI4COM, IY4FGM, GW4BLE, JA3ZOH, PY5EG, W3LPL, K3LR, VE3EJ, and so on (ON4UN was also mentioned — thanks Jim!). Each of these stations has elaborate and multiple rig setups, multiple antenna installations, many computers with each operating position networked with logging programs, band mapping programs, propagation monitoring radios, and so on.

"These Amateur Radio stations are all Contest/DX stations. They each have the most sophisticated and up-to-date technology possible. Each has invested into it what would be comparably invested into a stable of Ferrari racing automobiles; and I have seen both, especially in Southern California!

"These guys have pushed and pushed at manufacturers, at antenna designers, at software writers, and continue to do so. The station owners themselves are all first class operators and technicians. They spend virtually all of their time "tinkering" and pushing the state of the HF art, in every way feasible. Every station I listed represents thousands of man-hours of work, every year, to maintain and remain in the top ranks of competition in the sport of DX contesting, which of course has no more purpose than the sport of auto racing: fun to have and maintain and win with the *best*.

"The owners of these stations have pushed the state of the art of Amateur Radio every bit as directly as the owners of Indianapolis racing machines have pushed the state of the art of tires, lubricants, engines, brakes, frame design, and so on."

In the highly competitive sport of international contesting, it does not suffice to have the best car or engine; you also need the best drivers, the best mechanics and the best engineers. Contesting is indeed very much like car racing.

In DXing you can get the satisfaction and fulfillment of working all countries on 40 or 80 or even 160 meters. Once it's done, the game is over. Not that the game of working all countries on Top Band will ever be over, I guess. But in contesting, there is a new competition calendar every year. Every year you can measure the station's performance, you can measure your improvements, plan your progress and enjoy the fulfillment of your victories, over and over.

That is why international contesting is the ultimate playground of ever advancing, competitive and self-fulfilling Amateur Radio.

2. WHY CONTESTING?

Now and then I read on the Internet how a proud antenna builder tells us all about the wonderful performance of his new antenna. He is trying to convince the world by telling us all about the rare DX he worked with his new antenna. What does that prove? Very little, really. Working DX is in no way a proof of technical performance of an antenna. Not in the strictly technical sense, in any case.

Working rare DX can be the proof of outstanding operating, dedication and perseverance, when it's done from a modest QTH with small antennas.

If you want to prove the technical capabilities of the station, there are really only two ways. Number one consists of elaborate full-scale field testing and measuring in a precisely controlled environment. This is beyond the reach of almost every ham.

The second possibility consists of testing your weapons, not in a shooting range or in a lab, but on the battlefields. These battlefields are the major international contests. Contests are possibly as close as we can get to a controlled environment, simply because it is extremely unconditioned. In major international contests you are competing against all the best-engineered and equipped stations, under a variety of continuously changing conditions, which really makes it a fair and equal battle and test.

This is why, after having been an "occasional" contester for almost 30 years, I decided to get into some serious contesting, thereby putting emphasis on the low frequency capabilities of my station.

3. WHAT CATEGORY?

In order to convert my station into a successful contesting station the first decision was — "we want to win, but in which category?" In other words, what is the appropriate battleground (playground) for the weapons (toys) we have?

The biggest and probably best-known stations are the *multi-multi* stations. The most successful of them have two stations per band; that means 12 fully equipped stations. These stations are on each of the six bands, 24 hours per day, and they must catch every single opening. They need to have access to a wide variety of antennas with different wave angles. Therefore, they are generally equipped with various stacked Yagis for the HF bands, even including 40 meters. The second station, whose task it is to look for multipliers, generally has access to a simpler antenna setup. It is located as far away as possible from the running station's antennas, to minimize interference, although eliminating same-band interference is quite impossible.

Interstation interference is the most challenging technical challenge in multi-transmitter station design. With multi-multi contest stations, though, each of the band stations can be completely (galvanically) separated from each other, which certainly helps prevent leakage paths for unwanted coupling between stations. In this respect a multi-multi is simpler to design and make than a so-called small multi-single, where all of the antennas have to be accessible by both stations. This makes eliminating leakage paths much more difficult.

There are a number of *multi-single* stations, which as far as station design is concerned really fall in the category of multi-multi stations. I call them *big multi-single* stations. They have six well-separated stations, one of which "runs" while the five other stations are manned and are checking each of the five other bands simultaneously. In this configuration as well, it is possible to achieve better isolation between bands because there are six completely separate stations. These big multi-single stations normally also have antennas for each band on separate towers. To build a really top-notch multi-multi station you probably require at least 2.5 to 5 acres (1 to 2 hectares or 10,000 to 20,000 m^2) of land — and that does not include what you need for Beverage antennas.

For a big multi-multi setup, in addition to the financial limitations, there is simply not enough space for putting up additional towers in my backyard. This is why we decided, almost 20 years ago, on going for the category of multi-single, or — as I call it — *small multi-single*.

Small multi-single stations can be built much smaller than multi-multi stations. A small multi-single is a station with only two operating positions, one for the "run" station, and one for the "multiplier" station. The operator of the multiplier station has to scan all bands for multipliers. In general both the run as well as the multiplier station will have access to all antennas, which means that a fairly complex antenna switching system is part of the setup. Such switching systems increase the potential for unwanted coupling between the two stations. This is what makes designing a small multi-single station technically more difficult than a large multi-single or a multi-multi station.

While it is imperative for a multi-multi station to catch every single band opening, and therefore requiring antennas to match all possible elevation angles, this is not necessary for a multi-single station. The run station will run on the bands at the times the takeoff angle of his antenna matches propagation best. In other words, the height of the antennas should be such as to accommodate the average elevation angle, the angle that produces most QSOs for the longest period of time. This means antenna heights between 18 and 30 meters for 10 though 40 meters. The multiplier station may have to call a multiplier with an antenna that is not at the ideal height. He may not get through on the first call, but this is not as important for the multiplier station.

It goes without saying that a station designed for small multi-single is also well suited for *single-operator two-radio* (SO2R) contesting. In recent years SO2R has become quite popular. In concept and in layout it is very similar to a small multi-single station, except the two operator seats are replaced by a single one.

But there is not only multi-operator or SO2R all-band HF contesting. Any station that is successful in DXing could be a candidate for *single-operator contesting*.

And if the station is not equipped for all bands, a single-band effort can be contemplated. Also, if you are not 20 years young anymore, *single-band contesting* is attractive. If you operate the low bands, you have all day to rest. You can still prove the technical excellence of your station on the band of your choice!

Having passed the standard retirement age is a good excuse to take it a little easier. I use that excuse but I still enjoy very much single band single op (CW) contests on the low bands.

4. ANTENNAS

Almost 20 years ago the ON4UN (also OT7T) contest station was designed as a multi-single and a single-operator, two-radio station. **Fig 15-3** shows the QTH and some of the antennas. One tower supports the 40-meter Yagi (at 30 meters height) and the 20 meter Yagi (at 25 meters height). As they are both on the same tower, they cannot be rotated independently. A similar combination exists on tower number 2, where a 6-element 15-meter Yagi (at 24 meters) tops a 6-element 10-meter Yagi (at 19 meters). The third tower is quarter-wave 160-meter antenna, which also serves as a support tower for the 80-meter Four Square.

About 100 meters behind the house is a fourth tower (18 meters high) with a Force-12 C31XR triband Yagi. This is what we call the *multiplier antenna*. A Four Square serves also as multiplier antenna for 40 meters. See **Fig 15-4.** These

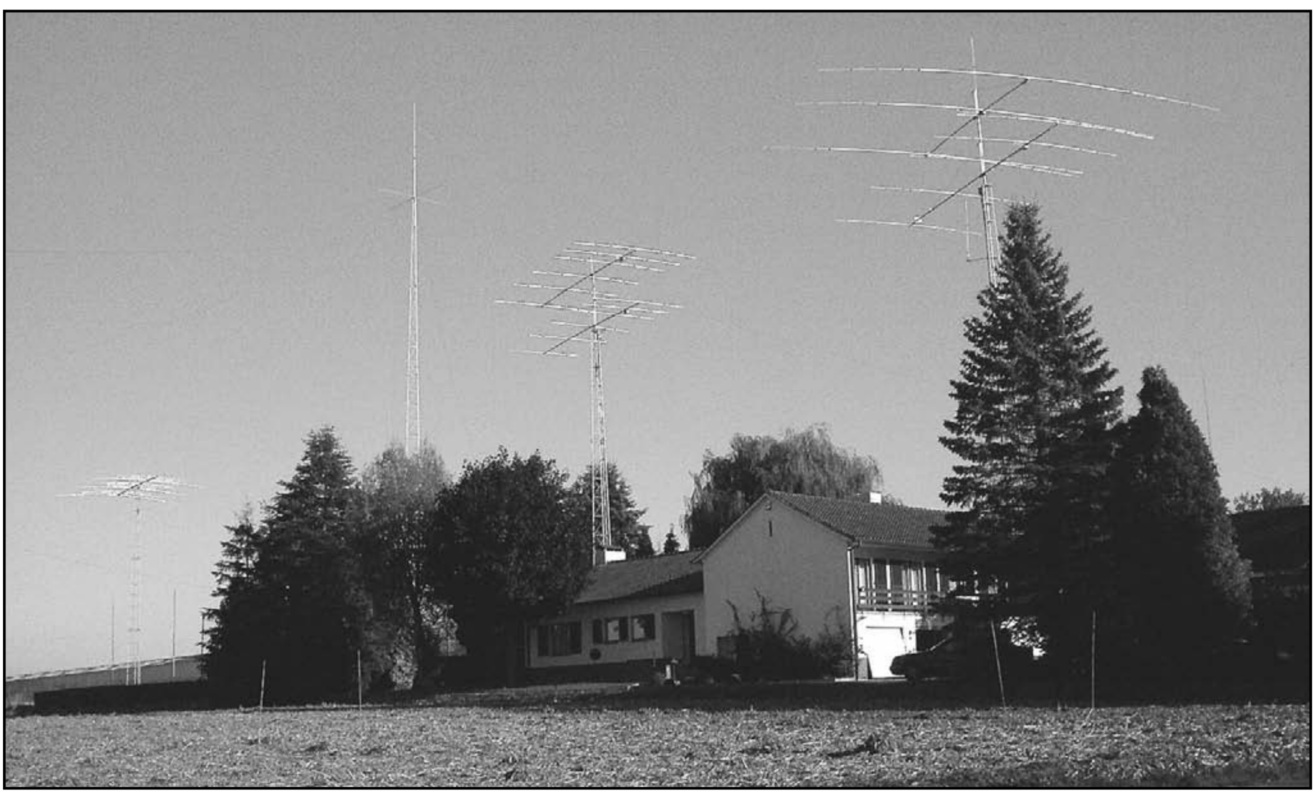


Fig 15-3 — The author's contest station is located on a 4000 square meter (2 acre) plot about 10 km from the nearby major city of Ghent. On this land are all antennas, except for Beverages, of course. In the foreground you can see part of the 30,000 square meters of farming land that is used in winter for putting up 12 Beverages, one every 30°.

Fig 15-4 — Way in the back, almost 100 meters behind the house, are the two "multiplier" antennas — a Force 12 C31XR (excellent triband antenna!) and a 40-meter four square with elevated radial system.

Fig 15-5 — SO2R setup at ON4UN using two Orion II transceivers (2005). On the extreme left of the picture we see the 12 position rotary switch for selecting the Beverage antennas.

multiplier antennas are used whenever the main antenna is not available for the multiplier station. For example, when the run station runs on 20 meters with the big Yagi, the 40-meter Yagi is not available, and the 40-meter Four Square must be used to work multipliers on that band. The 40-meter four-square is also often used to work a multiplier as switching directions goes much faster than rotating a big 40-meter Yagi.

What makes a low band station a good or excellent low band station is its special receiving antennas. Some people may say, "If I have a 3-element Yagi on 80 meters, I don't need Beverages." My answer to this statement is, "Can you change directions *instantly* with your beam?" You know the answer.

In a contest you must be able to switch directions all the time. I always used to say to a new operator who was not used to working with a range of Beverages: "I have 11 bad receiving antennas and one good one. Keep cranking the rotary switch to find out which is the good one." The aerial picture of my QTH, shown in Chapter 7 in Fig 7-102, gives you an idea of what's needed to be a station that hears (very) well on the low bands.

Since I have replaced the Orion II transceivers with Elecraft K3s (in 2008), I have largely rebuilt the station, which now no longer can accommodate either multi-single or SO2R, but I have kept all the antennas.

5. THE OPERATING POSITION

For a normal everyday DXing station there is practically no rule on how things have to work in the shack.

For a contest station equipped for a *team effort* it is different. Things have to be simple and ergonomic. A *hired-gun* contest operator is usually someone who doesn't read manuals. He or she wants to sit down and start operating right away. This means that the whole system must be simple and idiot-proof! I remember the days that we had no safety interlocks and that operators would start transmitting on the wrong antenna or with the amplifier set on the wrong band, resulting in inevitable damage and lots of frustration. Fortunately those days are over now.

It is not possible to describe what the ideal contest station should look like. There are too many variables involved. But a really well designed contest station is very different from a run-of-the-mill DXer station.

Building blocks are available from different sources. Array Solutions carries a wide variety of useful components, but many other suppliers have the individual parts. Check the *National Contest Journal* (ARRL publication), where they all advertise.

In the multi-single days at ON4UN there were two operating positions separated as widely as possible on the 7-meter-long operating desk. At one time there even was a sound absorbing mini wall set up between the two stations. In order to make SO2R possible, the two stations were later moved side by side as shown in **Fig 15-5**.

In 2008, with the introduction of two K3s, the station layout was changed again. The second station, which, in the SO2R days was situated immediately next to the "run" station, has been converted into an "easily movable station" (**Fig 15-6**). The K3 sits on top of a 5-cm high control unit (raising the K3 tuning knobs to a more comfortable operating height) as shown

Fig 15-6 — On the left is the "movable" 2nd station at ON4UN, set up in the corner of the shack where ON4UN does his book writing and other PC related duties (2009 picture). Each of the K3s has its own PC and two screens. The screen on top of the K3 only displays CW skimmer (VE3NEA) and its call sign list. The larger screen to the right displays several N1MM contest software windows as well as a *Power-SDR* (IF) spectral display. Both displays are generated using LP-PAN with the Creative 0202 USB sound card, both sitting on top of the PC. The two screens on the right are connected to the "general applications" PC, on which the 5th edition of this book was created.

in **Fig 15-7**. The control box contains antenna selection (main / second / both), direction switching for the 80-meter Four Square (plus band section selection: CW, mid band, SSB). It also provides rotator control for the different towers plus digital direction read-out.

The control unit is connected with a 25 conductor umbilical cord to the main station. This makes it possible to set up the second station anywhere in the house, even in the bedroom if necessary. This station normally drives an ACOM 2000A amplifier, which has its remote control unit set up next to the K3 control box.

Fig 15-7 — The K3 at the "movable" operating position, on top of the homemade control box, which includes a Microham interface/keyer and the switching circuitry that makes it possible to do all the remote antenna switching and more.

5.1. Antenna Switching at ON4UN

Since two different antennas are available on the higher bands (40 through 10 meters), I have made provision for either station to use either one of these antennas or split the power 50/50 into those two antennas. When the band is open in two directions, you can thus work in two directions simultaneously. Of course, you must realize that you have more QRM/noise and only half the transmit power in each direction. If you are using one antenna, you can quickly switch directions to work a multiplier. It is imperative that the SWR curves of both antennas are flat and similar, so that you do not have to retune the amp while switching.

Fig 15-8 shows the switching/combining unit I built. The unit contains four L-networks (10 through 40 meters) that convert 25 Ω back to 50 Ω. Simple ac-type relays are used for the switching, which results in quite a bit of wiring inductance (about 8 inches of wire). To compensate for the effects of these long wires, I put capacitors at both ends of each relay wire, creating pi networks. With about 30 pF on both ends, the SWR is less than 1.05:1 even on 10 meters.

The antennas are selected fully automatically, using the band output data from the transceivers. I built a control device using the band data output from the two transceivers to generate the logic signals for selecting the antennas. **Fig 15-9** shows the control unit. The switching logic must prevent the two stations from selecting the secondary (triband) antenna on 10/15/20 while the triband antenna is already in use by the other station. The logic has been developed such that when one station uses the tribander, the other one cannot get it, and will remain on the main antenna.

Relay logic is used throughout so the system is totally immune to RF and very reliable. A total of 11 small relays are used for logic switching. A three-position toggle switch is mounted on the control boxes of both K3s.

The three positions are:

1) Primary antenna
2) Secondary (multiplier) antenna
3) Both antennas in a 50/50 power split.

Fig 15-8 — On the left are the selector/combiner unit for 10 through 40 meters. The feed lines (⅞-inch) from the single-band antennas and the 10-15-20-meter tribander (C31XR) arrive on the left. From top to bottom: 40, 20, 10 and 15-meter L-networks. There are 15 relays in the box. On the right are four band outputs for 10 through 40 meters, which go to a home-built switch box with both N connectors and 7⁄16 inch connectors.

Fig 15-9 — The homemade antenna switching logic box. The inputs are the band data outputs from two transceivers and the selection data from the "main/multiplier/both" switches mounted on the K3 control boxes. The outputs are: switching data for bandpass filters and control voltages for switching the antenna relay box and for the combiner unit shown in Fig 15-8.

Fig 15-10 shows the block diagram of the system. On 80 and 160 meters, where only one transmit antenna is available, these antennas are fed directly from the six-pack switch. And if for any reason the band data from the switch and the data from the transceiver do not match, the transmitter in the transceiver is inhibited (in the K3 via the TX-inhibit line).

Note that the "movable" station (shown on the right in Fig 15-10) is connected to the main station hardware with only three cables: a 25 conductor cable, a length of RG-58 and the cable between the ACOM 2000A and its remote control unit.

5.2. Antenna Directions

On the receiving side the visual direction indication of the Beverage antenna selector proved to be very helpful. In Fig 15-5 and Fig 15-7 you can see the Beverage selector boxes, which use a 12-position rotary switch with LEDs for each azimuth direction.

Point-and-forget rotators: In the heat of the battle you don't want to sit and press that turn-left or turn-right button — just select your direction and press one button.

5.3. Radio-Computer Interface

Computers and contesting software are covered in detail in Section 7 of this chapter. Top-notch contesting without top-notch contesting software is no longer possible. All contesters control a great number of functions of their radios through the

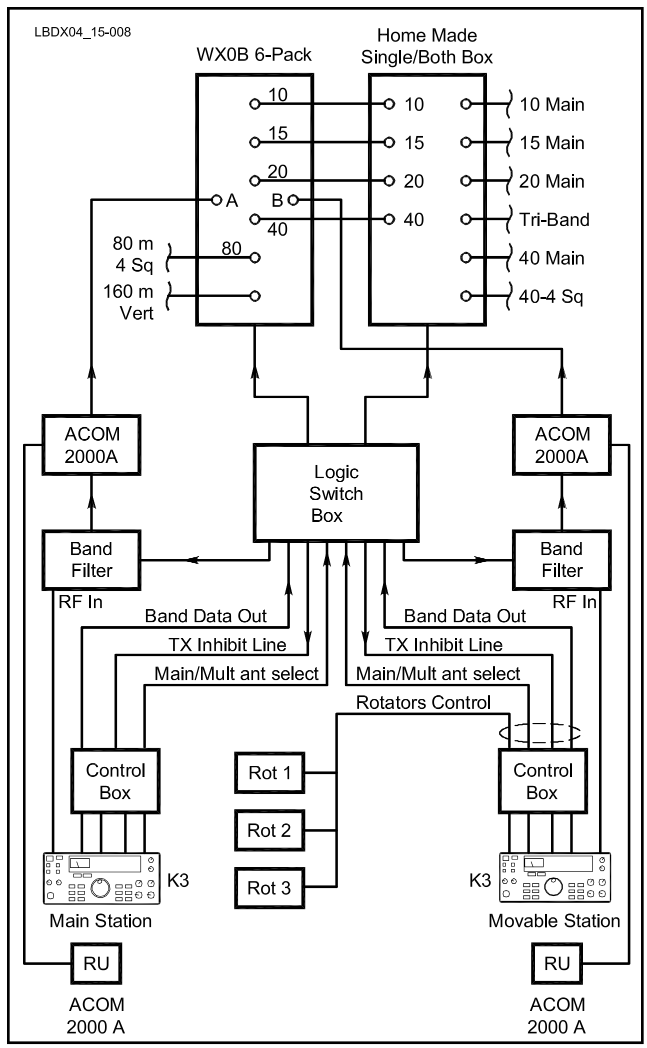


Fig 15-10 — Block diagram of the station and antenna control circuitry at ON4UN.

computer, or through a computer-controlled dedicated piece of equipment — for sending CW, changing frequencies, bands or modes and so on. You name it.

Until about five years ago connections between these boxes and the computers were all done by serial or even parallel ports. Modern computers no longer have these ports and communication with the dedicated boxes is now all done via USB ports. If the equipment still uses a serial port we need to use a converter for changing USB signals to serial-port signals.

5.4. Easy-to-Tune Power Amplifiers

In a typical multi-single or SO2R contest setup one transmitter (the run station) is tuned to the main band, where a pileup is worked. With the second transceiver the other bands are scanned, and multipliers are picked up between QSOs on the running band. In order to be able to concentrate fully on the operating aspects, band-switching should be automated as much as possible.

Tuning the second linear between bands often has been a problem, because:

- Contest operators often don't know how to properly tune a linear.
- It takes them too long.

The minimum to have is labels stuck to the linear front panel with all settings for all antennas and modes. Even that sometimes seems to be too difficult for the operators trying to concentrate on moving this very rare and weak KH8 station from one band to another. An amplifier that automatically switches bands, and automatically tunes to preset values of band switch, load C and tune C, or better yet, performs a fully automatic tune-up, is the answer to that problem.

Since 1998 the ON4UN station has been equipped with an ACOM 2000A linear. Five years later a second ACOM 2000A was added. The use of this fully auto-tune linear proved to be very helpful for working multipliers by quickly changing bands and antennas.

The ACOM 2000A is an auto-tune nominal 1500-W output amplifier (maximum 2000 W) using two Russian-made 4CX800A (or GU74B) tetrodes. This was the first real auto-tune amateur HF-amplifier I had ever seen. By pressing a button on the remote control panel it automatically tunes itself completely within a half second. The auto-tune function is not limited to recalling preset values — it actually tunes for a match for a load within the 2:1 SWR circle (on some bands up to 3:1)

The amplifier has an absolutely blank front panel, except for an ac on-off switch. This makes it possible to hide the amplifier in any convenient place. At my place the two amplifiers are in a little room adjacent to the shack (**Fig 15-11**). Doing this I have eliminated all noise and heat that is inevitably generated by a power amplifier. All control and monitoring functions are grouped on a remote small control box, which can easily be positioned next to the computer keyboard during operation. The ACOM amplifier can be connected via an RS-232 connector to a PC for either remote control or testing. Its processor keeps track of all the important data (currents, voltages, temperatures).

In case of a fault, you can send the information stored for the 12 most recent faults to the dealer or the factory by means of Baudot code on the telephone — simply put the microphone close to the tiny loudspeaker on the control box rear panel, or by means of a personal computer and its inherent communications channels (Internet, modem, etc). Needless to say, the use of this amplifier has greatly increased flexibility and efficiency at ON4UN's contest station. Since introducing the 2000A, ACOM has built a superb quality and service record, and continues to be an excellent choice for serious contesting.

Another popular legal-limit, automatic tuning amplifier is the Alpha 87A amplifier. This amplifier has very similar, but not identical, characteristics (also 1500-W nominal output), and is certainly a valid candidate for a top-notch contesting station linear as well. The Alpha 87A uses two 3CX800A7 triodes, which have the disadvantage of being more expensive than the tetrodes used in the ACOM. The Alpha 87A also has an RS-232 interface port, which allows the linear to be controlled remotely. In addition, key parameter measurement values can be monitored remotely. The newest auto tune Alpha amplifier is the model 9500 which uses an 8877 triode.

Several solid-state no-tune amplifiers are available as well. Models from Yaesu, ICOM, Tokyo Hy-Power, SPE and Ameritron are rated for 900 to 1500 W depending on model.

6. THE STATION AS SEEN BY THE TECHNICAL PERSON

It is clear that the technical requirements for a top-performing contest station are far superior to what's needed for casual, or even serious, DXing. Think of harmonic suppression. Stations built for multi-transmitter operation must transmit the cleanest signals, and their harmonics must be suppressed far in excess of the standard.

Another issue is to keep the contest station "up and running" all year long. It takes good mechanical engineering to keep the antennas up. A little side remark here: All the home designed and homemade antennas that you can see in Fig 15-3 now have been up nearly 20 years, and they still look and perform like new. Needless to say, over the years they have gone through quite a number of storms with over 100 km/h winds. Well, if you hate to climb towers, like I do, the only way to avoid it is to build your antennas strong enough!

Fig 15-11 — At ON4UN the two ACOM 2000A amplifiers are located in a small room next to the shack, in order to keep all the heat (cooling heat) and noise out of the shack. You really don't have to do it for the noise as these units are actually quieter than many PCs.

6.1. The Operating Desk

Even though in the early 1990s we were housed in a small shack of 3 by 3.5 meters, we nevertheless managed two multi-single first places in Europe. But we were almost sitting on each other's laps! So, one wall was taken out, which made it possible to install a single 7 meter long by 1 meter wide operating table.

The new shack layout was conceived with contesting in mind. To provide the best possible RF and safety ground, the underside of the table was entirely covered with a 1-mm thick aluminum sheet. This sheet provides maximum capacitance to the equipment standing on the table, and minimum resistance and especially inductance for good RF grounding. Forty ac outlets are mounted on the aluminum sheet, providing the shortest possible safety ground return for the outlets. Short and wide straps are connected to the sheet and are available on the back side of the table to ground various equipment.

The table is separated from the wall by approx 15 cm, which allows wires to pass and for ventilation as well. The aluminum ground sheet is grounded with a short strap to an excellent RF ground just outside the shack, with a 40-cm heavy-gauge cable going right through the wall.

The table is equipped with three separate mains distribution circuits, each equipped with a professional-grade mains filter.

6.2. The Monitor Scope

In the heat of a phone battle, operators sometimes have the tendency to crank up the microphone gain, resulting in poor and distorted audio, unnecessary splatter and so on. I always have a monitor scope connected to the output of each of the stations. I use a second-hand commercial 20-MHz 'scope, and tap off a little RF using a resistive voltage divider across the output of the linear. This way the operator always has the pattern of the transmitted signal right in view (see Fig 15-7).

6.3. The Problem of Interband Interference

In a two or more station setup, interband interference is the number one technical problem. But as the saying goes, every problem is an opportunity. In this field lies the opportunity to excel. Here also lies the opportunity for equipment manufacturers to improve their equipment. Interference can be minimized by using the following techniques:

- Use a transceiver with lowest possible VCO noise (see Chapter 3, Section 1.3.5) and keep pushing the equipment manufacturers to produce transmitters with the lowest possible in-band noise output.
- Separate the antennas as much as possible.
- Use vertical and horizontal polarization to take advantage of the additional attenuation of unlike polarization.
- Use band-pass filters between the exciter and the amplifier.
- Use amplifiers with pi-L networks, not simple pi networks.
- Suppress common-mode currents on the feed lines.
- Galvanically separate the feed lines of the separate bands.
- Use band-reject filters between the amplifier and the antenna.

It is obvious that interference will be heard on the harmonic frequencies. This poses much more of a problem on CW than on phone. The CW band segments are all in the low end of the bands, and the harmonics of 3.503 will be 7.006, 14.012, 21.018 and 28.024 MHz — all right in the CW window. On phone, if you operate on 3.775 kHz, the harmonics will be on 7.550, 15.100, and so on, all outside the band. There is no real problem with the direct harmonic frequencies when operating phone.

Unfortunately most present-day transmitters do not transmit just the wanted signal; they also transmit a lot of noise around the transmit frequency. This noise can often make it difficult to copy, even many kHz away from the exact harmonic of the transmit frequency, unless effective filtering is applied. And even then, the final improvement will have to come from the designers and manufacturers of our transceivers, putting out equipment producing less in-band noise.

My friend George, W2VJN, covers all of these aspects very thoroughly in his excellent publication *Managing Interstation Interference,* of which a second and thoroughly revised edition can be obtained directly from him. If you are serious about tackling inter station interference, first read this excellent book.

6.3.1. Medium Power Band-Pass Filters

There are a few commercial sources for medium-power band-pass filters that are widely used in multi-station contest setups as well as during DXpeditions. I have experience with the ICE, Dunestar and W3NQN units. The ICE units are rated 200 W, and if the SWR is low they will indeed cope with 200 W. The Dunestar filters are rated at 100 W. I have been using all three of them and I did some comprehensive measuring on these units.

The ICE and Dunestar units have insertion losses of between 0.3 and 1.0 dB (that is a lot) depending on band. Due to the circuitry used, the Dunestar filters have significantly steeper shape factors. The W3NQN filters undoubtedly have the best characteristics and show 0.2 to 0.4 dB insertion loss, depending on band. If we look at the performance of the 40-meter filters, we see that the ICE filter will attenuate 20 meters about 32 dB, the Dunestar more than 50 dB and the W3NQN between 80 and 90 dB. **Fig 15-12** shows the response of a 7-MHz W3NQN filter. **Fig 15-13** shows six of the W3NQN filters connected to a switching system.

Six such filters are used between the K3 and ACOM amplifiers at ON4UN. The logic switch box (Fig 15-10) provides control voltages to select the appropriate filter.

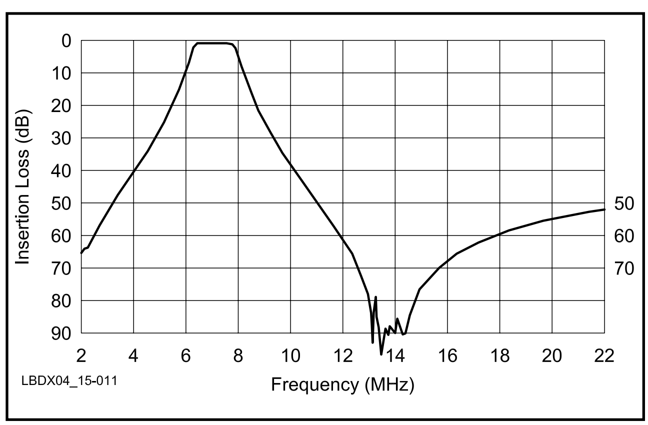
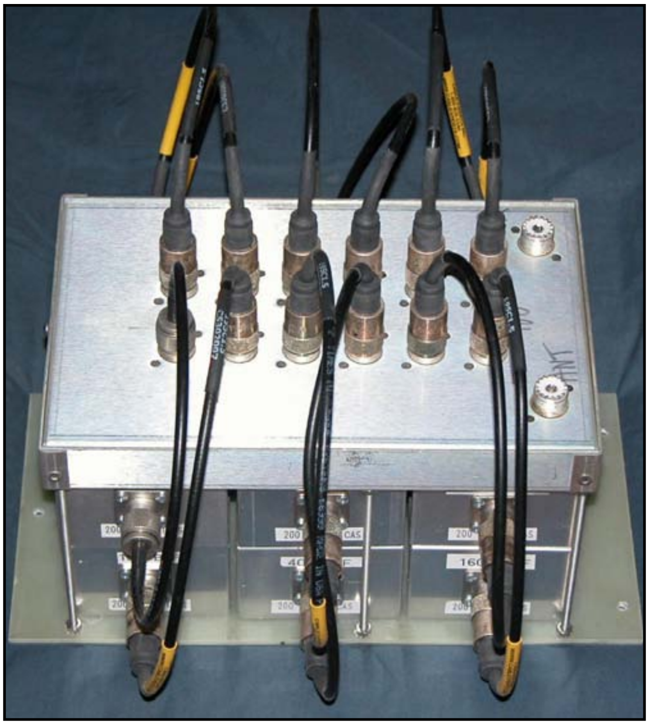
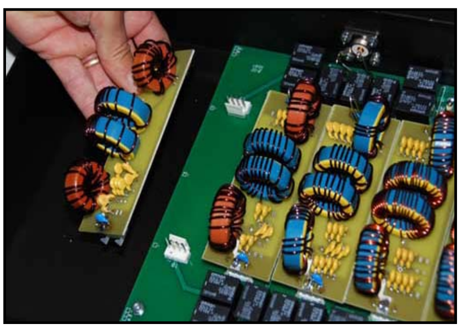


Fig 15-12 — ON4UN uses the W3NQN filters switched with the WXØB FM-6 filter box between the transceivers and the amplifiers. This is the response of the 40-meter W3NQN filter. Note the deep null at the second harmonic at 14 MHz.



(A)



(B)

Fig 15-13 — A: Six W3NQN filters connected to a relay matrix switching system, model FM-6 (source: Array Solutions). B: This is now also available mounted in a single metal box where the individual filters, mounted on a piece of PC board, can just be plugged in.

Table 15-1 lists some of the major characteristics I measured for 160, 80 and 40 meters.

It is important that the filters be operated at a low SWR. If not, you will likely blow the capacitors. It is important, when driving a linear amplifier through one of these filters, that the linear is switched to the right band. If not, a high input SWR may result. If the exciter is equipped with a built-in tuner, it may try to get the full power into the filter, at a very bad mismatch, which guarantees fried components. Therefore, you shouldn't switch the automatic tuner on when operating with band-pass filters. Also, it is a good idea to control the selection of the appropriate filter right from the transceiver's band data output, so you do not dump RF of the wrong band into the filter. There must be many contesters and DXpeditioners who have done that. I am sure replacement capacitors for these units must be a hot selling item!

6.3.2. High Power Filters

If you run power, it really is a must to run filters beyond the amplifier as well, because the amplifier also generates harmonic power. These filters should not only be designed to suppress harmonics; in addition they should attenuate signals on all bands, also on frequencies below the transmit frequency. It is not uncommon for signals from one of the stations of a multioperator station to mix in the linear with other signals (broadcast band signals or those from another amateur band) and create unwanted mixing products. The ultimate filter is indeed a filter that attenuates all other bands but gives the highest attenuation to the second harmonic.

The most common way of achieving out-of-band attenuation is by using band-reject filters. These can be made with discrete components or by using coaxial cable.

6.3.2.1. Using Discrete Components

High-power filters using discrete components can be made much smaller than those using coaxial cable, but the components are hard to come by (high-power, high-voltage, high-current capacitors) and the design requires some expertise and the use of a quality network analyzer.

I have designed a series of such filters, which perform very well. **Fig 15-14** shows a 10-meter band-reject filter that will take 3-kW continuous-duty power. I built it in a box measuring 25 × 6 × 6 cm. The box is made of double-sided glass-epoxy

Table 15-1
Bandpass Filter Measurements by ON4UN

		Attenuation					
Filter	*Band*	*1.8 MHz*	*3.5 MHz*	*7 MHz*	*14 MHz*	*21 MHz*	*28 MHz*
ICE	160 m	0.4 dB	15 dB	27 dB	40 dB	>45 dB	>45 dB
Dunestar	160 m	1.0 dB	28 dB	>45 dB	>45 dB	>45 dB	>45 dB
W3NQN	160 m	0.2 dB	50 dB	>80 dB	65 dB	50 dB	58 dB
ICE	80 m	25 dB	0.34 dB	17 dB	30 dB	>40 dB	>45 d8
Dunestar	80 m	42 dB	0.64 dB	37 dB	>45 dB	>45 dB	>45 dB
W3NQN	80 m	53 dB	0.4 dB	70 dB	62 dB	55 dB	49 dB
ICE	40 m	>45 dB	38 dB	0.8 dB	32 dB	>45 dB	32 d8
Dunestar	40 m	>45 dB	43 dB	0.6 dB	>50 dB	>45 dB	33 dB
W3NQN	40 m	68 dB	45 dB	0.3 dB	>80 dB	52 dB	48 dB

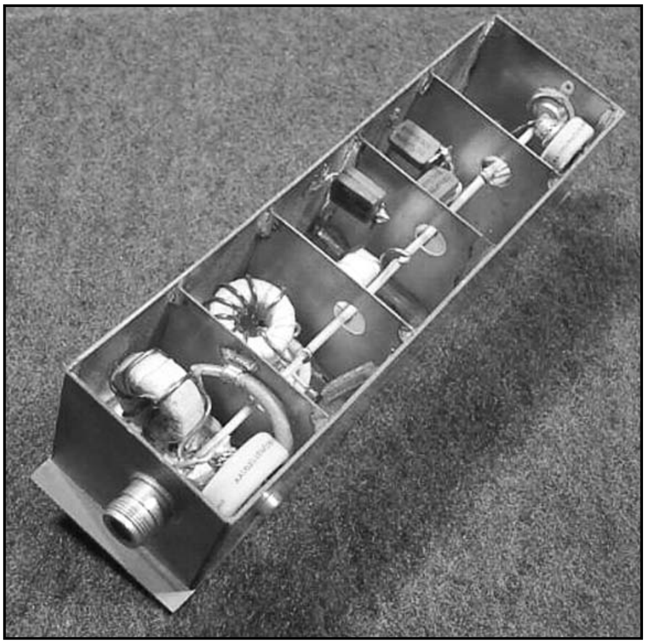


Fig 15-14 — Ten-meter high-power band-reject filter. This simple unit uses one series-tuned trap for each of the five bands to be rejected.

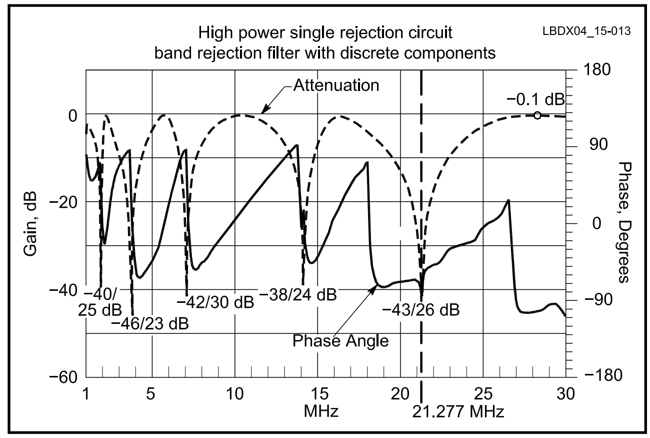


Fig 15-15 — Rejection curve for the homemade high power 10-meter filter in Fig 15-14. The rejection figures quoted are for the band center and band edges. For example, the rejection in the center of the 15-meter band is 43 dB, and 26 dB on the band edges. This measured plot was generated using the Alpha/Power network analyzer.

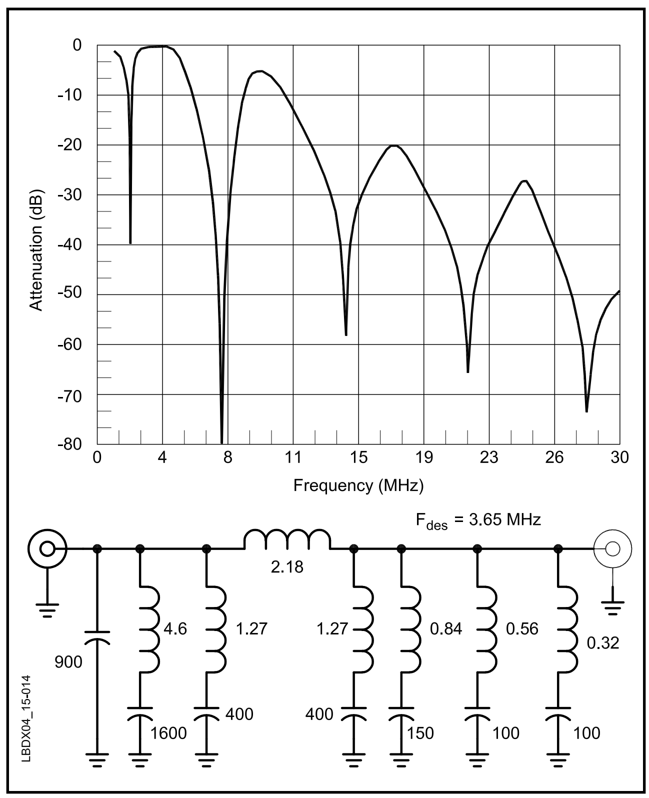


Fig 15-16 — High performance 80-meter filter, using discrete components and two 40-meter traps. This filter has a rejection of 80 dB on 40 meters, and between 60 and 70 dB on 20, 15 and 10 meters. The insertion loss is less than 0.1 dB.

printed board material, which is ideal for the application. With single-pole series-tuned circuits for each band, an attenuation of 38 to 46 dB was obtained on all five bands, with an insertion loss of about 0.1 dB. **Fig 15-15** shows the response from 1 to 30 MHz for this filter.

The principle for designing such band-reject filters is really quite simple. You design five series-tuned circuits, each one tuned to the frequency you want to suppress and simply connect all these traps in parallel. For the 10-meter filters, all these tuned circuits will exhibit an inductive reactance on 10 meters. You can easily calculate this value. Calculate the impedance of all the coils and all the capacitors (five of each) used in this filter. Since they are connected in series (for each band), you can simply add the values, taking the sign (+ for a coil, – for a capacitor) into account. Then calculate the parallel value of all of these, just as you calculate parallel resistors. Now we can "tune" out this positive reactance by using a parallel capacitor, which resonates the whole thing on 28 MHz. It really is that simple.

For other bands, series-tuned circuits below the design frequency will show as inductors on the design frequency, and as capacitors above the design frequency. By judiciously choosing the LC ratio of the series-tuned traps, you can now design filters where the positive reactance of a group of traps will cancel the negative reactance of another group, which means there will be no need for a parallel capacitor or inductor to tune the filter to a 1:1 SWR on the operating frequency.

Fig 15-16 shows a high-performance 80-meter filter using a pair of 40-meter traps for improved rejection. The basic configuration is a low-pass section. The effect of the low-pass section can clearly be seen at the overall shape of the rejection curve. Filters like this can easily be modeled using computer-aided design software. In this case I designed a symmetrical low-pass filter, and arranged the traps on both sides of the inductor to obtain the same capacitance value. A capacitor (900 pF) had to be added on one side to tune the low-pass filter. The value of the inductor can easily be calculated using the "Line Stretcher" module of the *New Low Band Software*.

If you want to design your own filters, the sky is really the limit. The biggest problem in making such filters is to obtain

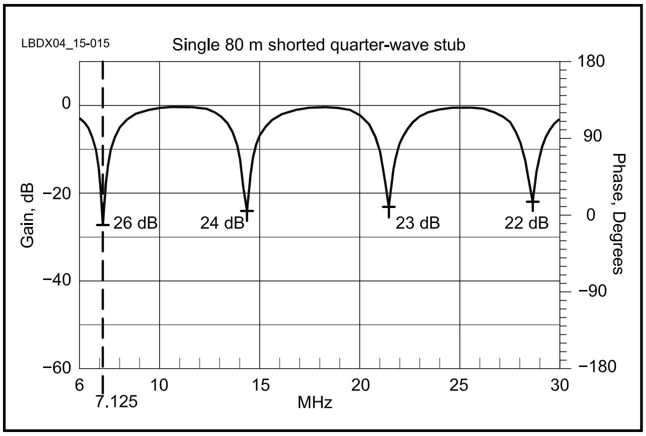


Fig 15-17 — Attenuation "dips" are obtained on all of the harmonically related frequencies.

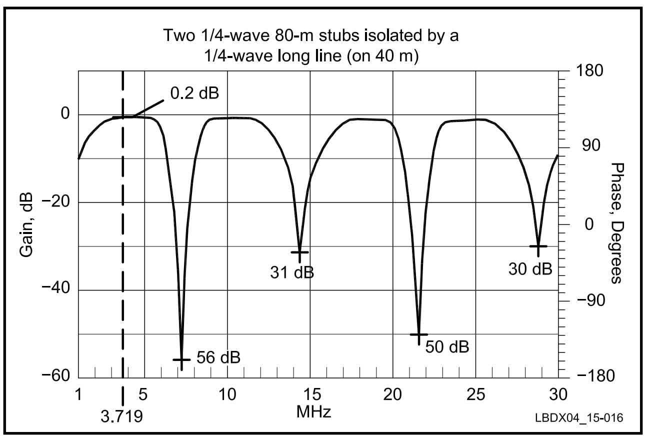


Fig 15-18 — Attenuation curve for two stubs separated by a ¼ λ feed line (see text).

suitable capacitors. Inductors can be wound on powdered-iron toroidal cores (#2 material). Make sure you calculate the estimated power that will be dissipated in the cores. On adjacent bands there may be a substantial amount of heating in the cores, and 2-inch cores may be required in some circumstances.

It is beyond the scope of this book to deal with the concept, design and construction of such filters, but they are necessary to make a multi-transmitter station fully competitive.

6.3.2.2. Using Coaxial-Cable Stubs

Let's work out a situation where we want to operate an 80-meter station and a 40-meter station simultaneously in the CW contest. A single quarter-wave long shorted stub, made of RG-213, cut for 80 meters, will provide about 26 dB attenuation on 7 MHz, 24 dB on 14 MHz, 23 dB on 21 MHz and 22 dB on 28 MHz (see **Fig 15-17**). The insertion loss on 80 meters will be less than 0.1 dB.

A quarter-wave RG-213 stub cut for 20 meters typically shows an attenuation of 37 dB. The same stub with RG-58 shows about 25 dB of attenuation. A 10-meter stub made of RG-213 can achieve 40 dB of attenuation.

We can use two identical stubs to almost double the attenuation, but *not* by merely connecting them in parallel! Connecting a short across a short can, in the best case, when the two shorts are equally "good" or "bad," bring you 6 dB additional attenuation. There is one way, however, to obtain much more attenuation.

Look from the linear amplifier into the feed line. With a well designed and built amplifier, the pi-L filter will provide a good deal of attenuation on 40 meters. But there is some 40-meter power at the linear output. Assume it is 50 dB down from the 80-meter fundamental. At the output of the linear we assume a low Z for the second harmonic, an acceptable assumption at the output terminal of the pi-L filter. If we now put the stub (which is a short on 40) right at the output of the transmitter, we are putting a short across a short, which is not very effective. If we insert a quarter-wave coaxial line between the output of the amplifier and the stub, we have transformed the very low impedance point (on 40 meters) to a high-impedance point. If we now connect the stub at that point, we will have the most effect of the short that the stub represents. All of this holds true only if the output of the amplifier represents a low Z for the second harmonic (40 meters).

In practice it is a good idea to experiment: install the stub right at the output of the amplifier and check the attenuation on 40. Then insert the quarter-wave line between the linear and the stub. If the attenuation is better (which is likely), leave it there. In theory a quarter-wave gives best results (maximum transformation ratio), but anything from ⅛ to ⅜ λ should be OK, as long as you stay away from the region of ⅜ to ⅝ λ. Fig 15-17 shows the attenuation of a single 80-meter shorted stub (between 6 and 30 MHz).

But you wanted more than 25 dB. You can install another quarter-wave *isolation line* between the first and the second stub. I call it an isolation line because it effectively isolates the two stubs. The reasoning is the same as explained above. Two stubs isolated by a quarter-wave coax (on 40 meters) now exhibit 56 dB of attenuation on 7 MHz, and 50 dB on 21 MHz, but only 31 dB on 20 meters and 30 dB on 10 meters. This is logical, since the isolation line must be an odd number of quarter-waves long on the reject frequency to do its job. This is true on 7 and on 21 MHz only. On 20 and 10 meters we only get the predicted 6-dB improvement. In this case using an isolation line of ⅛ λ on 40 meters would result in good attenuation on 20 (where the isolation line would be ¼ λ, on 15 (⅜ λ), but not on 10 meters where the isolation line would be ½ λ. See **Fig 15-18**.

Stubs can also be used as elements in a low-pass configuration, in combination with discrete components. The example in **Fig 15-19** is a combination of a simple 160-meter low-pass filter with four stubs. The attenuation pattern is amazingly clean, and gives better than 70 dB on all bands.

The impedance of the coaxial cable used for making stubs is irrelevant. The cable loss is important, however. An 80-meter stub made with RG-58 will yield approximately 15 to 17 dB attenuation, while RG-213 gives 25 dB and ⅞-inch Hardline will give 40 dB!

In a contest station setup, you can also install fixed s tubs at the feed points of single-band antennas. At my QTH I have a 160-meter quarter-wave transmit antenna that stands right in the middle of the 80-meter Four Square. You can hardly imagine how to obtain more coupling between these antennas. Without some kind of stub an ACOM amplifier feeding one of these antennas would switch off when some-

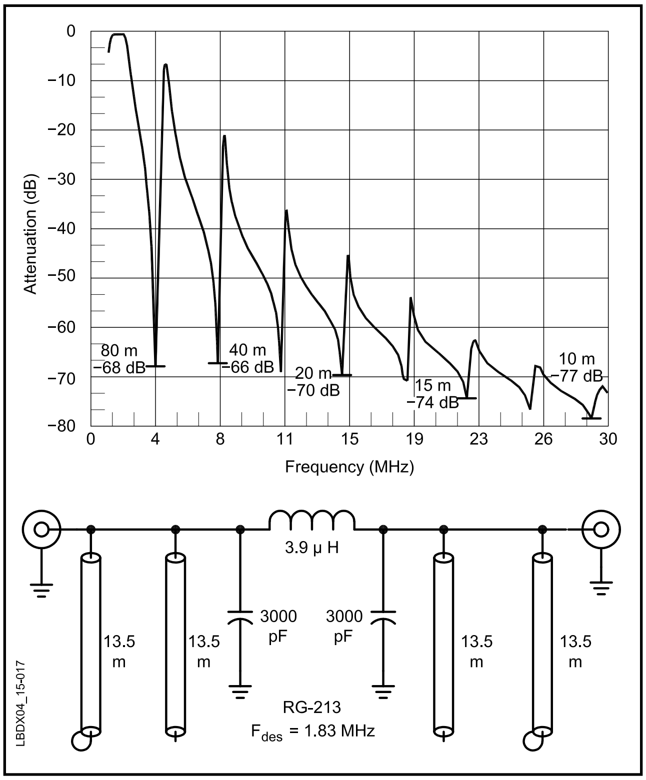


Fig 15-19 — Attenuation curves for stubs used in conjunction with discrete components to form a low-pass filter. See text for details.

one would transmit on the other antenna because of the large amount of "alien" RF it saw. I put a shorted ¼ λ 160-meter stub at the base of the 160-meter antenna and an open-circuit version at the feed point of the 80-meter antenna.

7. COMPUTERS AND SOFTWARE

While I was a very enthusiastic user of *CT* contest software way back in DOS days, I quickly became one of the very early users of *Writelog* when 32-bit *Windows* was first introduced. About seven years ago I switched to N1MM's contesting software (freeware). I use one computer for each of the two radios at ON4UN. Performance and speed go hand-in-hand, so they are state-of-the-art computers, running the latest operating software.

In addition both computers are networked by LAN with the main computer. All of this has the advantage of having instantaneous backups on the other machines, just in case something goes wrong. If one of them is a laptop, you are even protected against sudden power mains drop out. This computer is connected to the Internet via a broadband router, becoming the "main" computer.

In a multi-single contest environment this PC is what we call the "Master of Ceremony" computer. We used to have a full-time "Master of Ceremony Operator" whose task it was to make sure that we all get the Packet Cluster information available. His job was also to interpret that information and to decide when and where to run, and when and where to work multipliers. He was the man who was really in charge. In the days we ran multi-single this technique was used quite satisfactorily for years.

7.1. Connecting Computers

For our first multi-single as OT3T in 1993, we had linked a variety of PCs ranging from 286s to a 386 66-MHz machine (a fast machine in those days!) with copper-wire serial cables. Despite pounds of ferrite rods and toroids, we certainly did not achieve a totally RF-free situation. Often a computer would hang, without apparent reason. Several times during a contest, logs had to be merged and computers started up again. All of this was certainly far from ideal!

In the second phase we replaced all the copper links with fiber-optic links. This certainly was an expensive improvement, but still not 100% bulletproof. Then we went to an Ethernet solution using software written by David Robbins, K1TTT, and Wayne Wright, W5XD.

Later we found that twisted pair Ethernet cabling worked fine, even in a strong RF environment. I assume the sensitivity of PCs to strong RF fields has greatly improved over the years, and it must be over 10 years ago that we had problems with RF ingress into our PCs.

7.2. Connecting to the DX-Cluster

Here too evolution has been staggering. What was advanced five years ago, looks like museum technology today. Having had access to wideband Internet for several years now, I have quit using VHF or UHF radios to connect to the DX-cluster system. In the past I used ON5OO's excellent *DXConcentrator* software through which you could connect to a great number of DX-clusters simultaneously. That provided more and faster information than if you connected to only one DX-cluster. Nowadays the great majority of DX clusters are linked over the Internet, which means it does not pay off to check in with different clusters.

7.3. Computer Noise

In all the years we have been using various computers, I have never had a problem with direct radiation from a computer. Of course, if you use an antenna inside the shack very close to the computer, you will pick up all kinds of hash. It is very important that your antennas are at a sufficient distance from the computers, that the feed lines are well-shielded and well-grounded and that the coax connector makes perfect contact with the receptacle.

Computer screens can be very noisy though. If you buy a new display, make sure you can return it if it radiates too much. You may also want to add extra ferrite cores on the cable between the PC and the monitor. Better still, use an LCD monitor, but I've heard of cases where the monitor's switching power supply needed few more ferrites on the 12-V power cable to silence it. LCD flat-screen monitors are much less fatiguing for the eyes, and this can be important during long contest periods.

8. RESULTS

Remember why we participate in contests. There are the Formula 1 operators who want to win going full bore. But there is also the technical guy, who wants to see the fruit of his labor, his engine, his station, his antennas win in an international competition. This is what it was all about for me.

But at the same time I met a lot of good regular operators and technicians (the always available helpers from the local radio club). This is undoubtedly the important social aspect of contesting.

We also should not forget the majority of contesters, those who just take part in a contest because they enjoy contesting, those for whom winning is not so important, the guys with the little stations. These are the stations the big guns need to work if they want to win the contest.

Since the early 1990s ON4UN's station has been tested in well over 100 international contests (mainly those organized by either the ARRL or CQ), which resulted in nearly 70% first places (Europe or worldwide), plus 20% second places. This proves that the antennas as well as the station are capable of winning top-notch contests.

The low band performance of the station was further confirmed by the 80-meter country and zone totals during CQ World Wide CW contests during the same period (1993-1997). During the 1994 and the 1996 CQ World Wide CW contests we scored 5-band DXCC in one weekend, 10 through 80 in 1994, and 15 through *160 meters* in 1996. The 100-plus countries during a single weekend on 160 meters was a first, I believe.

Of all entries in the CQ World Wide 160-meter CW contests over the past 20 years, 90% were first places (either Europe or worldwide). Phone participation (with hired guns — I don't like phone contests on Top Band) yielded 70% first place Europe or worldwide. Note also that during most of these contests we scored the highest number of country multipliers worldwide. Some of these are still European records. For example, my friend George, K2UO, placed 2nd worldwide and set a new European record in the assisted category on 160 meters in the 2008 CQ World Wide phone contest (**Fig 15-20**).

The results on Top Band and 80 meters not only speak for the performance of the transmit antennas (four-square on 80 and single quarter-wave vertical on 160) but also, and even more importantly, of the receiving capabilities of the station, which we have continued improving over the years. Chapter 7 of this book gives you all the details behind those improvements over the past years.

It is obvious that taking part in the big contests is very often helpful to increase one's DXCC status as long as your county total is not nearly at the top. Up until a DXCC country score of well over 200, I worked a new country in almost every big contest on Top Band. All big DXCC scores on Top Band are from stations which, over the years have been active in the big contests.

This brings me to the topic of ethics and related issues. Achieving good results in contesting, as well as in DXing, if done according to the obvious *ethical standards*, can bring a high degree of satisfaction and even pride, and rightfully so. Pretending that you are a good DXer or a good contester through cheating is about the lowest you can fall in our wonderful hobby. But, after having been around in this hobby for almost half a century, I must admit that apparently not all individuals behave according to the same ethical standards. It makes me sad. The only thing you can achieve through cheating is to finally become an outcast, someone who's not worth being part of our wonderful hobby. There simply is no excuse for it.

To prevent such practices, I support the idea of making all contest logs and all the credit details for DXCC and other high ranking awards openly available on the Internet. If we play the game correctly, what do we have to hide? To the contrary, if we are very good, we have a good reason to show how we did it.

Fig 15-20 — George, K2UO, operating ON4UN's station on 160 meters during the CQ World Wide phone contest in October 2008, using the new K3. He worked 90 countries in 16 zones with over 1300 QSOs, good for second place worldwide and a new European record in the assisted category. Note how George is cranking the Beverage 12-position rotary switch. One week after his K3 experience in Belgium, George ordered his own K3.

9. FURTHER IMPROVEMENTS

To stay competitive in international contesting you must improve the station year after year. Our competitors do the same. This is the driving force that leads to technological and conceptual improvements. Really, it is almost the opposite from DXing. The more successful you are in DXing, the more countries you have worked, the less there are left for you to work, the less pressure there is — the more you can relax. The more successful you are in DXing, the easier your call will be recognized in the pileups (that helps, too). No need to add another couple of dB for those last two or three countries.

With contesting it is just the opposite. Competition grows and improves steadily, and if you don't match their efforts, you'll be at the losing end. Within the limits of where I live there is not much more I can do antenna-wise. Over the years competition got fiercer, especially on the low bands. I hope and believe that some of the work I have put into publishing the techniques and art of Low Band operating has contributed to the raising of the standards and the increasing of the performance from stations on 160 and 80 meters from all around. That makes me happy.

Another evolution, over the years, is the increase of man-made noise. Whereas 15 years ago, I considered myself living in a semi-rural area where manmade noise was pretty low, I now consider myself living in a semi-residential area! Not that I have any more close neighbors, but there now are two small-industry industrial parks within 2 km of my house.

Every year I spend days chasing noise sources and trying to kill them. In the coming years we will have to continue to fight for our RF spectrum. Also the threat of BPL (PLC) certainly still exists today, although I am convinced that the inferiority of the technical concept of sending RF signals via the mains network will make sure that this technique will never be successful at a very large scale.

10. THE FUTURE

My first contest I took part in was the 1961 UBA CW contest. I was licensed less than a month before, and I won it. And yes, it was on CW. And no, there were only a few contenders… yes, I still have the cup I earned. It's one of my dearest trophies.

I now have been quite active in contesting over the past 20 years, and I have proven what I wanted to prove, which makes me happy. I built a competitive contest station, so my success in DXing on the low bands (worldwide highest DXCC score on 80 meters, and highest "honest" DXCC score on 160 meters outside the USA), are no coincidences.

Next year I'll turn 70, and for some time already I have no longer the urge to take part in every contest and to win it. I can now just participate for a couple of hours and give out points, and just simply enjoy myself doing it.

I just don't derive the same pleasure from fighting the contest battles as I did years ago. But that is normal, I guess. I get more pleasure in helping other hams in all fields related to our wonderful hobby. Having written the five editions of this book, over a span of 25 years, gives me probably more satisfaction and pleasure than scoring high in DXCC or in world wide contests.

By electing me to the CQ Contest Hall of Fame in 1997 and to the DX Hall of fame 10 years later, my friends and fellow contesters and DXers have told me I was a good contester and a good DXer, and that makes me proud and happy.

As I wrote before, ham radio is about enjoyment, satisfaction, self fulfillment and maybe a little recognition as well. If we all keep thinking about ham radio in these terms, I am sure we're on the right track!