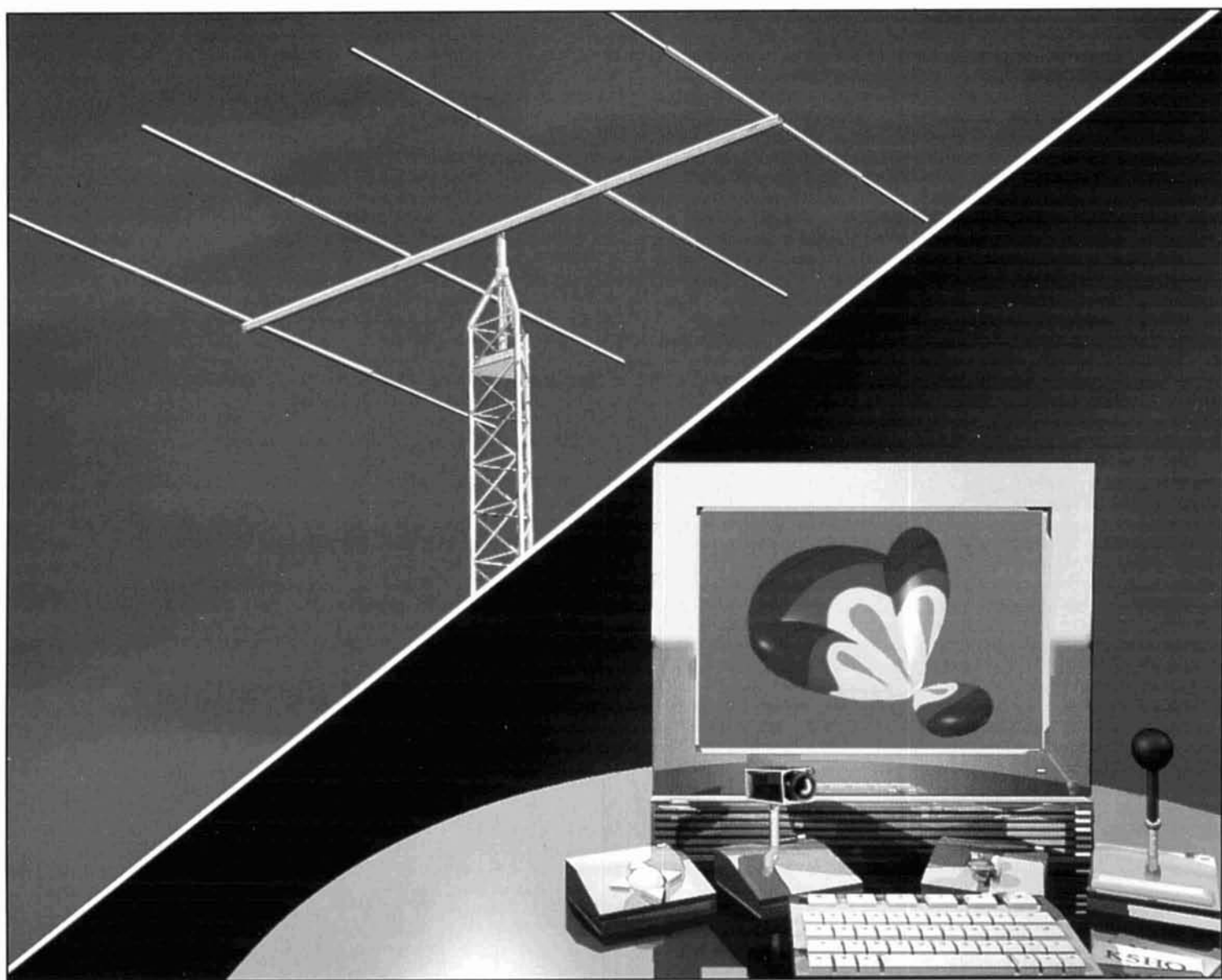


COMMUNICATIONS QUARTERLY

THE JOURNAL OF
COMMUNICATIONS
TECHNOLOGY

Spring 1993

\$9.95



- Determination of Yagi Wind Loads Using the "Cross-Flow Principle."
- The 1993 Perseids: A Meteor Storm?
- Radio Frequency Transistors: Principles and Practical Applications
- Build a High-Performance Low-Profile 20-Meter Beam
- Building the Perfect Noise Bridge
- 1.5:1 and 2:1 Baluns
- FSK Signal Monitor
- Small Loop Antennas: Part 2
- ONETWORES
- Gilbert Cell Active Mixers

REMOTE CONTROL JUST GOT EASIER



Full remote control
by a transceiver.

DR-600T *New*

We invented the remote control transceiver in 1990, and we are redefining it today. Hams have been asking for a high-power remote-control rig that is more flexible, more powerful, and less complicated: This is it.

Standard features include FULL remote control with or without a security code. The user can exercise control over both bands from one control channel. This means that a single band HT is all that is required to take full advantage of this feature. The DTMF decoder board that makes all this possible is now factory installed as standard equipment. Also included is DSQ for private paging, autodialer channels, direct frequency input from microphone, CTCSS encode, and odd splits on every memory channel.

This rig features state of the art sensitivity and selectivity, including Airband reception capability with a simple modification. For security and convenience, the head can be mounted separately from the main unit with the optional separation kit.

**Check out the affordable technology of the 90's.
Check out ALINCO.**

Alinco New Warranty program
One Full Year, Parts and Labor Factory Warranty from Alinco Electronics.
Extended Coverage Plan (ECP) available.
See your Authorized Alinco Dealer for details.

 **ALINCO**
ELECTRONICS INC.

ALINCO ELECTRONICS INC.
438 Amapola Avenue, Unit 130, Torrance, CA 90501
Tel.(310)618-8616 Fax(310)618-8758

Specifications and features are subject to change without notice or obligation.

Depend on the #1 Handheld in the World



Incredibly Tough!



DJ-580T Dual Bander
Photo for Demonstration only-
Abuse to the transceiver will void
the warranty.

twin - band HT you will find with so much power going for it.

Great sound, excellent sensitivity, and a comfortable, ergonomic design make this mini very hard to beat.

Alinco's DJ-580TA has Full-Duplex Cross Band Operation and Cross Band Repeater Functions with real world power and excellent sensitivity. Airband receive with simple modification.

This unit has built in DSQ for paging, CTCSS encode and decode standard, various scanning functions, 3 power level selections for each band, bell function, and an illuminated keypad.

Model DJ-580TA

The new DJ-580TA takes the lead in super-compact, handheld technology. Generously loaded with all the best features, this is the smallest

New MCF function allows you to set the 40 memory channels regardless of which channels you want for VHF or UHF. Any combination is possible.

If the battery is depleted to less than 5 volts, Alinco's Patented Super Low Battery Consumption Function is automatically activated. You can continue to operate the radio all the way down to 3.5 volts. This feature is effective with dry cell batteries only.

**Check out the affordable technology of the 90's.
Check out ALINCO.**

 **ALINCO**
ELECTRONICS INC.

ALINCO ELECTRONICS INC.
438 Amapola Avenue, Unit 130, Torrance, CA 90501
Tel: (310) 618-8616 Fax: (310) 618-8758

Specifications and features are subject to change without notice or obligation.

Alinco New Warranty program
One Full Year, Parts and Labor Factory Warranty
from Alinco Electronics.
Extended Coverage Plan (ECP) available.
See your Authorized Alinco Dealer for details.

KENWOOD

...pacesetter in Amateur Radio

NEW

3-Way Split

Mobile transceivers with stylish sophistication

Kenwood's new TM-742A (144MHz/440MHz) and TM-942A (144MHz/440MHz/1200MHz) FM multibanders offer prime performance plus unparalleled freedom of choice for installation (optional kit).

- **High power**
Maximum RF output is 50 watts (144MHz), 35 watts (440MHz), and 10 watts (1200MHz).
 - **Wideband receiver coverage**
The TM-742A receives from 118 to 174MHz and 410 to 470MHz; transmit ranges are 144-148MHz and 438-450MHz. The TM-942A adds the 1240-1300MHz range.
 - **New & improved detachable front panel**
The display and control sections can be separated for 3-way convenience. (with DFK-3,4,7)
 - **100 memory channels**
The 100 multi-function memory channels (all available for split operation) can be grouped into 5 banks for added convenience.
 - **Multiple scan modes**
Choose from 8 scan modes per band, plus CO (carrier-operated) and TO (time-operated) scan stops.
 - **Tri-band receive/display**
The TM-942A can receive/display all three bands (144MHz/440MHz/1200MHz) simultaneously. For the TM-742A there are four optional band units: 28MHz (50 watts), 50MHz (50 watts), 220MHz (25 watts), and 1200MHz (10 watts).
 - **Cross-band repeater, dual-in repeater, fixed-band repeater**
 - **Single-bander simplicity**
Independent SQL and VOL controls for each band enable rapid response.
 - **5 meter squelch and auto squelch**
Weak signals can be shut out. Noise squelch is also available.
 - **Supplied multi-function microphone**
Enables direct frequency entry.
 - **Clock and timer**
Includes stopwatch, alarm and on/off timer functions.
 - **Wireless remote control function**
A DTMF transceiver can be used to control various settings on the TM-742A/942A.
 - **Built-in DTSS and pager function**
The TM-742A/942A offers DTSS (Dual-Tone Squelch System) for selective calling and paging using standard DTMF tones. Elapsed time is shown by the tone alert system.
 - **Supplied accessories**
Mounting bracket, DC cable, fuses, mic, mic hook.
 - **Choice of accessories**
A full line of mics, speakers, and other accessories is available. See your authorized Kenwood Amateur Radio dealer for details!
- Specifications guaranteed for Amateur band use only.*



TM-742A/942A Mobile Transceivers

KENWOOD COMMUNICATIONS CORPORATION
AMATEUR RADIO PRODUCTS GROUP
P O BOX 22745, 2201 E. Dominguez Street
Long Beach, CA 90801-5745

KENWOOD ELECTRONICS CANADA INC.
6070 Kestrel Road, Mississauga,
Ontario, Canada L5T 1S8

93-ARD-0593

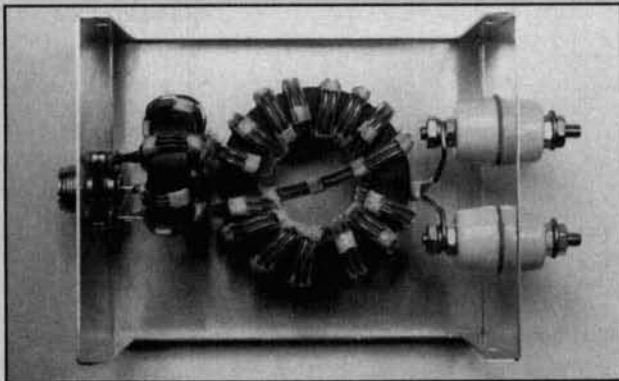
COMMUNICATIONS QUARTERLY

THE JOURNAL OF
COMMUNICATIONS
TECHNOLOGY

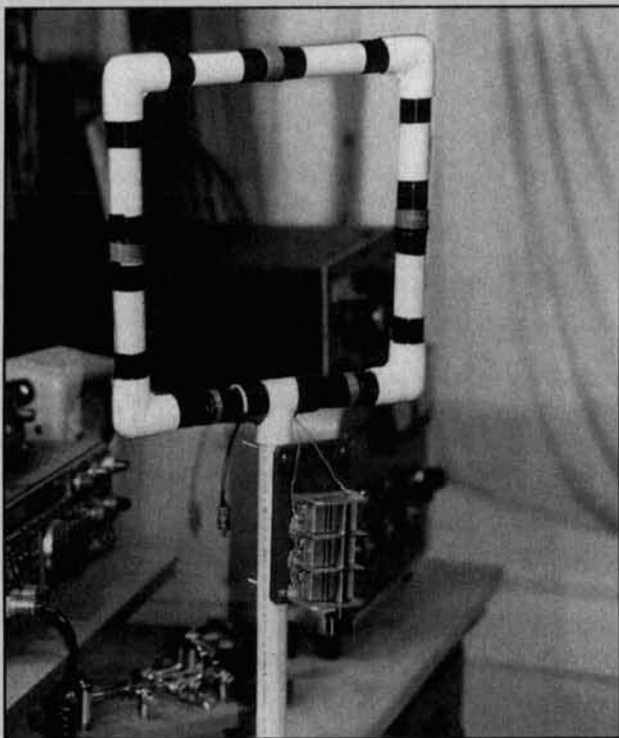
CONTENTS

Volume 3, Number 2

Spring 1993



Sevick, page 39



Carr, page 71

**13 Determination Of Yagi Wind Loads Using The
"Cross-Flow Principle"**

Dick Weber, K5IU, PE

31 1993 Perseids—A Meteor Shower?

Joseph L. Lynch, N6CL

**37 Radio Frequency Transistors: Principles And
Practical Applications**

Reviewed by Rick Littlefield, K1BQT

39 1.5:1 And 2:1 Baluns

Jerry Sevick, W2FMI

45 FSK Signal Monitor

Claton Cadmus, KA0GKC; Bruce L. Meyer, W0HZR/NNN0BQH

51 The Solar Spectrum

Peter O. Taylor

55 Building The Perfect Noise Bridge

A.E. Popodi, AA3K/OE2APM

65 Quarterly Devices

Rick Littlefield, K1BQT

71 Small Loop Antennas: Part 2

Joseph J. Carr, K4IPV

81 Tech Notes

Peter Bertini, K1ZJH

**85 Build A High-Performance Low-Profile
20-Meter Beam**

Cornell Drentea, WB3JZO

91 ONETWORES

Yardley Beers, W0JF

99 Gilbert Cell Active Mixers

F. Dale Williams, K3PUR

104 The Final Transmission

Anne Prather, KA9EHV

Cover photo: A beam against the evening sky and a communications workstation displaying the 3-D radiation pattern from the beam are depicted in this computer graphic by Bryan Bergeron, NU1N.

EDITORIAL

Working The Meteors

Amateur radio operators communicate via a number of different propagation modes. There's sporadic E, troposcatter, EME, auroral propagation, lightning scatter, rain scatter, and aircraft scatter—a whole host of ways to work other amateurs. One mode in particular, meteor scatter (bouncing a radio signal off a meteor trail), is a prime one for VHF/UHF DXing.

The meteor connection

The Earth is bombarded by large numbers of meteors every day. Ranging in size from that of a grain of sand to a pebble, these meteors burn up in the atmosphere, leaving a trail of ionized gases in their wake. On a clear night, you can see quite a few trails. The best time to view them is after midnight, when the number of meteors increases. Although the individual trails are very small and don't last long, they are capable of reflecting radio signals of up to 150 MHz—or more, in some instances. In meteor scatter communications, messages are transmitted via either SSB or CW (high-speed Morse is popular in Europe). These messages are repeated many times in order to enable receiving stations to pick them up. In the continental United States, the best months for random meteors are June through August.

Although meteors enter the atmosphere every day, their number increases greatly during certain times of the year. These incidences of increased meteor activity are called meteor showers.

Showers	Date
Quadrantids	January 1–6
April Lyrids	April 19–24
Eta Aquarids	May 1–7
June Lyrids	June 10–21
Ophiuchids	June 17–26
Capricornids	July 10–August 15
Delta Aquarids	July 15–August 15
Pices Australids	July 15–August 20
Alpha Capricornids	July 15–August 25
Iota Aquarids	July 15–August 25
Perseids	July 25–August 18
Orionids	October 16–October 26
Taurids	October 20–November 25
Cepheids	November 7–November 11
Leonids	November 15–November 19
Geminids	December 7–December 15
Ursids	December 17–December 24

Table 1. Major meteor showers.

In amateur radio circles, meteor scatter communication revolves around the various meteor showers that occur throughout the year. The major showers are listed in **Table 1**. Some showers last only a few hours, others last for several days. They can peak any time of the day or night. Because the number of meteors present in a meteor shower is so concentrated, the probability of completing a QSO is greater.

An excellent opportunity

Meteor scatter is a challenging mode to work and this summer presents an excellent opportunity to do so. It's been predicted that this year's Perseids meteor shower may, in fact, be a meteor storm. Consequently, the opportunity to make contacts will increase exponentially. Joe Lynch, N6CL, author of *CQ* magazine's "VHF Plus" column brings you up-to-date this issue with his article, "The 1993 Perseids: A Meteor Storm?" Along with a short history of meteor scatter propagation, Joe discusses the predictions for this year's Perseids and gives a run-down on how to make meteor scatter contacts.

What do I need; what do I do?

What equipment do you need to listen in or make contacts yourself? On 6 meters, a basic radio and a 3 to 5-element quad will do. For 2-meters, the equipment required is a bit more involved. You'll need a 150-watt amplifier, a sideband radio, and a 12 to 15-element 2-meter beam. As for 10 meters, you've probably already made meteor scatter contacts without even knowing it. To give it a try, simply listen for a series of bursts from a transmitting station, and return the call. Read Joe's article for helpful hints, and practice on some of the showers listed to hone your skills.

Set time aside this August to work meteor scatter during the Perseids. It's an opportunity you don't want to miss!

Terry Littlefield, KA1STC
Editor

Bibliography

1. Ian Poole, G3YWX, "Meteor Scatter—The Basics," *Practical Wireless*, April 1993, pages 28–32.
2. Joe Reisert, W1JR, "VHF/UHF World," *Ham Radio*, June 1984, pages 82–92.
3. Joe Reisert, W1JR, "VHF/UHF World," *Ham Radio*, July 1984, page 21.

EDITORIAL STAFF

Editor

Terry Littlefield, KA1STC
Consulting Technical Editor
 Robert Wilson, WA1TKH
Senior Technical Editor
 Peter Bertini, K1ZJH
Assistant Editor
 Nancy Barry

EDITORIAL REVIEW BOARD

Forrest Gehrke, K2BT
 Michael Gruchalla, P.E.
 Hunter Harris, W1SI
 Bob Lewis, W2EBS
 John Marion, W1QM
 Walter Maxwell, W2DU
 Jim McCulley, P.E.
 William Orr, W6SA1

BUSINESS STAFF

Publisher

Richard Ross, K2MGA
Advertising Manager

Arnie Sposato

Sales Assistant

Tracy Walsh

Controller

Frank Fuzia

Circulation Manager

Catherine Ross

Data Processing Manager

Melissa Kehrwieler

Data Processing

Carol Licata

Customer Service

Denise Pyne

PRODUCTION STAFF

Art Director

Elizabeth Ryan

Asst. Art Director

Barbara Terzo

Artist

Susan Reale

Electronic Composition Manager

Edmond Pesonen

Production Manager

Dorothy Kehrwieler

Photographer

Pat Le Blanc

A publication of

CQ Communications, Inc.

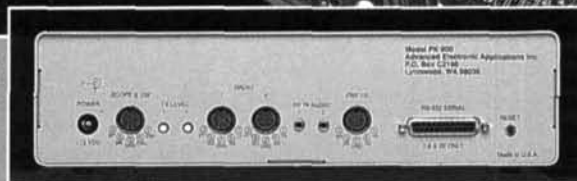
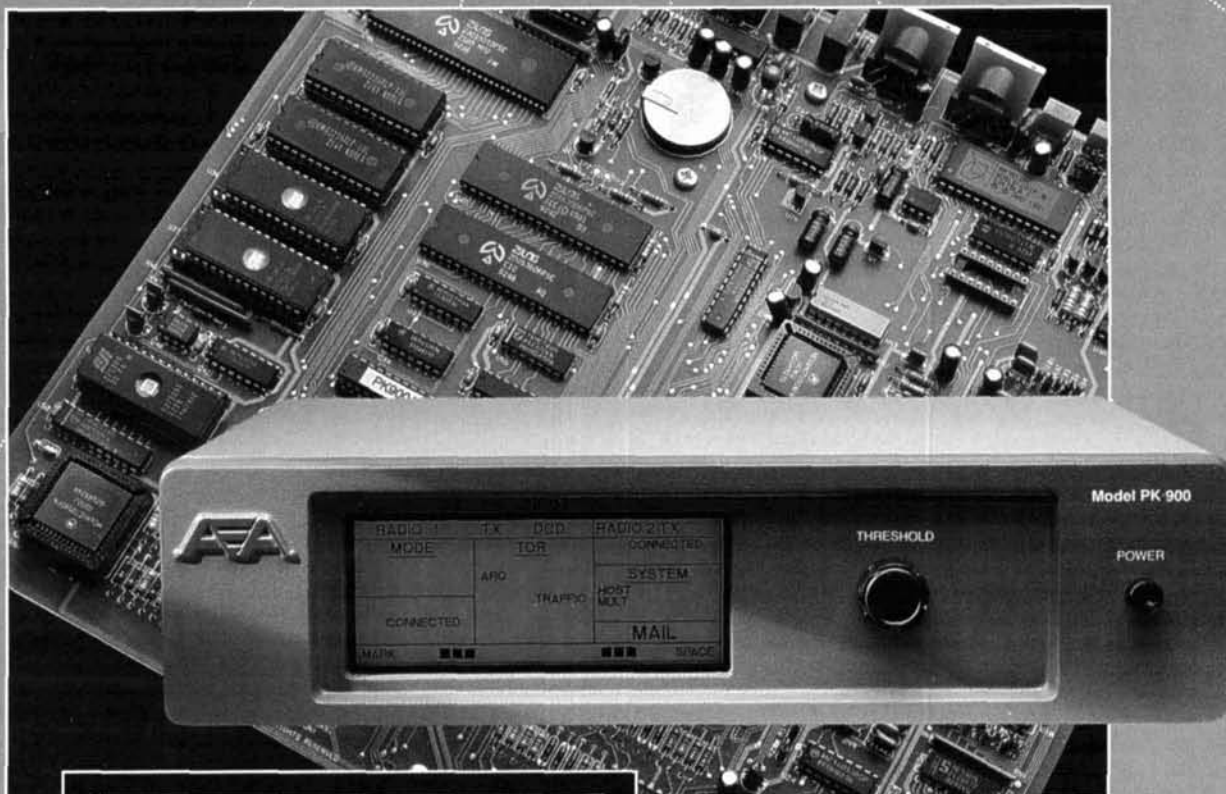
76 North Broadway

Hicksville, NY 11801-USA

Contents copyrighted CQ Communications, Inc. 1993. **Statement of Ownership, Management and circulation** Feb. 1, 1993. Communications Quarterly, 76 North Broadway, Hicksville, NY 11801. Publication #1053-9344. Issued four times a year, subscription price \$29.95 per year. Publisher: Richard A. Ross; Editor Terry Littlefield. owned by CQ communications, Inc. Stockholders: Richard A. Ross, Alan M. Dorhoffer, Thomas Kneittel, Arthur S. Salsberg. Circulation (Average of Preceding 12 Months): Net Press Run 6,204, Sales Through Dealers and News Agents 662, Mail Subscriptions 4,242, Total Paid 4,904, Free distribution 300, Total Distribution 5,204, Copies Not Distributed 1,000, Returns from News Agents 0, Total 6,204. Circulation (Single issue nearest filing date): Net Press Run 6,828, Sales through Dealers and News Agents 925, Mail Subscriptions 4,603, Total Paid 5,528, Free distribution 300, Total Distribution 5,828, Copies Not distributed 1,000, Returns from News Agents 0, Total 6,828. /s/ Richard A. Ross, Publisher. Second-class postage paid at Hicksville, NY and additional mailing offices.

Postmaster: Please send change of address to Communications Quarterly, CQ Communications, Inc., 76 North Broadway, Hicksville, NY 11801. ISSN 1053-9344
 Printed in the United States of America

INTRODUCING THE PK-900... NEW FROM THE INSIDE OUT!



**IT'S THE NEXT GENERATION
IN MULTI-MODE CONTROLLERS.**

Now, there's a new standard of excellence in multi-mode digital controllers...the new PK-900 from AEA. It incorporates all of the features which made the PK-232 the most popular multi-mode controller in the industry. But that's just the start. AEA's new PK-900 also features dual port HF or VHF on either port; low cost 9600 baud plug-in option; memory ARQ and VHF DCD state machine circuit; powerful triple processor system; zero crossing detector for the sharpest Gray Scale FAX you've ever seen; and many other new software selectable features.

Inside and out, the new PK-900 from AEA is what other multi-mode controllers will now be measured against.

- Processors used: Zilog 64180, Motorola 68HC05C4, Motorola 68HC05B4
- Data rates: 45 to 1200 baud standard, up to 19.2K baud with external modems
- Dimensions: 11.75" (29.84cm) x 11.75" (29.84cm) x 3.5" (8.89cm) Weight: 4.6 lbs. (2.08 kg)
- Power requirements: 12 VDC at 1.1 amps

Connect with us for what's new in multi-mode controllers.
Call our literature request line at 1-800-432-8873.

Advanced Electronic Applications, Inc.

PO Box C2160, 2006 - 196th St. SW, Lynnwood, WA 98036



Connect with us

LETTERS

Don't lose sight of your objective

Several days ago I received the spring Communications Quarterly and was disappointed to observe from the article "Radio Receivers of the Past" that your magazine is perhaps losing sight of the original objective of providing the latest communications technology?

The article by Joseph Carr is well written and interesting, but is clearly published in the wrong magazine. I am really surprised it was accepted for publication in your magazine. There are numerous other magazines which are more appropriate for an article of this kind.

If you are struggling for suitable technical articles, I can appreciate you have a problem and I sincerely hope this is not the case; but if this article was chosen ahead of article addressing current technical issues, I deplore your choice.

It would be nice to see an article on practical application of digital signal processing for a local oscillator of an HF transceiver.

I normally really enjoy your magazine and I wish you all the best for its future.

I sincerely believe if you lose your direction and move away from solid technical articles concerning the present and future, your subscribers like me, who only want this type of article, will drop the magazine. Remember, the other sort of articles are already well accommodated.

**Ian Williams, VK3MO
Victoria, Australia**

A reader disagrees

I must say that I am not in agreement with your editorial in the Summer 1992 issue. Your definition of your friend getting his license in the "traditional way" does not (necessarily) coincide with mine. You'd be surprised at the number of incredulous looks I get whenever I describe the earlier days of amateur radio when there were *no* novices. I have been told—several times—"But there have *always* been novices!"

I have always had a less than pleasant feeling about the lowering of the entrance requirements to amateur radio since the early '50s. *No*, it's not a case of "since I had to pass 13 wpm CW"—it's a feeling that the license should not be *given* away.

You are surprised that the advent of the no-code license had brought more into the fold? My God! If Ford *gave away* Aerostar

vans and people flocked to get them for nothing, would you applaud the method and be surprised that a large crowd arrived to get a freebee? I really think that's being a tad naive, no?

I don't care whether it's CW or having to climb a greased pole, the license should involve some *work* and should *require* a strong desire to achieve the granting of it. Otherwise, its gaining has *no meaning* to the recipient. He or she gains no self-respect, no self-esteem and the license ends up not being a device that elevates the person.

The idea of "try before buy" doesn't wash, either. I was licensed in 1947—no code tapes, no computer CW programs, no study guides, no digital tutorials. Just hard studying, hard listening to an old *Comet Pro* receiver to learn the code. Maybe a *year* of intensive study to take a single level entry test—the (traditional) CLASS B license exam! Thirteen words per minute (sending and receiving) and a comprehensive theory/regulations test. No *novice* entry, no stepped upgrade levels—only the CLASS A privileges to study further for and to wait at least one year to test for. You *had* to *want* the ticket *before* you went through all of that! And once you held it in your hands, you *knew* that you would never do anything that would tarnish that piece of paper or all the hams who had gone before and earned the privileges you now were party to.

Why not give away college degrees? Boy Scout merit badges? Purple Hearts? Let 'em *try* one on before you give one to them, just to see if they *like* it? Be honest! The swelling of the ranks is only desirable because of the sheer *numbers*, the political *clout*. I've never heard a *single* word about the *quality* of the new recruits—only their quantity.

Make the license just a little more difficult to obtain. Increase the technical requirement just a little, put in more questions about proper operations and the regulations. Make sure they *read and understand* Part 97—all of it! Make pure memorization less easy. Then we all can be proud of the new additions and welcome them to the ranks with open arms and companion pride in their first accomplishments.

**Joe Weite, KH6GDR, T32BF, OX5JW,
WA6VTX, WA6VTX/KL7, W2WJM,
W2WJM/VO1
Makakilo, Hawaii**



JRL-2000F

Fully Automatic MOSFET HF LINEAR AMPLIFIER

- 1 kW NO-TUNE POWER AMPLIFIER
- 48 MOSFETs SINGLE ENDED PUSH-PULL (SEPP) DESIGN
- BUILT-IN AUTOMATIC ANTENNA TUNER
- HIGH-EFFICIENCY SWITCHING POWER SUPPLY



The JRL-2000F is the world's first MOSFET HF linear amplifier, designed using the same high technology found in JRC's professional high-power radio transmitters. Featuring a heavy-duty power amp that incorporates 48 RF power MOSFETs to ensure low distortion and clean output up to 1,000 watts (100% duty cycle, 24 hour) SSB/CW, plus a

high-speed automatic antenna tuner with memory capacity of 1820 channels for instant QSY. Plus a high efficiency switching power supply (80V-264V) with power factor correction to suppress AC line currents, an automatic antenna selector for up to four antennas and a wireless remote control unit.



Japan Radio Co., Ltd.

430 Park Ave, 2nd Floor New York, NY 10022

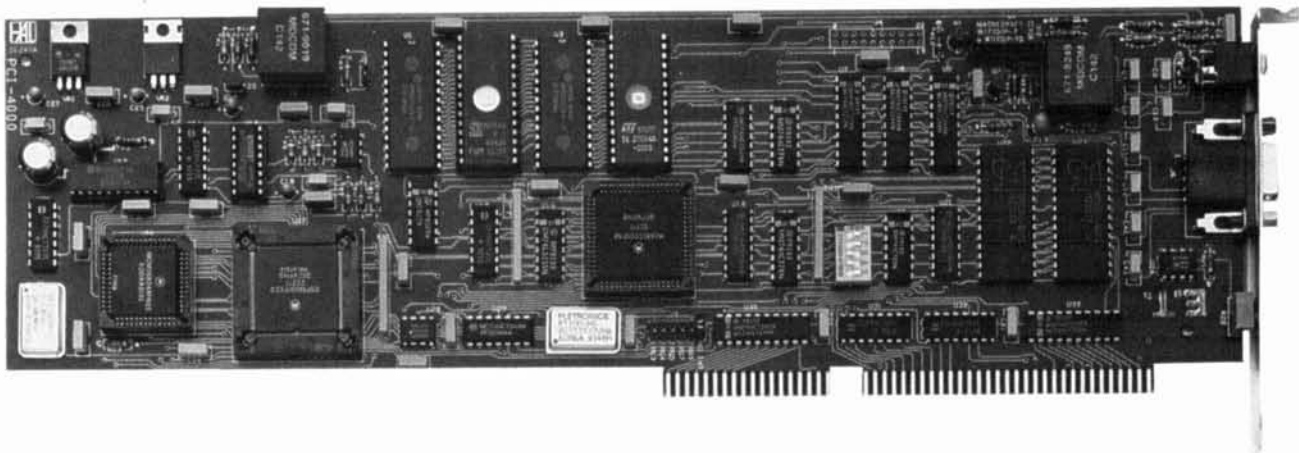
Phone: (212)355-1180 Fax: (212)319-5227

Telex: 961114 JAPAN RADIO NYK

NEW!

HAL Announces the PCI-4000 PC-CLOVER System

For Fast, Bandwidth-Efficient HF Data



The PCI-4000 uses the latest development in HF data transfer methods—CLOVER-II. CLOVER-II is designed to maximize the amount of data which can be transferred in a narrow bandwidth over HF radio frequencies. It uses a combination of four tone frequencies with phase and amplitude modulation to achieve data transfer rates as high as 60 characters per second—about ten times faster than AMTOR. The PC-CLOVER system incorporates Reed-Solomon error correction, not simply a retransmission scheme. The PCI-4000 is a full-sized PC card which operates in a 80286-based PC or higher.

The PCI-4000 PC CLOVER system features:

- ♣ Higher throughput than RTTY, AMTOR, Packet, or PACTOR on similar HF channel
- ♣ Simple pull-down menu operation
- ♣ Signal bandwidth of 500 Hz (@ 50 dB down)
- ♣ Plugs into your PC (286, 386SX, 386, or 486 machines)
- ♣ Easy interface to your transceiver
- ♣ Automatically adapts to HF band conditions
- ♣ Error correcting using Reed-Solomon error correction

You've read about it in the articles. Now you can operate CLOVER!
Order your PC-CLOVER system today from HAL Communications Corp.

PCI-4000 PC-CLOVER System

Only \$995.00



HAL Communications Corp.
P.O. Box 365
Urbana, IL 61801
Phone (217) 367-7373
FAX (217) 367-1701



The Fastest Counter Starts with the Fastest Chip

Optoelectronics Puts the OE10 Inside

Since the 70's the frequency counter chip most used was the 7216 (8 digits,

10 million counts per second). The only way to have a faster more reliable counter was to design our own chip. So we did — the OE10. Faster — 10 digits,

220 million counts per second. Reliable — zero field failures.



MADE IN THE USA

And now we have added a microprocessor to unleash the power of the OE10.

Optoelectronics introduces the 3000A Multifunction Micro HandiCounter® with these new and exciting features:

- **Digital Filter Technique** (patent pending)
Filters stable signals from background clutter
- **Digital Auto Capture** (patent pending)
Holds present reading and stores previous reading for hands free unattended operation
- **Works Even in Presence of Strong RF Signals**
- **Stores Frequencies**
- **Increased Sensitivity**
- **Synchronous RF Detection**
Super sensitive 16 segment bargraph operates independently from counter
- **High Speed ASIC**
220 million counts per second.
- **10Hz-3GHz**
- **Multifunction HandiCounter®**
- **Dual High Impedance Amplifiers**
For ratio and time interval functions
- **High Resolution**
Direct count to over 200MHz
- **Extended Battery Use**
- **1 Year Full Warranty**
- **All This and More for the Low Price of \$279.**

Options for the NEW 3000A

- TCX030
± .2ppm\$100.
- BL30/Backlight45.
- Antenna - TA100S
Telescoping Whip Antenna12.
- Antenna Pack 2
Includes RD27, RD50, RD100
RD440, RD800 - Save \$3299.
- CC30
Padded black vinyl
carrying case15.

Factory Direct Order Line

1-800-327-5912

305-771-2050 • Fax 305-771-2052

5821 NE 14th Ave, Ft. Lauderdale, FL 33334

5% Ship/Handling (Max \$10) U.S. & Canada. 15% outside continental U.S. • Visa, Master Card, C.O.D. (Cash or Money Order only).

OPTOELECTRONICS
BE SURE YOU HAVE AN OE10 INSIDE

VERSATILITY PLUS +



L.L. Grace introduces our latest product, the **DSP-12 Multi-Mode Communications Controller**. The **DSP-12** is a user programmable, digital signal processing (DSP) based communications controller.

FEATURES

- Multi-tasking operating system built in
- PC-compatible (V40) architecture allows development of custom applications using normal PC development tools and languages
- Motorola DSP56001 DSP processor
- Serial interface speeds from 110 to 19200 bps
- Optional 8-channel A-to-D & DAC for voice and telemetry applications
- 12-bit conversion architecture
- V40 source code and schematics available
- RAM expandable to one megabyte. Useable for mailbox feature, voice mail and development
- EPROM expandable to 384k bytes
- Low power requirements: 10-15vdc, 750ma
- 3 analog radio connectors. RX & TX can be split in any combination. Programmable tuning outputs are available on each connector
- Many modems available in the basic unit, including Packet, RTTY, ASCII, and PSK modems for high speed packet and satellite work
- Both V40 and DSP programs can be down-line-loaded from your PC or a bulletin board. You can participate in new development!
- Built in packet mailbox
- V40 and DSP debuggers built in
- Open programming architecture
- Free software upgrades
- Low cost unit
- Room for future growth

APPLICATIONS

- HF Packet
- HF RTTY & ASCII, including inverted mark/space and custom-split applications
- VHF Packet
- 400bps PSK (satellite telemetry)
- 1200bps PSK (satellite & terrestrial packet)
- V26.B 2400bps packet
- 9600bps direct FSK (UO-14)
- Morse Code

CUSTOM APPLICATIONS

- Voice compression
- Telemetry acquisition
- Message Store-and-Forward
- Voice Mail

COMING ATTRACTIONS

(Remember, software upgrades are free!)

- WEFAX and SSTV demodulators
- NAVTEX
- AMTOR and SITOR
- Multi-tone Modems
- ARINC ACARS

Commercial inquiries are welcomed. We offer rapid prototyping of custom commercial, civil, and government applications including intelligent radio, wireline, and telephone modems.

DSP-12 Multi-mode Communications Controller	\$ 595.00
One Megabyte RAM Expansion Option	149.00
Date/Time Clock Backup Option	29.00
8-Channel A-To-D Telemetry/Experimentation Option	49.00
Wall-Mount Power Supply for DSP-12 (110 vac)	19.00

We accept MasterCard & VISA and can ship C.O.D. within the USA. All orders must be paid in US Dollars. Shipping & Handling: \$5 (\$20 International).

L. L. Grace Communications Products, Inc.

41 Acadia Drive, Voorhees, NJ 08043, USA

Telephone: (609) 751-1018

FAX: (609) 751-9705

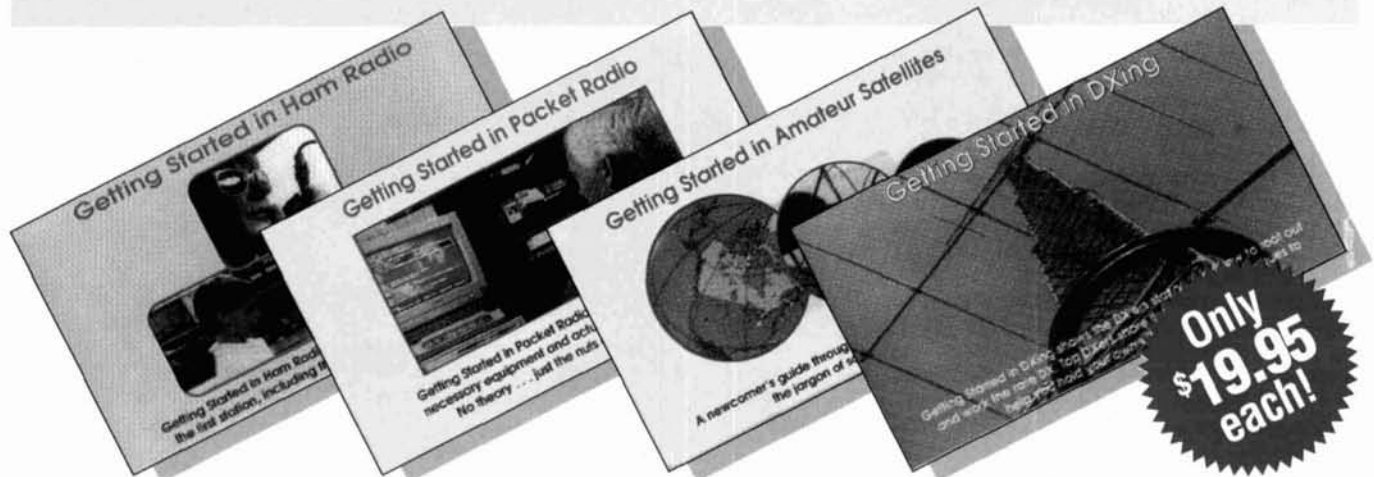
CompuServe: 72677,1107

1/91

L. L. Grace also manufactures the Kansas City Tracker family of satellite antenna aiming systems. Call or write for more information.



CQ's new videos bring the exciting world of amateur radio into your living room!



Getting Started in Ham Radio.

The authoritative fast paced video introduction to the fascinating world of ham radio. Shows how to select equipment and antennas. How to use repeater stations for improved VHF coverage. The importance of grounding and the basics of soldering. How to get the most out of your station, whether it's home-based, mobile or hand held!

Getting Started in Packet Radio.

This video will help de-mystify the exciting but sometimes confusing world of packet radio. Shows you how to get started in using your computer on the radio. Includes step-by-step instructions on making packet contacts and using packet bulletin boards, networks and satellites.

Getting Started in Amateur Satellites.

Shows you how veteran operators set up their satellite stations and how to find and track ham satellites with ease. How to access current satellites and contact far ranging countries around the world. This video is filled with easy to understand advice and tips that can't be found anywhere else.

Getting Started in DXing.

Top DXers share their experience on equipment, antennas, operating skills and QSL-ing. See hams work rare DX around the world. If you're new to DXing this video is for you. All the valuable information presented here may well give you the competitive edge you need to master the exciting world of DXing.

The New CQ Video Library!

Now, for the first time, CQ brings you all our experience in an exciting new format. Our videos will help you get started and improve your existing skills in amateur radio with our informative, helpful and easy-to-follow instructional videos.

Whether you're a new ham or just getting started in Packet, Satellite or DXing we've got your video. Each tape gives you first-hand tips from hams who are on the air. You'll get advice from some of our hobby's most experienced users. They take you under their wing and show you just what it takes to get on the air!

Filled with step-by-step tips in setting up your first station, operating satellites, using packet radio, maximizing your DXing potential, these videos offer insights for beginners and experienced operators alike. You can order just the ones you want, or all four. It's up to you.

Send for your first video today and start growing now in the exciting and adventurous world of amateur radio!

CQ Communications
76 North Broadway, Hicksville, New York 11801
For Faster Service Call 516-681-2922 or FAX 516-681-2926



The new R-2550.
COMMUNICATIONS SYSTEM ANALYZER

SPECIAL INTRODUCTORY PRICE

At only \$8695, it produces like a money tree.

In today's economy, your shop needs maximum value – top performance at lower cost.

Motorola introduces the R-2550, with the combination of performance features you need – variable level duplex generator

• auto calibrated spectrum analyzer • auto calibrated field strength meter • large (7" diagonal) CRT • digital storage oscilloscope • dozens of additional features – all at a price that will make you think you planted a money tree.

For information, call:

1-800-235-9590

Or write P.O. Box 2606
Scottsdale, AZ 85252



MOTOROLA

Quality means the world to us.™

Determination of Yagi Wind Loads Using the "Cross-Flow Principle"

Find Yagi wind loads with this well-known method

I'd like to talk about a well-known and proven method of determining the wind loading on structural members. I've applied it here to a Yagi with horizontal elements to find the forces caused by wind loading. After defining and discussing the forces, I'll recommend a standard for comparing the wind loading of Yagis.

In the course of my work, I learned that the method currently used to define wind loads on Yagis is incorrect. Although this method has been used for over twenty-five years, it has gone unquestioned. Today, this incorrect method is used widely in the amateur radio antenna industry, and in amateur radio literature and software. I hope a simple standard based on sound, proven scientific methods will be adopted, and that these methods will be used throughout the amateur antenna industry. I hope the work presented here will help people become familiar with the concepts upon which the suggested standard is based.

Forces Due to Wind

When an object—like an element or boom of a Yagi antenna—is in a wind stream, a force results when the wind strikes the object. The magnitude of the force depends on the velocity of the wind, the size of the object, the shape of the object, and the object's orientation relative to the wind stream. As the wind strikes an

L/D	C
0 – 4	0.7
> 4 – 8	0.8
> 8 – 40	1.0
> 40	1.2
$L/D = \frac{\text{tube length}}{\text{tube diameter}}$	

Table 1. Drag coefficient for round tube.

V (mph)	70.0	80.0	86.6	100.0
P (psf)	12.5	16.4	19.2	25.6

Table 2. Values of P at several values of wind velocity.

object, it creates a dynamic wind pressure that acts on the object's area. The higher the wind velocity, the higher the pressure. It's possible to calculate the dynamic wind pressure using **Equation 1**.

$$P = 0.5 m V^2 \quad (1)$$

where:

P = dynamic wind pressure
m = air mass density
V = air velocity

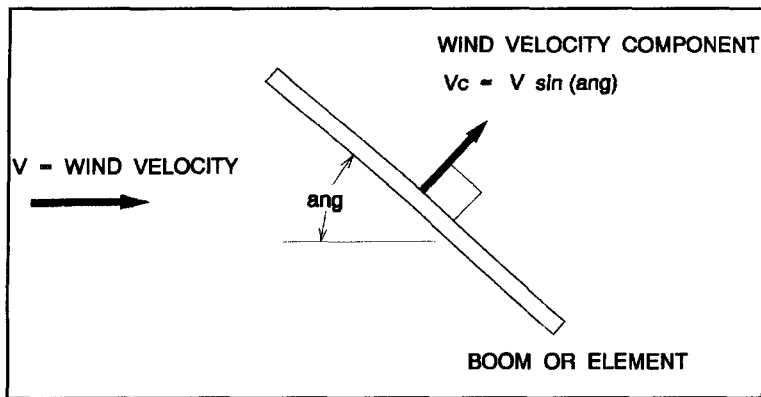


Figure 1. Wind velocity component as a function of angle.

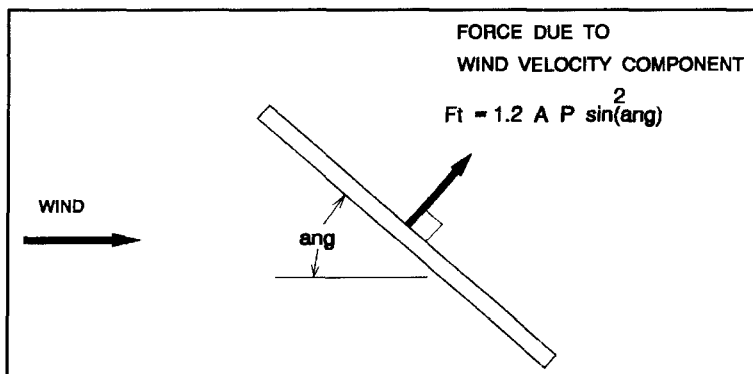


Figure 2. Force on round tube inclined to wind.

Equation 1 shows that the dynamic wind pressure depends on wind velocity and air mass density. To simplify matters, we use the density of air at “standard sea-level” With wind velocity in MPH and using “standard sea-level” air mass density, **Equation 1** is simplified to provide **Equation 2**.¹

$$P = 0.00256 V^2 \quad (2)$$

where:

$$P = \text{dynamic wind pressure (psf)}$$

$$V = \text{wind velocity (MPH)}$$

Now that we can determine the dynamic wind pressure, we need a relationship that accounts for the size and shape of the object.

Equation 3 gives this relationship for wind perpendicular to the object.

$$F_w = C A P \quad (3)$$

where:

$$F_w = \text{force (lbs)}$$

$$P = \text{dynamic wind pressure (psf)}$$

$$A = \text{projected area (square feet)}$$

$$C = \text{drag coefficient (no units)}$$

Consider the projected area of an object as the object’s shadow or broadside area. For a round tube, we would find the projected area—multiplying the tube’s diameter by its length. This tells us something about the tube size, but nothing about the object’s streamlined nature. Enter a term called the drag coefficient. The lower the drag coefficient, the lower the force. For a round tube that’s long compared to its diameter, $C = 1.2$.² For round tubes that aren’t long compared to their length, C is less. There’s a logical reason for this. Air only flows over the sides of an infinitely long tube, while shorter tubes experience flow over both the sides and the ends. With flow over the ends, there’s less drag produced relative to the overall length of the tube. **Table 1** shows this effect. For HF Yagis, L/D is generally greater than 40. As a result, we’ll use $C = 1.2$ for round tubes in this article. As a point of interest, a flat plate which has $C = 2.0$ is a common shape used for mounting plates and brackets.²

Clarification of Wind Pressures

Table 2 gives values of P at several levels of wind velocity. The pressures in **Table 2** may appear to be lower than those found in various books or articles discussing wind loads on towers and antennas. Admittedly, this is an area of much confusion. Some people are familiar with standard EIA/TIA 222-E, while many more are familiar with the terminology of a prior version—EIA 222-C—issued by the Electronic Industries Association. In these references, factors for gusts and other “modifiers” are included in equations used to define wind pressure.^{3,4} Here, **Equation 2** determines the dynamic wind pressure at the exact wind speed, V .⁴ No additional factors are embedded to account for gusts or other effects.

Wind loading on an inclined member

An HF Yagi antenna usually consists of an assembly of tubes. Round tubes are used for the elements and another round tube, or assembly of round tubes, is used for the boom. Because the elements are mounted perpendicular to the boom, finding the resulting wind load of a complete antenna requires knowledge to determine the forces generated on each member when it’s not broadside to the wind, but at some angle of attack. It is important to be able to determine the total wind load on the antenna for all angles relative to the wind to ascertain the worst case force that must be handled by the supporting tower system.

Figure 1 shows a tube inclined to a wind

stream. When the angle is 90 degrees, we can use **Equation 3** to find the resulting force. For angles less than 90 degrees, the issue becomes more complex. To begin, we'll use the "cross flow principle."⁵ To quote Dr. S.F. Hoerner from his book *Fluid-Dynamic Drag*, "At an angle of attack, ang, flow pattern and fluid dynamic pressure forces of such bodies only correspond to the velocity component (and the dynamic pressure) in the direction normal to their axis."⁵ In **Figure 1**, the wind velocity component perpendicular to the tube causes a dynamic wind pressure that acts on the entire length of the tube. In other words, we have found the component of the wind velocity perpendicular to the tube and, knowing this, can use **Equation 2** to find the dynamic wind pressure perpendicular to the tube. We then use **Equation 3** to find the resulting force. With the tube inclined to the wind at an angle ang, **Equation 4** gives the velocity component perpendicular to the tube.

$$V_c = V \sin(\text{ang}) \quad (4)$$

where:

- V_c = wind velocity component perpendicular to the tube (MPH)
- V = air velocity (MPH)
- ang = angle of attack (deg)

Now that we can find the velocity component, V_c , we can use **Equation 2** to determine the dynamic wind pressure acting on the tube. Using **Equation 2** with V_c gives:

$$P_c = 0.00256 V_c^2$$

where:

- P_c = dynamic wind pressure on inclined tube at angle ang (psf)

Substituting **Equation 4** for V_c gives us **Equation 5**, which is in terms of the wind velocity and angle of attack.

$$P_c = 0.00256 V^2 \sin^2(\text{ang}) \quad (5)$$

In **Equation 5** the term $0.00256 V^2$ is the value of P broadside to the wind, as defined in **Equation 2**. Based on this, we can rewrite **Equation 5** to obtain **Equation 6**.

$$P_c = P \sin^2(\text{ang}) \quad (6)$$

This equation shows the resulting dynamic pressure on the tube, where P is the dynamic wind pressure broadside to the wind. Now that we can find the dynamic wind pressure acting on the tube, it's possible to combine **Equations**

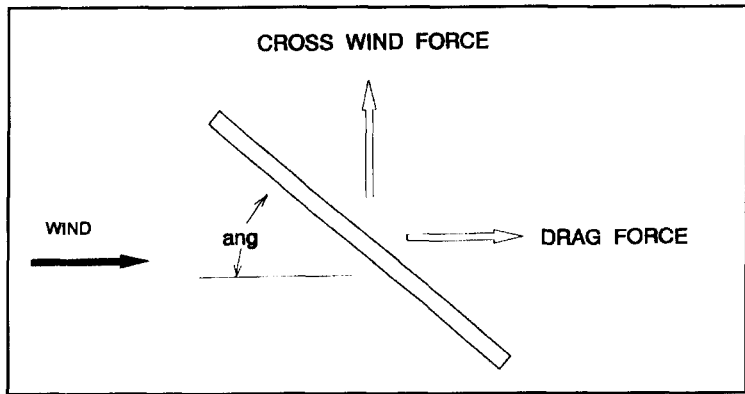


Figure 3. Drag and cross-wind forces on inclined tube.

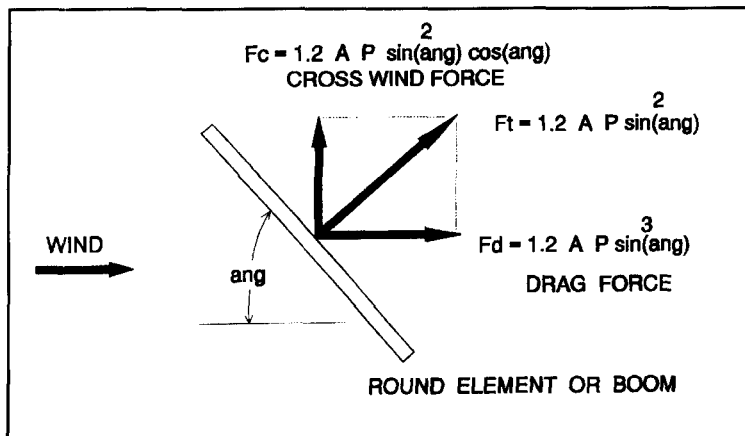


Figure 4. Drag and cross-wind forces.

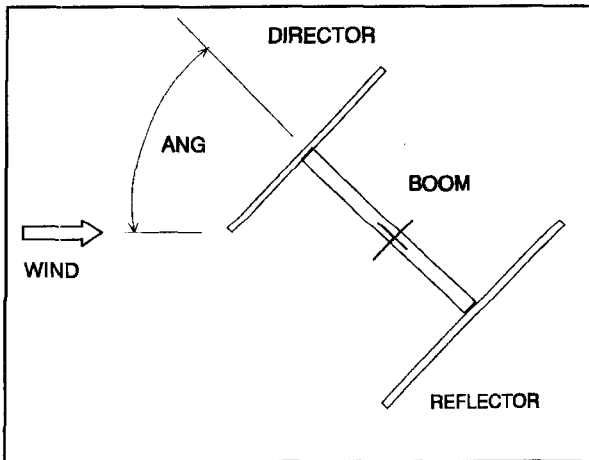


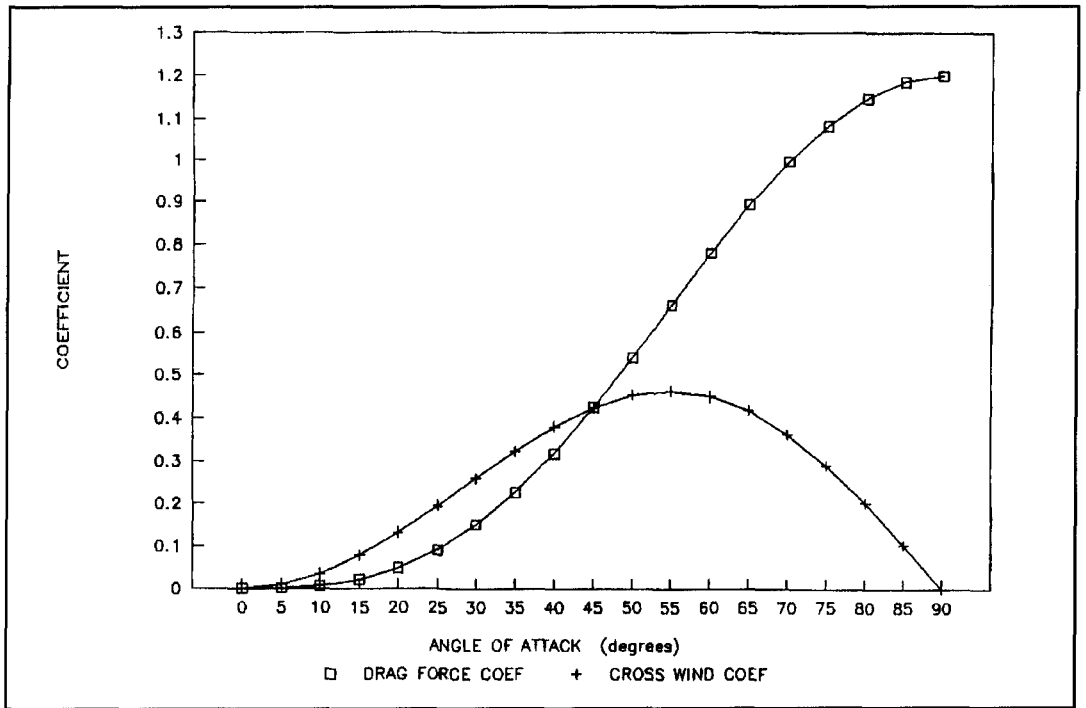
Figure 5. Simple example Yagi.

6 and **3** to form an equation which gives the force perpendicular to a object when it is inclined to a wind stream.

$$F_m = C A P \sin^2(\text{ang}) \quad (7)$$

where:

- F_m = force perpendicular to object (lbs)



Graph 1. Drag and cross-wind coefficients versus angle of attack.

Because we are mainly concerned with round tubes, C can be set equal to 1.2 in Equation 7 to provide Equation 8. Figure 2 illustrates Ft acting on a round tube inclined in a wind stream. Note that the “cross-flow principle” applies to all shapes, not just round objects. If the object was a flat plate, the value of the drag coefficient would be C = 2.²

$$F_t = 1.2 A P \sin^2(\text{ang}) \quad (8)$$

where:

Ft = force perpendicular to round tube (lbs)

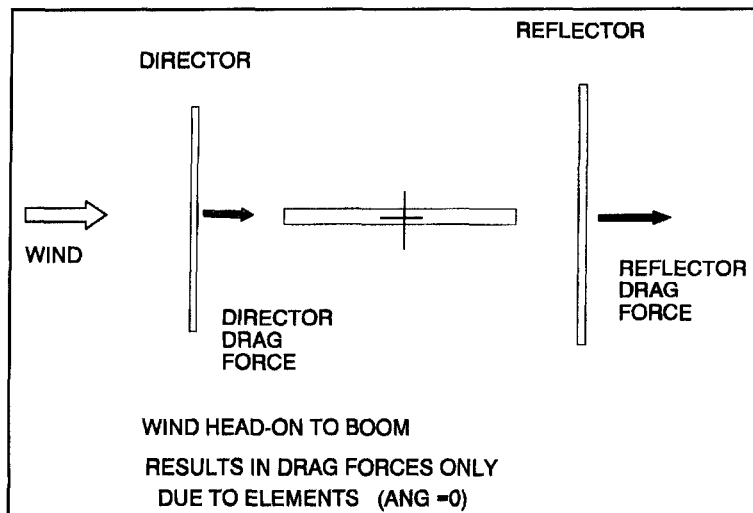


Figure 6A. Wind head-on to boom. (ang = 0).

P = 0.00256 V² (psf)
V = wind velocity (mph)
A = projected area of tube (square feet)
(= length x diameter)
ang = angle of attack (deg)

When ang = 90 degrees, the wind velocity component perpendicular to the tube is the same as the stream velocity. With ang = 90 degrees, Equation 8 yields:

$$F_t = 1.2 A P$$

which is the same as Equation 3 with C = 1.2. This is as it should be, because Equation 3 is for a tube broadside to the wind.

Please note that the projected area of the tube is still the broadside area found by the product of its length and diameter. In determining the force on an inclined tube, it is the component of the wind velocity that changes as a function of angle causing a variation in pressure loading. The change in force is **not** due to a difference in the tube area being exposed to the wind stream.

Drag and cross-wind forces

Equation 8 is used to find the force that results when a round tube is inclined at an angle in a wind stream. It is perpendicular to the tube, as shown in Figure 2. When there are multiple tubes in an assembly, it's common to simplify matters by finding the components of the tube forces which are in line and perpendic-

ular to the wind. These forces are called the drag and cross-wind forces, respectively.^{5,6,7} **Figure 3** gives an example that shows the cross-wind force, F_c , and the drag force, F_d , for the case in **Figure 2**. When there are multiple tubes, drag forces for all tubes are summed as are all cross-wind forces—making the accounting of the forces manageable. To do this, F_t is broken into the drag and cross-wind force components as shown in **Figure 4**. F_c and F_d are found using **Equations 9** and **10**, respectively.

$$F_c = F_t \cos(\text{ang}) \quad (9)$$

where:

$$F_c = \text{cross-wind force (lbs)}$$

$$F_d = F_t \sin(\text{ang}) \quad (10)$$

where:

$$F_d = \text{drag force (lbs)}$$

Substituting **Equation 8** for F_t in **Equations 9 and 10** yields **Equations 11 and 12** for F_c and F_d , respectively.

$$F_c = 1.2 A P \sin^2(\text{ang}) \cos(\text{ang}) \quad (11)$$

$$F_d = 1.2 A P \sin^3(\text{ang}) \quad (12)$$

If **Equations 11 and 12** were rewritten in the general form of **Equation 3**, a similar form would result, as shown in **Equations 13 and 14**. While **Equation 3** uses a fixed drag coefficient, C , **Equations 13 and 14** have an angle-dependent cross-wind coefficient, C_c , and an angle-dependent drag-force coefficient, C_d .

$$F_c = C_c A P \quad (13)$$

$$F_d = C_d A P \quad (14)$$

Where:

$$C_c = 1.2 \sin^2(\text{ang}) \cos(\text{ang}) \quad (15)$$

$$C_d = 1.2 \sin^3(\text{ang}) \quad (16)$$

The equations for C_c and C_d describe a variable coefficient as a function of the angle of attack. They are normally referred to as the cross-wind and drag coefficients for inclined tubes.^{5,7,8} The behavior of these coefficients are shown in **Graph 1** which shows C_c and C_d as a function of angle. The drag force coefficient is maximum at $\text{ang} = 90$ degrees. This is expected as the tube is fully broadside to the wind. What's interesting is that the cross-wind coefficient peaks at 54.75 degrees, which

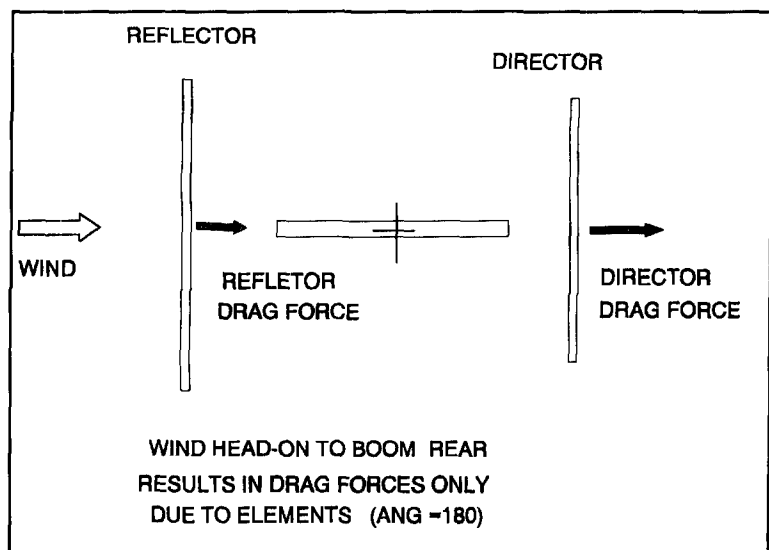


Figure 6B. Wind head-on to boom rear. (ang = 180).

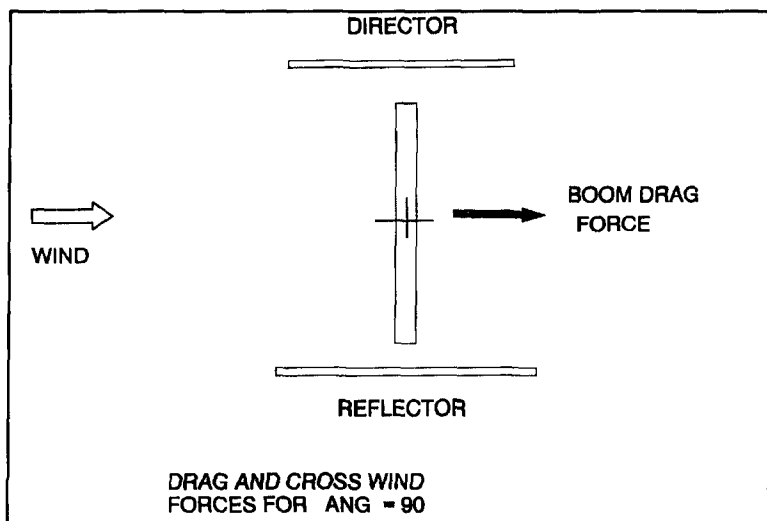


Figure 6C. Drag forces for ang = 90.

means the cross-wind force also peaks at 54.75 degrees. It's important to note that the cross-wind and drag-force coefficients are equal at 45 degrees. With these coefficients equal, the force produced in line with the wind is equal to the force produced perpendicular to the wind when a tube is 45 degrees relative to the wind.

There have been numerous experimental efforts that have verified C_c and C_d . Some were undertaken as early as seventy years ago.⁵ Although not currently used in the amateur radio antenna field, many engineering books and references contain values for C_c and C_d . One of the more common references for engineers is *Mark's Standard Handbook for Mechanical Engineers*. On page 11-78 in the 9th edition, there's a chart for values of C_c and C_d at various angles between 0 and 90 degrees that agree with **Graph 1**.⁷ *Mechanical*

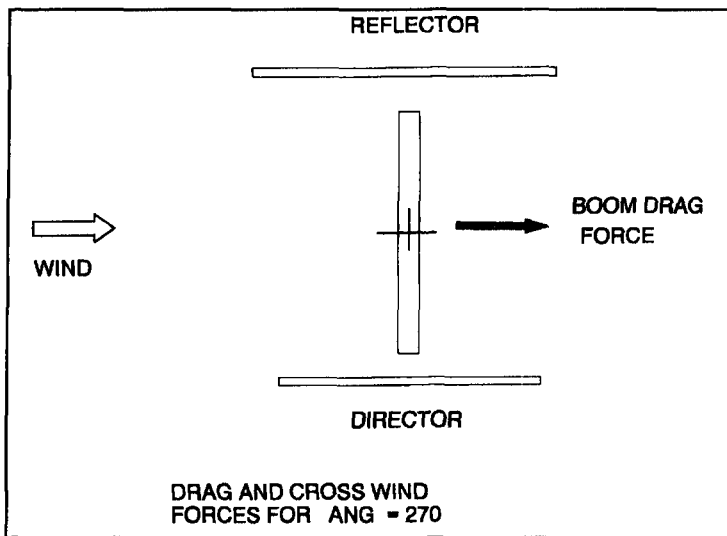


Figure 6D. Drag forces for $\text{ang} = 270$.

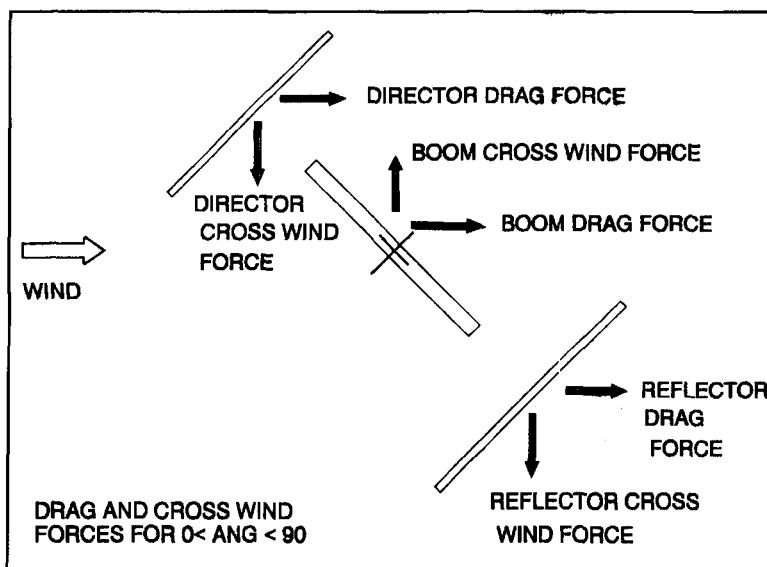


Figure 7A. Drag and cross-wind forces for $0 < \text{ang} < 90$.

Engineering in Radar and Communications is a second book commonly used by mechanical engineers involved in antenna structural design. On page 165 of the 1969 edition, a diagram and chart are shown that also list the values of C_c and C_d for 0 to 90 degrees.⁸ These, too, match **Graph 1**. The EIA/TIA 222E is a third reference that shows **Equations 15** and **16**. You'll find them on page 10.⁹ And just recently, I found values for C_c and C_d in the building code for Richardson, Texas.¹⁰

Those interested in a discussion of the "cross-flow principle" may want to look at graphs of theoretical and experimental values of C_c and C_d on page 3–11 of *Fluid-Dynamic Drag*, 1965 edition.⁵ As Hoerner's book illustrates, some of the earliest experimental verifi-

cation for values of C_c and C_d was performed in 1919 by Alexandre Gustave Eiffel—a well-known engineer and aerodynamic researcher.⁵ By the way, his tower is still standing!

Wind loads on Yagis with horizontal elements

Now that we know how to find cross-wind and drag forces for round tubes, we can find the total force loading of a Yagi. **Figure 5** depicts a simple Yagi with two horizontally mounted elements that we will use as an example. As the antenna is held at various angular positions relative to the wind, the loading gets somewhat complex. However, there are four positions that are easy to evaluate. We'll look at two positions where the elements are broad side to the wind and two where the boom is broadside to the wind, as shown in **Figures 6A** through **6D**. In the first two cases, only the elements experience the resulting forces; in the latter two cases, only the boom experiences the resulting force.

For angles between these four positions, the force loading becomes more complicated. **Figures 7A** through **7D** illustrate these cases. Two things become apparent. First, all drag forces are in the same direction. Second, the boom and element cross-wind forces are opposite each other because the two members are mounted at 90 degree angles. It's easy to see why it is desirable to have the dynamic wind force broken into components which are in line and perpendicular to the wind. Doing so makes determining the net force a relatively easy matter using the force sign convention shown in **Figure 8**.

Because the elements are perpendicular to the boom, we use a slightly different set of equations to find the element cross-wind and drag forces. First, **Equations 11** and **12** are renamed to apply for the boom that results in **Equations 17** and **18**. **Equations 19** and **20** are for the elements, and account for the 90 degrees between the boom and the elements. Because elements are parallel to each other, their projected areas can be summed to A_e , while the boom projected area is A_b .

$$F_{cb} = 1.2 A_b P \sin^2(\text{ang}) \cos(\text{ang}) \quad (17)$$

$$F_{db} = 1.2 A_b P \sin^3(\text{ang}) \quad (18)$$

where:

F_{cb} = cross-wind force for round boom
 F_{db} = drag force for round boom
 A_b = boom projected area
 ang = angle of attack of boom relative to wind (degrees)

$$F_{ce} = 1.2 A_e P \cos^2(\text{ang}) \sin(\text{ang}) \quad (19)$$

$$F_{de} = 1.2 A_e P \cos^3(\text{ang}) \quad (20)$$

where:

F_{ce} = cross-wind force for round elements

F_{de} = drag force for round elements

A_e = total element projected area

Let's start by concentrating on the case in **Figure 7A**. The cross-wind forces for the boom and elements are summed based on the sign convention in **Figure 8** to give the net cross-wind force.

$$F_{cn} = F_{cb} - F_{ce} \quad (21)$$

where:

F_{cn} = net cross wind force (lbs)

The substitution of **Equations 17 and 19** into **Equation 21** produces a single equation for the net cross-wind force due to the boom and elements.

$$F_{cn} = 1.2 P [A_b \sin^2(\text{ang}) \cos(\text{ang}) - A_e \cos^2(\text{ang}) \sin(\text{ang})] \quad (22)$$

Similarly, we sum drag forces for the elements and boom to yield the net drag force.

$$F_{dn} = F_{db} + F_{de} \quad (23)$$

where:

F_{dn} = net drag force (lbs)

Substituting **Equations 18 and 20** into **Equation 23** gives a single equation for the net drag force due to the boom and elements.

$$F_{dn} = 1.2 P [A_b \sin^3(\text{ang}) + A_e \cos^3(\text{ang})] \quad (24)$$

Some examples

Now that we've developed equations to describe the net cross-wind and net drag force, we'll consider several examples. First, let's look at a rather large Yagi consisting of horizontal elements with a boom area equal to that of the elements, $A_b = A_e$. To add some sense of magnitude, we'll set the areas at 10 square feet and calculate loads at an exact wind speed of 70 MPH. From **Table 2**, this puts $P = 12.5$ psf. The boom cross-wind force, element cross-wind force, and net cross-wind force for these values are shown in **Graph 2**. The boom, element, and net drag forces are shown in **Graph**

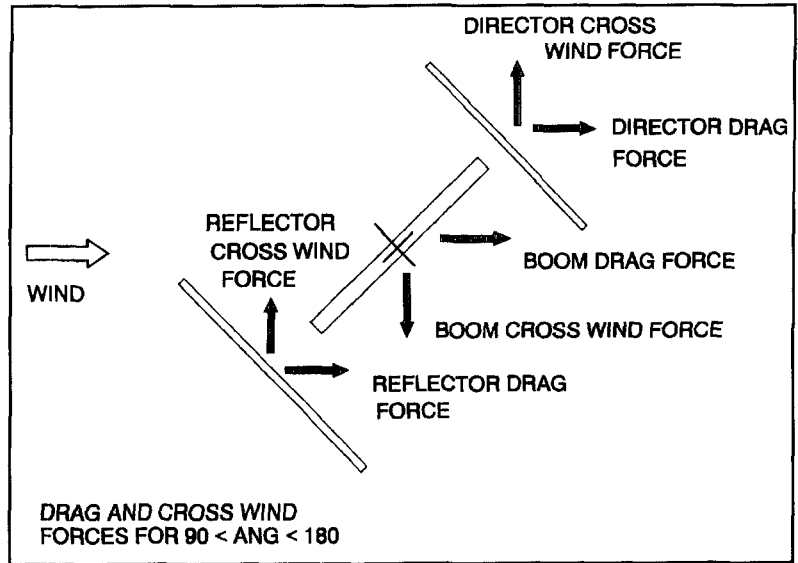


Figure 7B. Drag and cross-wind forces for $90 < \text{ang} < 180$.

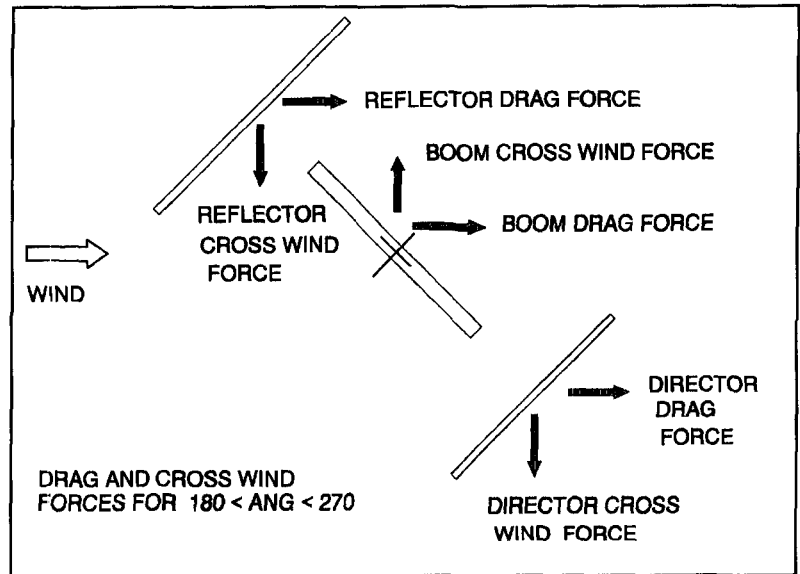


Figure 7C. Drag and cross-wind forces for $180 < \text{ang} < 270$.

3. Graph 2 indicates that the net cross-wind force is less the two individual cross-wind forces. This is expected, because the boom and element cross-wind forces are opposite each other, and their effects tend to cancel each other out. At the same time, **Graph 3** shows a dip in net drag force at 45 degrees.

Originally, forces were broken into components in line and perpendicular to the wind. This was done to allow the orderly summing of multiple forces. Now that we know the net cross-wind and drag forces, we must combine them to obtain the net force on the tower. This is done vectorially to provide F_{tow} , as shown in **Figure 9**. The angle w_t is the angle of the net force on the tower relative to the wind. F_{tow}

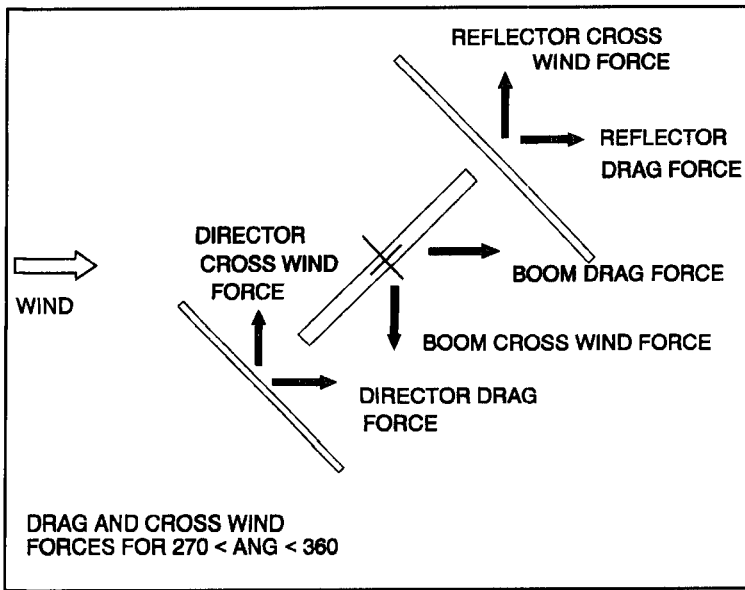


Figure 7D. Drag and cross-wind forces for 270<ang<360.

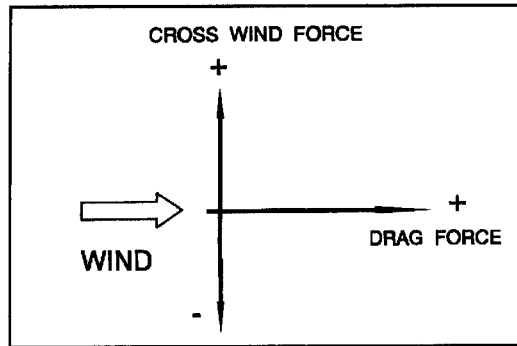


Figure 8. Force sign convention.

and w_t are found using **Equations 25 and 26**, respectively. However, because the wind can come from any direction, knowing the w_t is of little importance.

$$F_{tow} = [F_{cn}^2 + F_{dn}^2]^{1/2} \quad (25)$$

where:

F_{tow} = net force on tower (lbs)

$$w_t = \tan^{-1} \left(\frac{F_{cn}}{F_{dn}} \right) \quad (26)$$

where:

w_t = angle of F_{tow} relative to wind direction (degrees)

Graph 4 gives the net force on the tower, F_{tow} , for the case shown in **Graphs 2 and 3**. The force levels at 0 and 90 degrees are expected to be the same, as the element and boom

areas are equal. What may not have been expected is the force level at 45 degrees, where a minimum occurs. What does the method in current use say F_{tow} should be?

The “variable area” method (my name for it until a better one is suggested) says that F_{tow} is found by determining the total of the equivalent areas of the boom and elements broadside to the wind stream. The total broadside area is found using **Equation 27**. Then, **Equation 28** is used to find F_{tow} .^{11,12}

$$A_{eff} = C [A_e \cos(\text{ang}) + A_b \sin(\text{ang})] \quad (27)$$

$$F_{tow} = P A_{eff} \quad (28)$$

In **Equation 27**, the terms $A_e \cos(\text{ang})$ and $A_b \sin(\text{ang})$ stand for the magnitudes of the element and boom areas that appear broadside to the wind stream. This is because it’s assumed that the resulting wind forces are only in line with the wind, and a function of the broadside areas.¹³ As you can see, this method doesn’t account for the cross-wind forces or the true dynamic behavior of the loading. An additional misconception results from the “variable area” method. The method says there is an intermediate angle of attack between 0 and 90 degrees where F_{tow} peaks because A_{eff} has a corresponding maximum. The peak value of A_{eff} and the angle at which this occurs are found with **Equations 29 and 30**.¹³

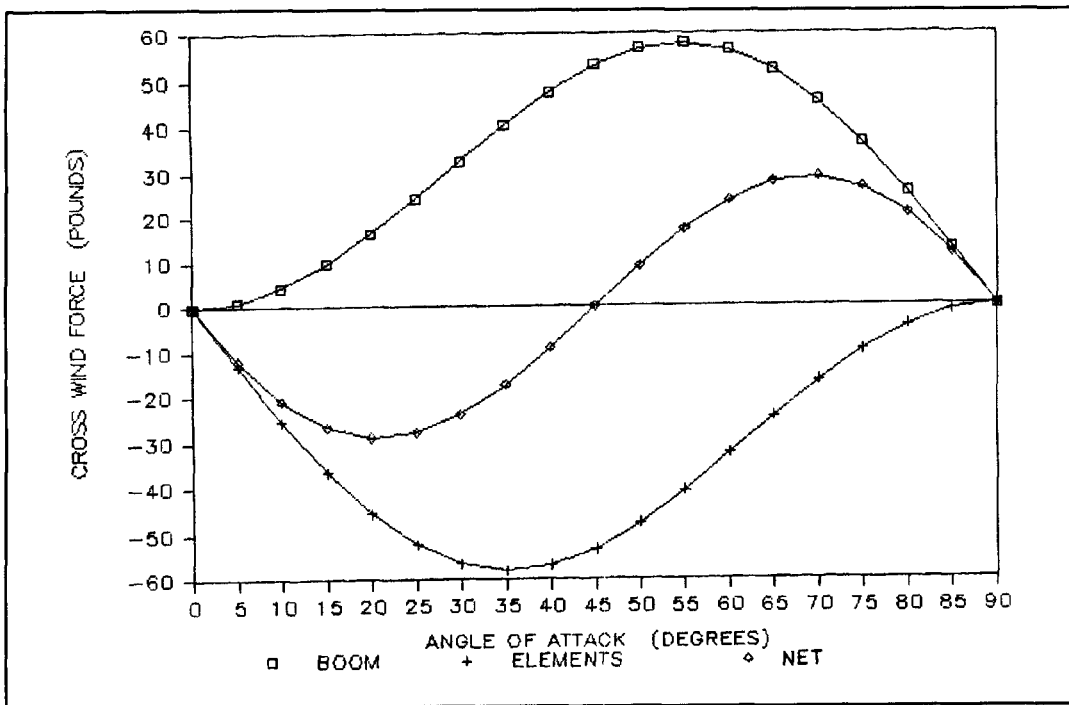
$$A_{peak} = C [A_e^2 + A_b^2]^{1/2} \quad (29)$$

$$\text{ang}_{pk} = \tan^{-1} \left(\frac{A_b}{A_e} \right) \quad (30)$$

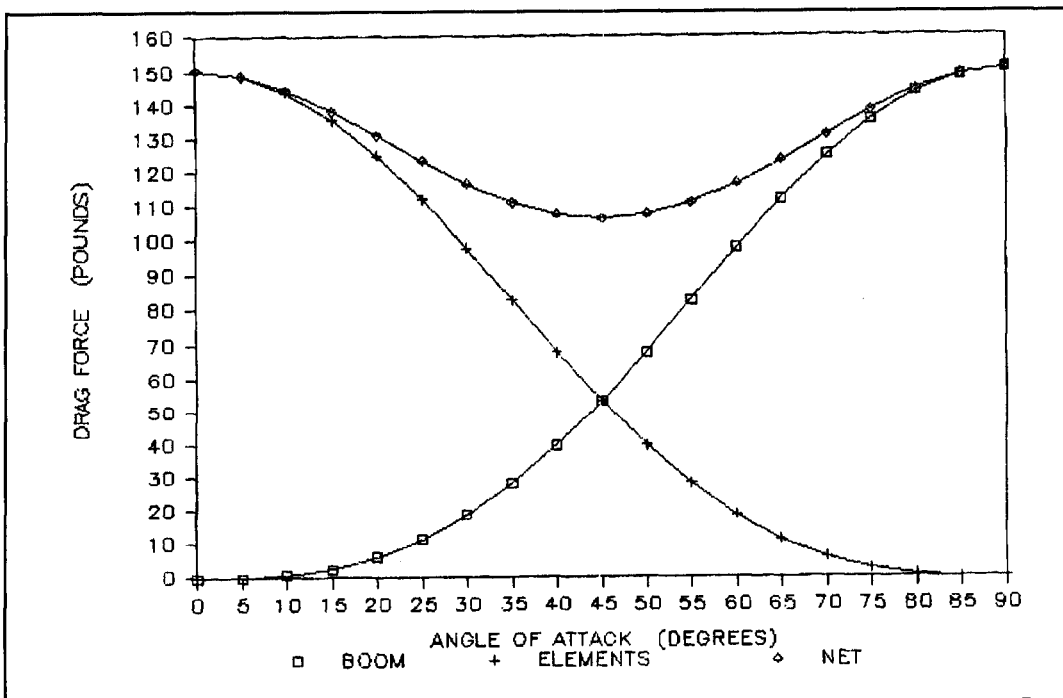
where:

A_{peak} = maximum effective area
 ang_{pk} = angle of attack at which maximum area occurs(degrees)

Using the same boom and element areas from the present example, **Graph 5** shows a comparison of F_{tow} derived by the “cross-flow principle” and the “variable area” method. Two things are evident. First, the magnitude prediction for F_{tow} is 41.4 percent too high when determined by the “variable area” method. Second, the angle of attack for maximum F_{tow} is off by 45 degrees. At first, it might seem illogical that the maximum net wind load isn’t at some angle between 0 and 90 degrees. This is easily explained. As **Graph 3** illustrates, the element drag force falls off more rapidly than the boom drag force increases, as the angle of attack increases from 0 degrees. At the same time, there isn’t much contribution to F_{tow} from the cross-wind forces, as they tend to



Graph 2. Yagi cross-wind forces. $A_b = A_e = 10$ square feet/70 mph.



Graph 3. Yagi drag forces. $A_b = A_e = 10$ square feet/70 mph.

negate each other. Also, because the net cross-wind force is relatively small when compared to the net drag force, its contribution is minimized when combined vectorially with the net drag force.

Let's look at two more examples using a wind velocity of 70 mph. The first has a boom

of 5 square feet, $A_b = 5$, and a total element area of 10 square feet, $A_e = 10$. **Graph 6** shows the element, boom, and net cross-wind forces. Again, we can see that the net cross wind is less than the largest individual contributor. **Graph 7** gives the element, boom, and net drag forces. This behavior isn't unexpected at 0 and 90

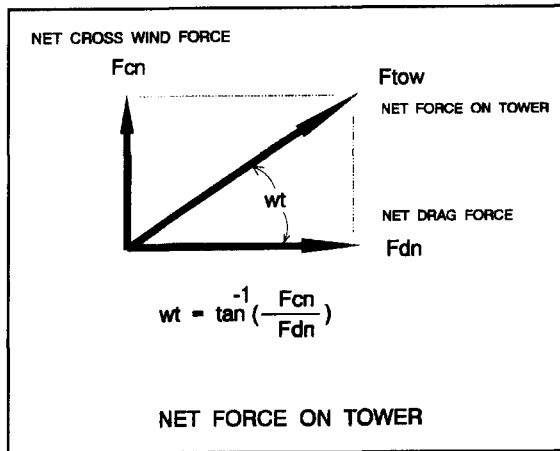


Figure 9. Net force on tower.

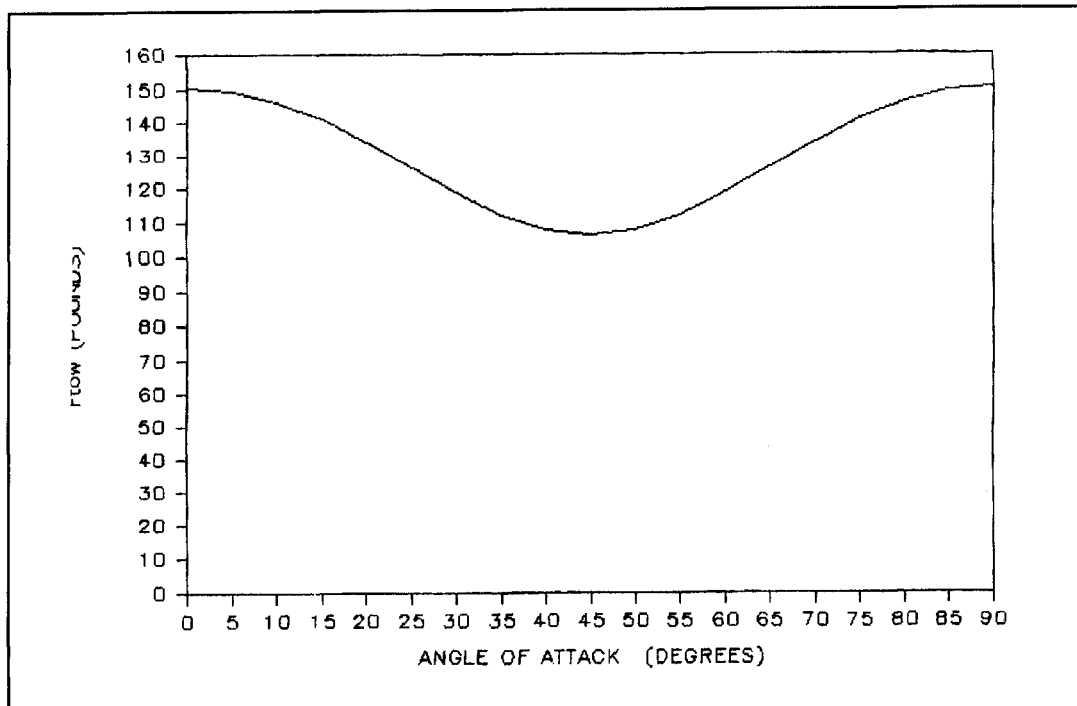
degrees due to the different boom and element areas. **Graph 8** shows the net force on the tower, F_{tow} . It also shows that the highest loading is encountered when the elements are broadside to the wind, and not at some intermediate angle of attack. In the second example, the element and boom areas are reversed, $A_b = 10$ square feet and $A_e = 5$ square feet. **Graph 9** shows the cross-wind forces, **Graph 10** the drag forces, and **Graph 11** the net force on the tower. **Graphs 8** and **11** once again indicate that there's no intermediate angle of attack where F_{tow} is greater than the force produced when the larger of the boom or total element area is broadside to the wind.

When the boom has a greater area than the

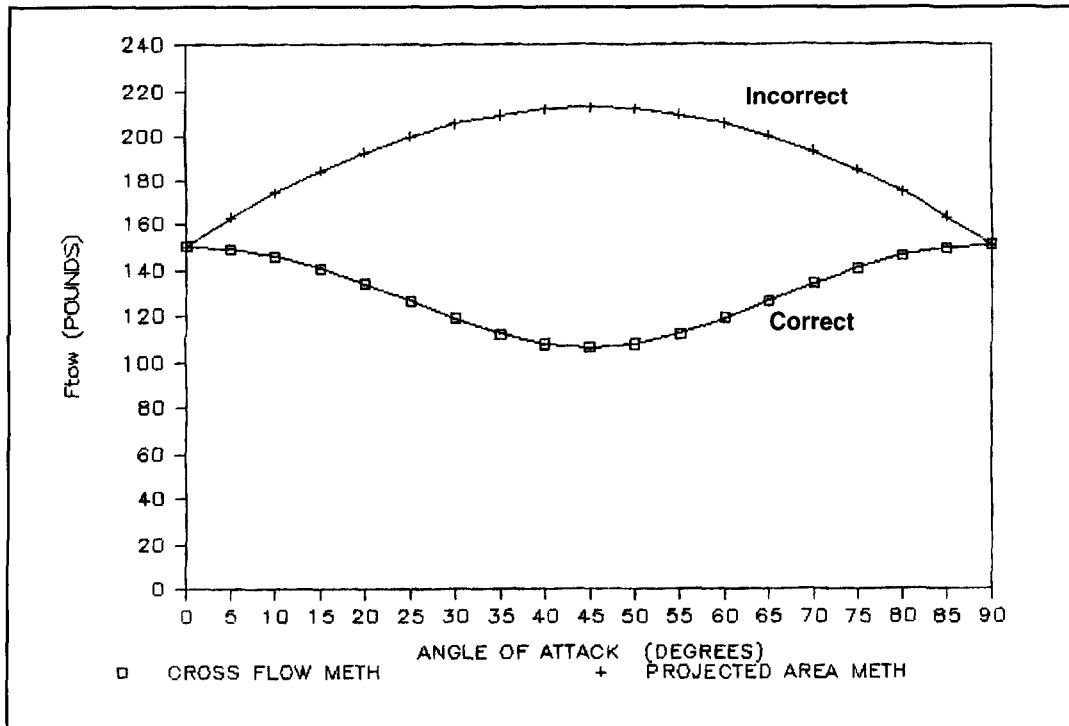
elements, the boom area determines the highest load imposed on the tower. The opposite is also true. When the elements have more total area than the boom, the element area determines the highest tower load. For the unique situation where the boom area and total element areas are the same, there are two angles of attack—0 and 90 degrees—where the same peak value of F_{tow} is developed. **Graphs 4, 8, and 11** show F_{tow} for 0 to 90 degrees. **Graphs 12, 13, and 14** show F_{tow} for 0 to 360 degrees of rotation for the same cases, respectively. These last graphs show that F_{tow} behaves as expected for a full 360 degrees of rotation.

Summary review

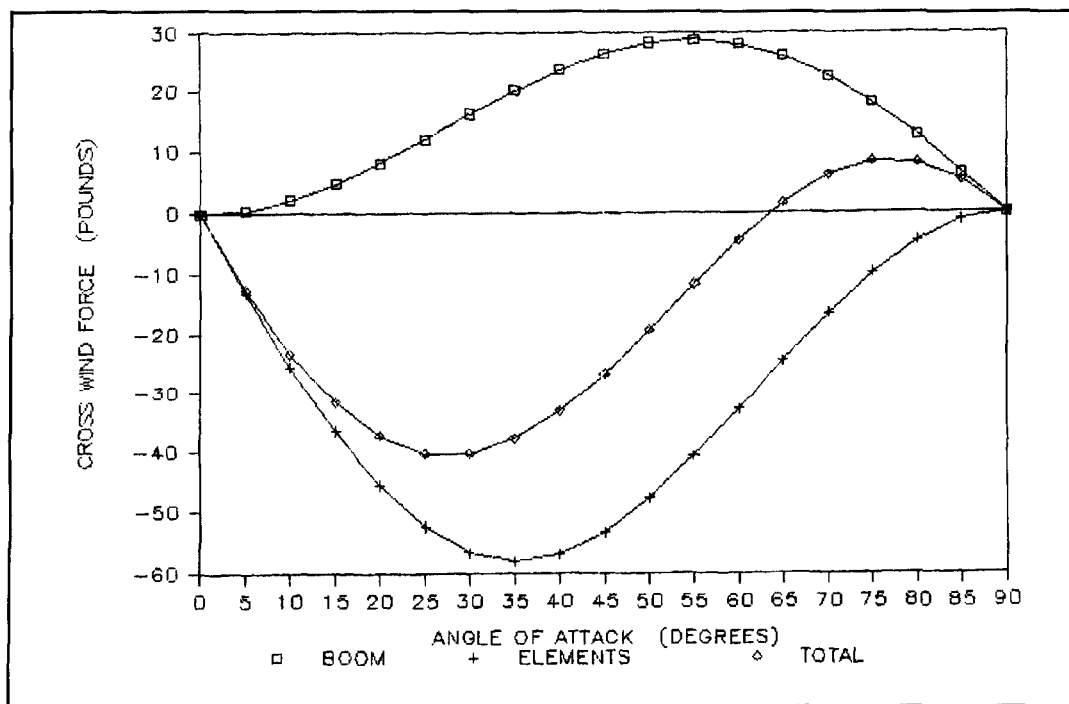
The force developed on a round tube inclined to a wind stream is perpendicular to the tube as shown in **Figure 2** and is found using **Equation 8**. In **Equation 8**, A is the projected area of the tube, which is its length times its diameter. To make it a simple matter to find the net force on a tower, forces produced by horizontally mounted elements and boom are broken into components perpendicular to the wind and in line with the wind. The forces perpendicular to the wind are the cross-wind forces found using **Equations 17** and **19**. The drag forces found by **Equations 18** and **20** are forces in line with the wind. Building on these equations, three others result that let us find the force on the tower, F_{tow} , directly. These are **Equations 22, 24, and 25**. **Equation 22** gives



Graph 4. Net force on tower, F_{tow} . $A_b = A_e = 10$ square feet/70 mph.



Graph 5. Net force on tower, F_{tow} . $A_b = A_e = 10$ square feet/70 mph.



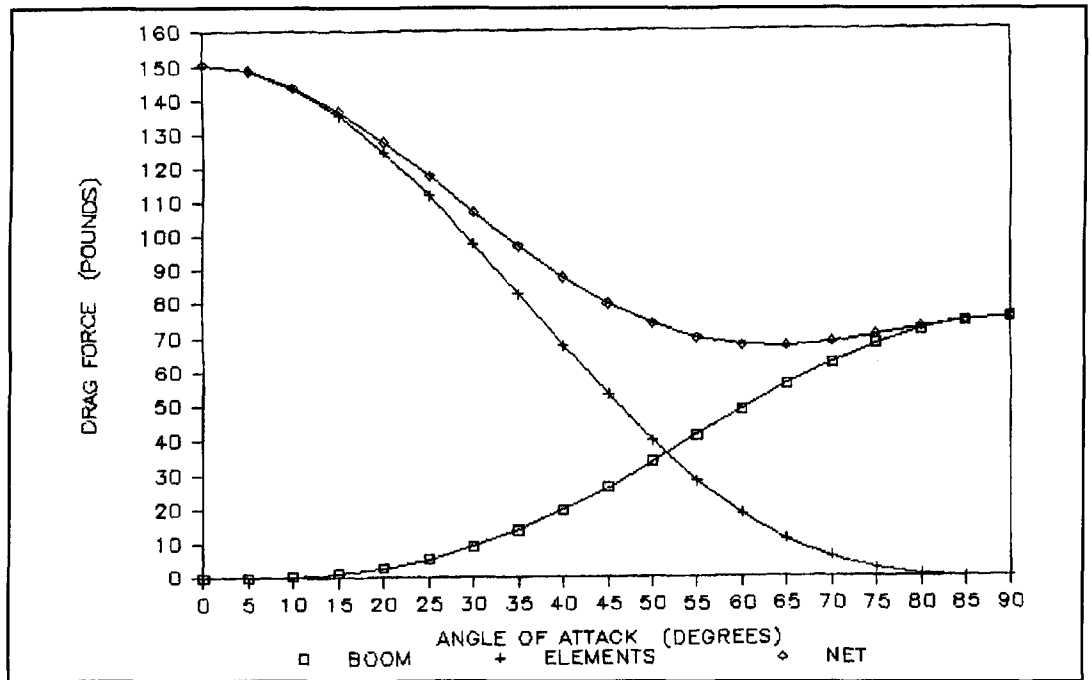
Graph 6. Yagi cross-wind forces. $A_b = 5$ square feet/ $A_e = 10$ square feet/70 mph.

the resultant cross-wind force, **Equation 24** the resultant drag force, and **Equation 25** the vector sum of the cross-wind and drag forces. F_{tow} from **Equation 25** provides the net loading on the tower.

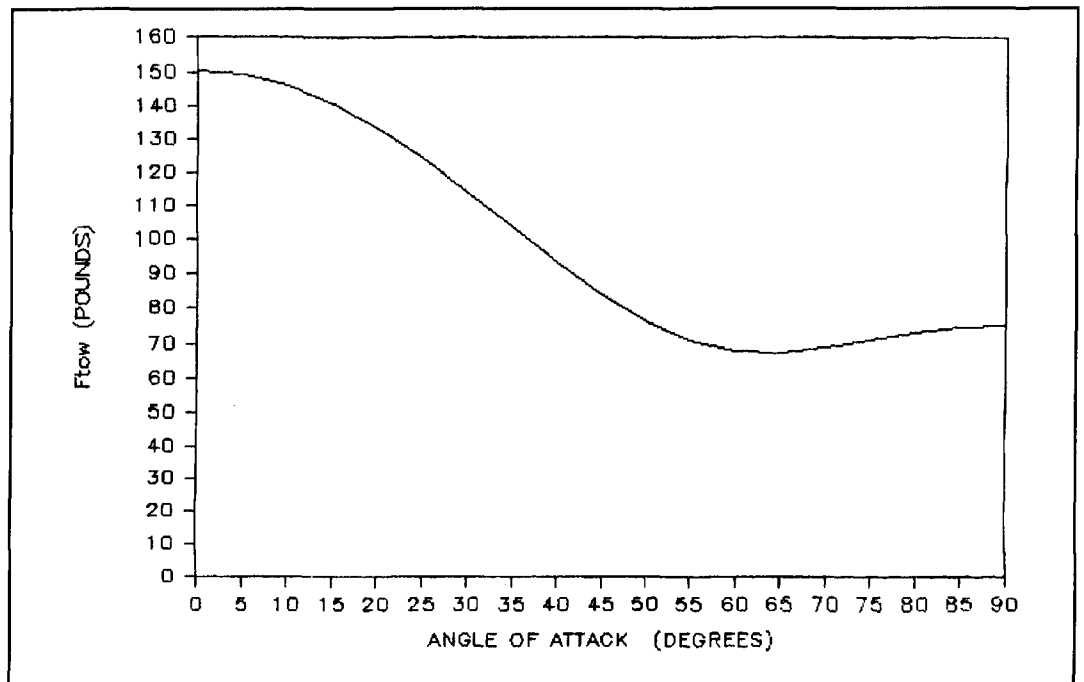
For more information on using the “cross-flow principle”, review **References 5 and 6**.

These will show the mathematical formulation and verifying experimental results. Although the use of the “cross-flow principle” to evaluate wind loads on amateur Yagis is new, the method is not. Some of the earliest experiments verifying it are over seventy years old.⁵

Reference 13, which discusses the incorrect



Graph 7. Yagi drag forces. $A_b = 5$ square feet/ $A_e = 10$ square feet/70 mph.



Graph 8. Net force on tower, F_{tow} . $A_b = 5$ square feet/ $A_e = 10$ square feet/70 mph.

method, states: "The wind force acts only in the direction of the wind..."¹³ The following simple test demonstrates that force components are also developed perpendicular to the wind stream direction. Two people take a 24 to 36-inch length of round tubing about 1 inch or more 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

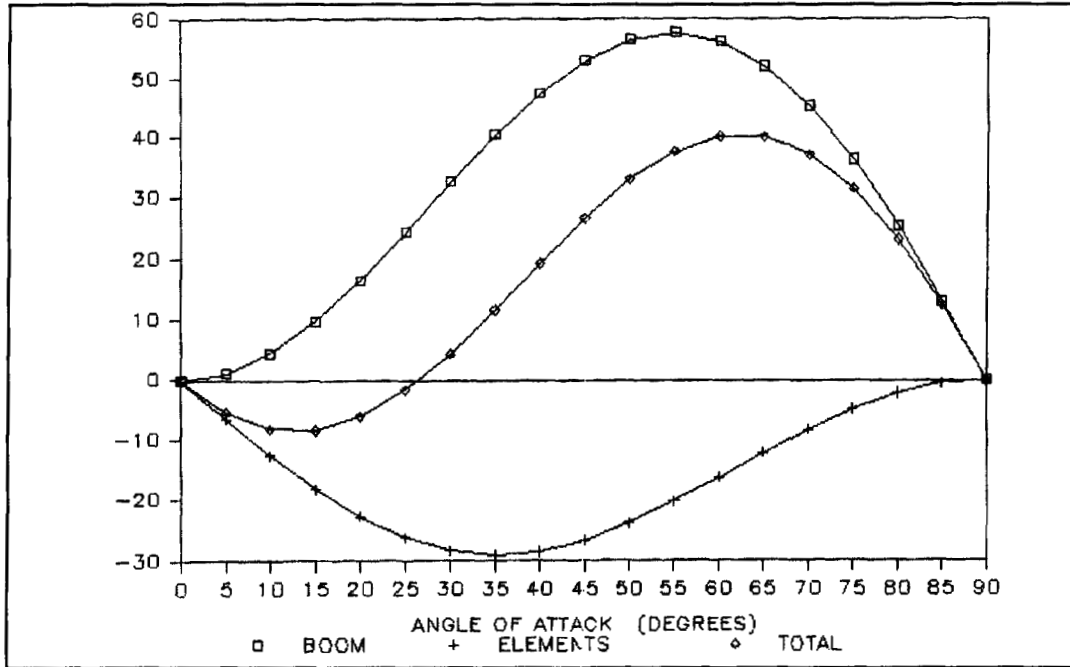
straight up and down. The tube is now perpendicular to the wind stream. With the tube broadside to the wind stream, it's easy to observe a force (drag force) in line with the wind. The passenger now rotates the tube 45 degrees, top end forward. There is not only a much smaller drag force, but also be an upward push due to the cross-wind force. At a different angle of attack, when the tube is turned 45 degrees to the wind, lower end forward, the

cross-wind force pushes downward. Using this simple test, it's possible to prove that the basic assumption of the "variable area" method is incorrect and, at the same time, verify the basic nature of the "cross-flow principle."

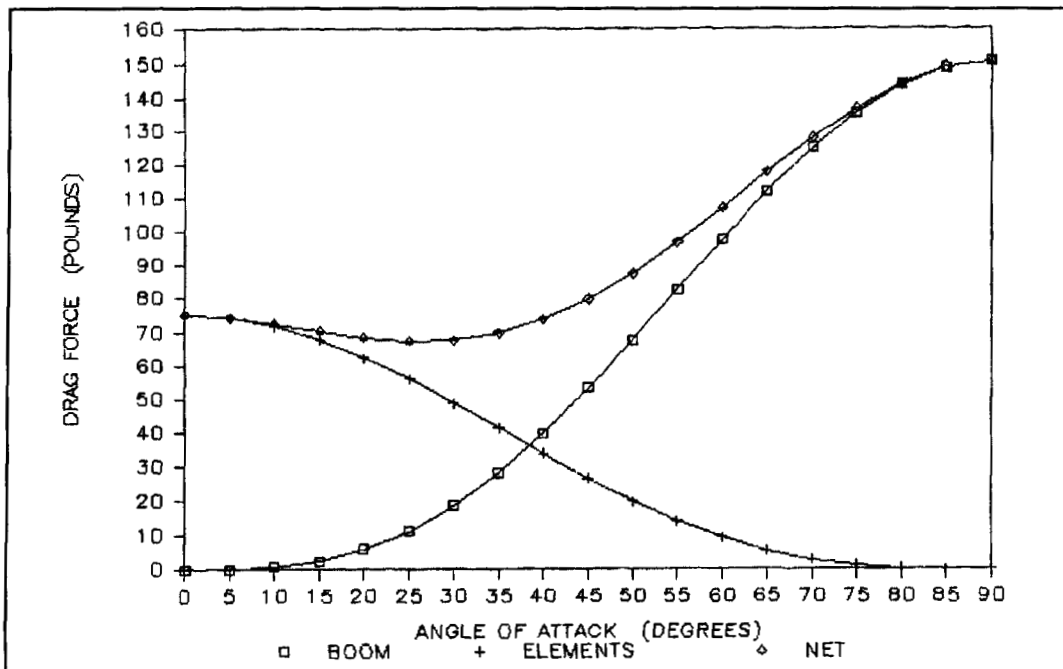
The Challenge

In the course of preparing this article, I wrote to five major amateur radio antenna manufac-

turers and asked how they determined wind loading effects for their Yagis. All but one, use the "variable area" method to derive a "surface area" (sic) for their antennas using **Equation 29**. The remaining manufacturer uses the sum of the element and boom areas. In both instances, the intent is for the stated surface area to be multiplied by the dynamic wind pressure to find the resulting wind load. Using these methods, all have overstated the parameters



Graph 9. Yagi cross-wind forces. $A_b = 10$ square feet/ $A_e = 5$ square feet/70 mph.



Graph 10. Yagi drag forces. $A_b = 10$ square feet/ $A_e = 5$ square feet/70 mph.

used to find the wind loads produced by their antennas. One manufacturer does so by taking an extremely conservative approach. The others, using the “variable area” method, also overstate the parameters used to determine the resulting wind loads—demonstrated by the example used to generate **Graph 5**.

A quick look at the surface area specifications for two very similar antennas from two different manufacturers shows that one, or both companies, is not doing what it says it does when determining the “surface area” of its antenna. Also, there is at least one manufacturer who doesn’t include the area of the boom-to-mast plate in the wind load. The manufacturer claims this plate is part of the mast. I’m not sure about the masts sold outside of Texas, but here, when you buy a mast, you don’t automatically get a boom-to-mast plate which, quite coincidentally, fits your beam. After a little investigation, it becomes apparent there is need for a standard method to find the wind loading effect of Yagis, and that it should be based on the “cross-flow principle.” This standard should be used by all manufacturers, because it would allow the consumer to make an accurate comparison of antenna specifications. Also, having a technically correct standard would allow for valid structural evaluations of Yagi installations. The challenge? To establish a technically correct standard.

Establishing the standard

Most Yagi antennas are made from round tubes and flat plates, although some designs have other shapes—like molded element insulators or formed mounting brackets. The challenge is to establish a standard that recognizes these different shapes. A few simplifications are required to make the standard straightforward. Otherwise, it will get bogged-down in the intricacies of defining the wind loading for all possible shapes used for element mounting and boom-to-mast plates.

The first simplification is that all parts of a Yagi which aren’t round must be treated as flat areas. Because Yagis mostly consist of round tubes and flat plates, any error induced by treating a molded part as a flat plate will be extremely small. The “cross-flow principle” applies to all shapes—not just round tubes. Therefore, the equations for the cross wind and drag forces for round tubes apply for flat plates with only a change in the drag coefficient. Although equations derived here have concentrated on round tubes with a drag coefficient of 1.2, the same equations apply for flat plates with a drag coefficient of 2.0. All that’s needed is an inventory of the round and flat areas. Round areas have a 1.2 drag coefficient applied

and flat areas have a 2.0 coefficient. This means that flat plates and round tubes can be combined to provide useful parameters that describe wind loading effects. The required parameters would describe instances when the elements are broadside to the wind and when the boom is broadside to the wind.

This can be handled by defining two load factors, E and B. E would account for wind loading effects with the elements broadside to the wind, and B when the boom is broadside to the wind. Factors E and B would have areas and respective drag coefficients embedded so a simple equation could be used to find the worst-case wind load.

Because there is no intermediate angle of attack between 0 and 90 degrees where F_{low} peaks, the larger of E and B would be used to find the worst-case wind load. To do so, one would use **Equation 31** with the larger of E and B.

$$F_{max} = (E \text{ or } B) P \quad (31)$$

where:

F_{max} = maximum value of F_{low}
 E = element load factor (square feet)
 B = boom load factor (square feet)
 (use the larger of E and B)

Equation 32 shows how to determine the boom load factor with the boom broadside to the wind. This equation takes into account the projected areas of the round and flat parts making up the antenna. It also considers the boom-to-element connection viewed “boom broadside,” as shown in **Figure 10**.

$$B = 1.2 A_b + 2.0 A_{bp} + 2.0 A_{et} \quad (32)$$

where:

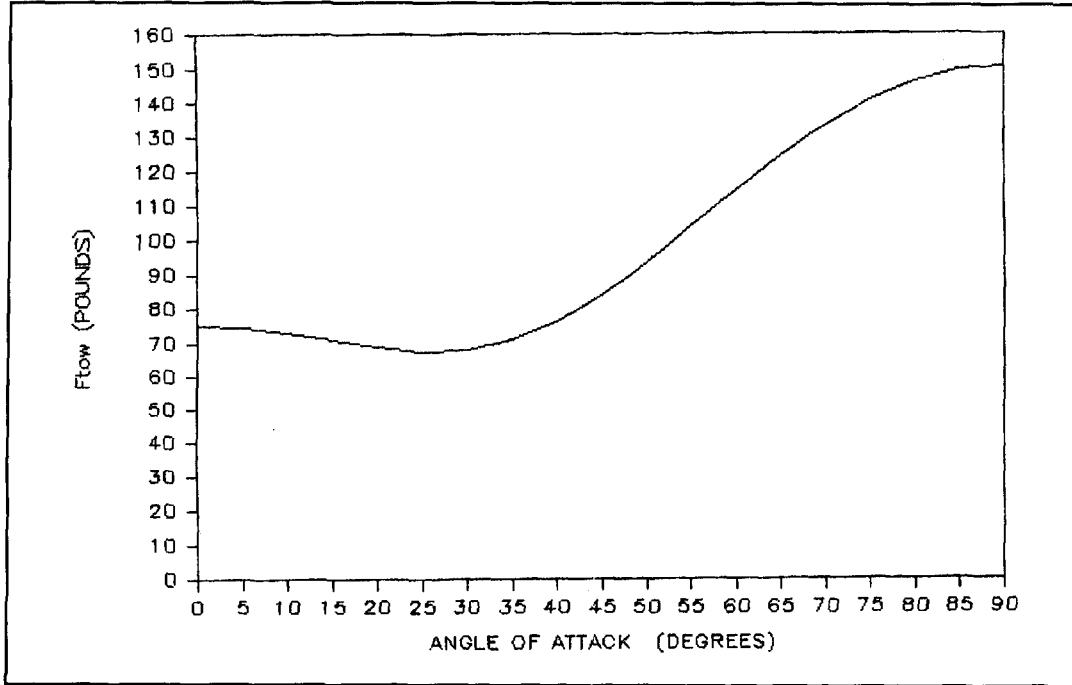
B = boom load factor (square feet)
 A_b = exposed boom area (square feet)
 A_{bp} = boom flat plate area (square feet)
 A_{et} = total element end area (square feet)

In the case of the boom-to-mast plate shown in **Figure 10**, the broadside area of the plate, A_{bp}, is used while the area of the boom behind it is not, because the plate shields the boom. Although the actual drag coefficient in this area is a little less than 2.0, there’s very little error introduced by following this methodology. More importantly, this methodology is easy to follow and understand. The area highlighted by the circle and enlarged in **Figure 10** shows the recommended way to handle the element mounting plate and the area presented to the

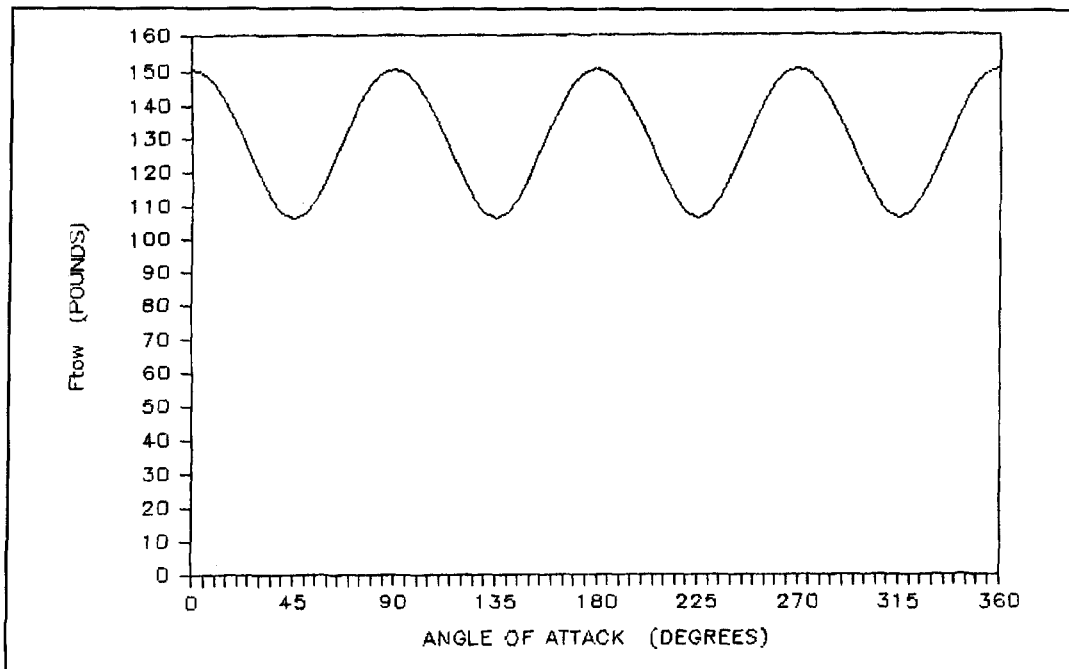
wind by the end of the element. The area of the box is taken as a flat plate area. This is done for each element and the areas summed to obtain A_{et} . The exposed boom area, A_b , is then the area of the boom not shielded by the boom-to-mast plate and not enclosed in the "boxes."

Determination of the element load factor follows a similar pattern as shown in **Equation 33** and **Figure 11**. The term A_{pe} denotes any flat

plate that may be attached to the boom or elements, is broadside to the wind, and is not used to mount the elements. Only in a very rare instance would there be such a plate. The boom end cross-section area is used to account for the boom end. For example, if the boom has a 2 inch diameter at its end, A_{be} would be the area of a 2-inch diameter circle. Admittedly, for HF Yagis the boom end area will be very small



Graph 11. Net force on tower, F_{tow} . $A_b = 10$ square feet/ $A_e = 5$ square feet/70 mph.



Graph 12. Net force on tower, F_{tow} . $A_b = A_e = 10$ square feet/70 mph.

compared to the area of the elements, but it may be a significant contributor to E for UHF and higher frequency Yagis.

$$E = 1.2 A_{el} + 2.0 A_{pe} + 2.0 A_{be} + 2.0 A_{em} \quad (33)$$

where:

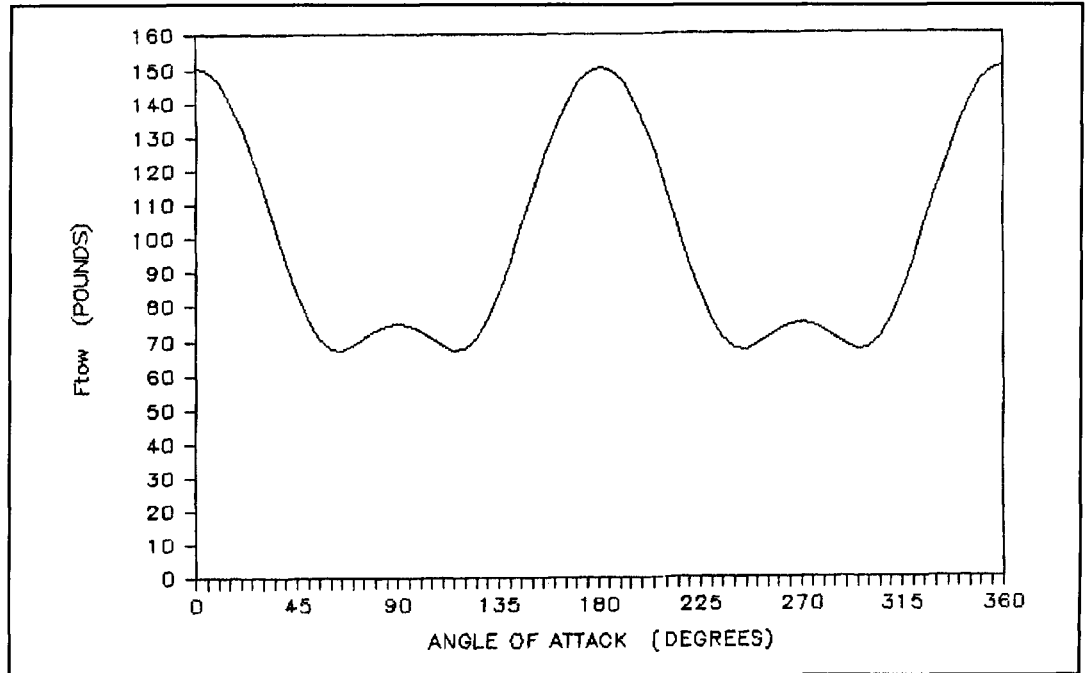
E = element load factor (square feet)
 A_{el} = total element area (square feet)

A_{pe} = flat plate area (square feet)

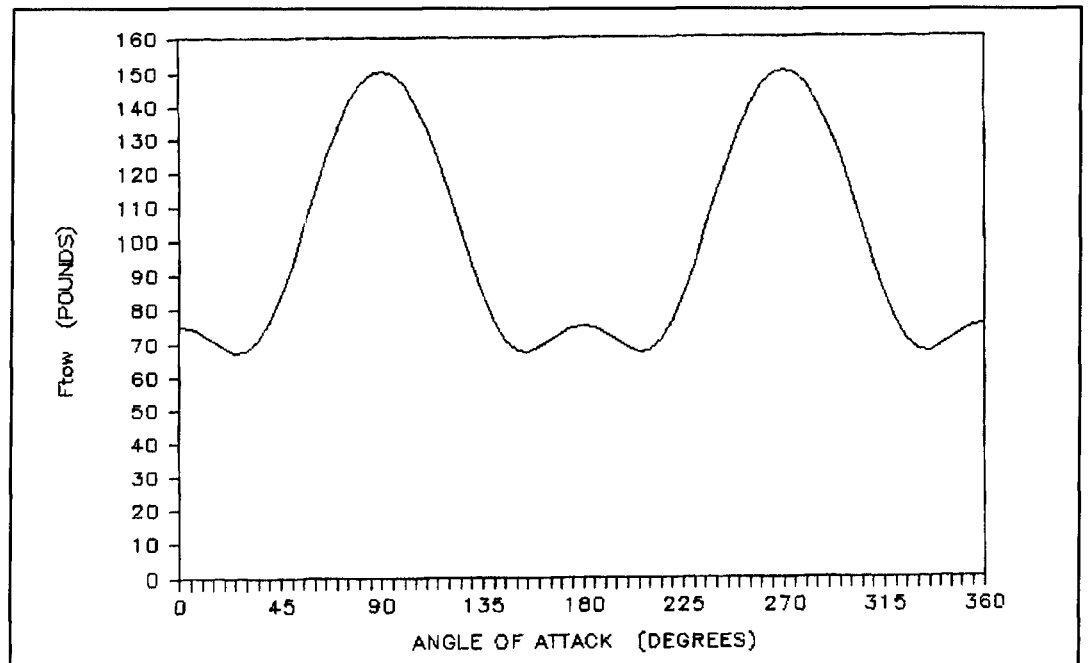
A_{be} = boom end area (square feet)

A_{em} = total element mounting area (square feet)

Most Yagis use relatively thin plates to mount elements to the boom. However, there are cases where molded insulators, which generally have considerable broadside area, are used to isolate the elements from the boom. To



Graph 13. Net force placed on tower, Ftow. Ab = 5 square feet/Ae = 10 square feet/70 mph.



Graph 14. Net force on tower, Ftow. Ab = 10 square feet/Ae = 5 square feet/70 mph.

account for the contribution of mounting plates or insulators, I recommend the method shown in **Figure 11**. The enlarged area shows the mounting plates and the mating element length enclosed by a “box.” Find the area of the “box” for each element and then sum the areas to obtain A_{em} in **Equation 33**. This area is also treated as a flat plate area. Obtain A_{el} by finding the broadside area of each element, but not using that within each “box,” and summing them to give the total element area. Again, there are some minor errors that occur when using the “boxes,” but the contributions of the enclosed items are accounted for using a simple, understandable, methodology.

Finding the maximum force

Follow the methodology for the boom-broadside case using **Equation 32** and the element-broadside case using **Equation 33**. After finding the load factors, select the larger of them to find the maximum force. Once you’ve found B and E (either by “cranking” the numbers or hopefully, in the future, obtaining them from a manufacturer’s specification sheet), use the larger in **Equation 31**.

The value of P depends on the wind conditions for your installation and the antenna height. There are several ways to find out the value of wind velocity and set of modifying factors for gusts and tower height you should use. EIA Standard EIA/TIA-222-E is one of the best sources of information.⁹ In this standard, wind conditions are now specified by county rather than in map form. Also given are equations to find allowances for antenna height and gusts. You can also check to see what is required by the building code for your area. Building codes generally state conditions deemed to be worst case for your area. In areas of very high wind, you might be well advised to consult local government agencies, like the weather bureau, to see what information they can provide.

Yagi wind loading standard

The standard recommended here is very simple. Use **Equations 32** and **33** to find boom and element load factors. Manufacturers should list these two parameters in their specification sheet. Manufactures should **not** state wind forces at a specific wind speed—as this doesn’t account for local conditions, gusts, antenna height, and other conditions that control the dynamic wind pressure. Once you know the two load factors, its simple to use **Equation 31** to find the maximum tower load based on the dynamic wind pressure unique to your location.

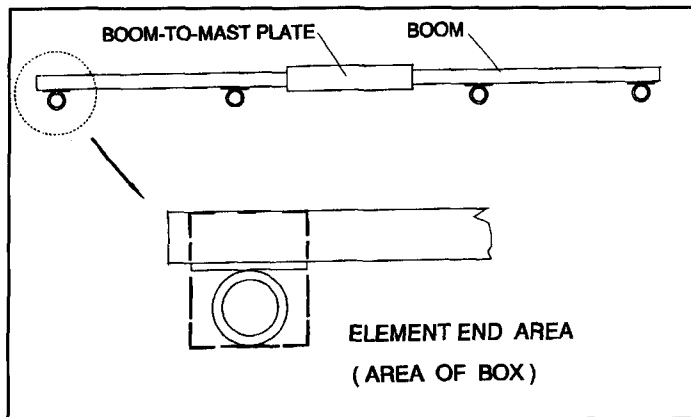


Figure 10. Boom broadside.

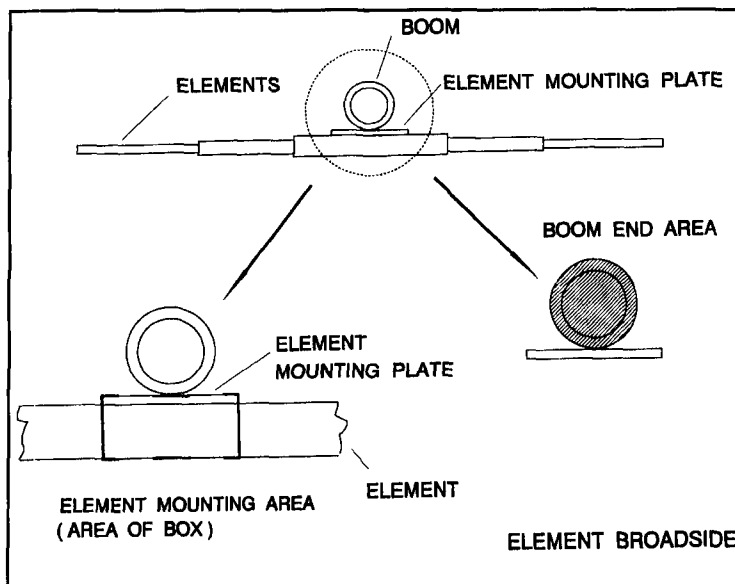


Figure 11. Element broadside.

When comparing different antenna models, you would compare the E and B load factors. Also, by knowing both factors, you would know which way to point your beam to minimize the wind loading during storms.

Summary

For at least the past 25 years, an incorrect method has been used to determine Yagi wind loads. The “variable area” method currently employed by manufacturers, discussed by numerous authors, and used in amateur mechanical design software has no basis in science, experimental evidence, or fact. Using the “cross-flow principle,” forces due to wind can be predicted correctly. The “cross-flow principle” has been known and used for many years by mechanical, civil, and aeronautical engineers. There is an abundance of experimental

verification regarding this principle—some of which dates back to 1919.⁵

Use of the “cross-flow principle” shows that larger wind loads result with the boom broadside to the wind and the elements broadside to the wind. The larger of the two presents the worst case. There is no intermediate position that produces a higher load; quite the contrary is true. There is a minimum between the two broadside positions. The only information needed to determine the wind loads at both broadside orientations are the E and B load factors. I recommend that manufacturers and individuals use these two factors when stating the wind loading aspects of their designs.

I hope that antenna manufacturers will adopt the methodology discussed here, and that the E and B load factors will be used in product specification sheets to correctly describe antenna wind loads for Yagis with horizontally mounted elements. I invite those interested in this topic to review the cited references. I’d like to state once again that the “cross flow principle” isn’t new. It just hasn’t been used, while the current method has gone unquestioned for

many years. Needless to say, I look forward to vigorous discussions on this topic. ■

REFERENCES

1. S.F. Hoerner, *Fluid-Dynamic Drag*, published by author, Bricktown, New Jersey, 1965, page 1–10.
2. C.J. Richards, *Mechanical Engineering in Radar and Communications*, Van Nostrand Reinhold Company, London, England, 1969, page 163.
3. EIA/TIA-222-E, *Structural Standards for Steel Antenna Towers and Antenna Supporting Structures*, Electronics Industries Association, Washington, D.C., 1991, pages 4–5.
5. S.F. Hoerner, *Fluid-Dynamic Drag*, published by author, Bricktown, New Jersey, 1965, page 3–11.
6. C.J. Richards, *Mechanical Engineering in Radar and Communications*, Van Nostrand Reinhold Company, London, England, 1969, page 162.
7. E.A. Avallone and T.B. Baumeister, *Mark's Standard Handbook for Mechanical Engineers*, 9th edition, McGraw-Hill Book Company, New York, New York, page 11–78.
8. C.J. Richards, *Mechanical Engineering in Radar and Communications*, Van Nostrand Reinhold Company, London, England, 1969, page 165.
9. EIA/TIA-222-E, *Structural Standards for Steel Antenna Towers and Antenna Supporting Structures*, Electronics Industries Association, Washington, D.C., 1991, page 10.
10. *Standard Building Code/1988*, Richardson, Texas, page 227.
11. Roger Cox, WB0DGF, *Determining Antenna Surface Area*, Telex-Hygain, Minneapolis, Minnesota, 1978, page 4.
12. David B. Leeson, W6QHS, *Physical Design of Yagi Antennas*, American Radio Relay League, Newington, Connecticut, 1992, page 6–3.
13. David B. Leeson, W6QHS, *Physical Design of Yagi Antennas*, American Radio Relay League, Newington, Connecticut, 1992, page 6–1.

PRODUCT INFORMATION

Radio Spectrum Explorer™ for Windows

Advanced Computer Controls introduces Radio Spectrum Explorer software for Windows. Explorer is a new user interface for your VHF and UHF communications receiver.

Explorer offers different perspectives on operating the radio through several windows—Frequency, Description, SuperMemory™, Spectrum Chart, and Map windows. The *Frequency window* most closely resembles a radio’s conventional front panel, although it is enhanced to make it easier to “tune around.” Other windows provide new perspectives on spectrum exploration.

The *Description window* displays information about the channel selected. It shows nationwide FCC allocation information, along with easily edited local information panels to show the radio user description, detail, locations, call sign, and service.

With Explorer, memories are organized in a hierarchy called *SuperMemories*. Folders contain frequencies, groups (like trunking groups), tuning ranges, and more folders. The multi-level folder hierarchy uses descriptive names and icons resembling file management on modern computers.

The *Map window* lets you tune using geographical maps of a specific area. Maps are obtainable from a program like *Street Atlas USA* or *AutoMap*, or by scanning printed maps into the computer.

The *Spectrum Chart window* is analogous to a printed spectrum wall chart, but appears on-screen, instead.

Radio Spectrum Explorer is fully Windows compliant, and its operation resembles other popular Windows programs. It has its own Help file which serves as a thorough on-line hypertext manual. A status line in each window continually feeds back information about how to use the program.

System requirements include a personal computer using a 386SX-25 or higher processor running Windows 3.1, and a receiver with a computer port and level converter. Several radios *without* computer ports will also be supported in the future with hardware options.

ACC produces microcomputer based control systems and software for amateur, commercial, and government radio users. For more information, contact Advanced Computer Controls, Inc., 2356 Walsh Avenue, Santa Clara, California 95051, or call 1-408-727-3330.

THE 1993 PERSEIDS: A METEOR STORM?

*Chances are good we'll see one
this August*

A brief introduction^{1,2}

There has been interest in using the ionization to the E-layer of the ionosphere caused by meteor vaporization, or burn-up, for propagation of amateur radio station signals since before World War II. However, interest in this form of propagation didn't really take off until Ed Tilton, W1HDQ, the then editor of *QST's* "The World Above 50 Mc." column, encouraged his readers to listen for something spectacular on the newly acquired 6-meter band during the predicted meteor storm caused by the Giacobinid-Zinner Comet.**

Those on the band the night of October 9, 1946 weren't disappointed. In his column,³ Tilton stated that the event "...produced such bursts, occurring over a wider area and for a longer period of time than ever before in the history of VHF work." Reports sent to Tilton from numerous amateurs indicated that the band opened around 8:30 p.m. He described the sound they heard as a "peculiar rumble" noticeable on all signals coming from beyond the horizon. Tilton wrote that reports he received indicated that signals from as close as 200 miles and as far as 1,200 miles were coming in all at once, causing QRM that "...had never before been experienced on the normally wide open space (of the 6-meter ham band)."

Very short skip contacts were possible because of the unusually low height of the ionization of the E-layer. The intensity of the ionization, allowed for contacts of long duration

(for meteor scatter work). Ed Ladd, W2IDZ, one of the participants that night described the QSOs as "finished." Not only were signal reports and other necessities exchanged, but unnecessary items such as equipment and weather reports were being swapped, as well. Additionally, Ladd reported that after QSOs were completed, operators could hear the other stations for two to three minutes following the contact. He described the sound of the contacts as fluttery, with intermittent drop outs of signal strength as the ionization disappeared and reappeared with each meteor vaporization.

The propagation lasted for around three hours. Most of Tilton's reports came from the eastern and the midwestern portions of the United States. For the first time in the history of ham radio, meteor scatter was put to genuine use for signal propagation.

This shower piqued the interest of hams and others about this form of propagation. John Stewart, and others, reported⁴ on the results of a U.S. Army radar experiment conducted during the shower that showed echoes on the 106-MHz "early warning" radar. Conversely, the article also reported that there was no evidence of echoes on radars operating on 600, 1,200, 3,000 or 10,000 MHz.

From this 1946 beginning, meteor shower-induced propagation became part of the VHF operator's communications repertoire. However it wasn't until 1953 that a 2-meter contact via this form of propagation took place.

In June 1953, Paul Wilson, W4HHK, and Ross, W4AO, were in contact with each other via a tropo path. After the tropo path fell apart, Paul continued to hear bursts of signals. Ross advised Paul that he was hearing meteor bursts.

¹Mr. Lynch is the VHF editor for our sister publication, CQ magazine.

²Meteor showers, and the more intense "storms" are caused by the Earth's collision with debris expelled by a comet as their two orbits intersect or come within close proximity of each other.



Photo A. Comet Swift-Tuttle is the bright object in the lower right of the photo shot by Dennis di Cicco on November 1, 1992. Notice the gas tail extending more than $\frac{1}{4}^\circ$ from the bright coma. Di Cicco made this 200 second exposure from a suburb 20 miles west of Boston, Massachusetts, with an SBIG ST-6 CCD camera attached to a 4-inch f/5 refractor. Photo courtesy Dennis di Cicco and *Sky and Telescope*.

Within a few days of this contact, Paul got a letter from Ralph "Tommy" Thomas, W2UK, asking to set up schedules for a possible 2-meter contact via any mode of propagation. Paul's answer indicated a desire to try to work him via meteor scatter.

Over the next several months, schedules were set without success. Then, on the morning of October 22nd, it all fell together when Tommy copied more than 2 minutes of transmission from Paul and he, in turn, was able to copy Tommy's confirmation and signal report. With that exchange, they had the first complete 2-meter QSO via meteor scatter.

Interest grew over the years. In April 1957, Walt Bain, W4LTU, wrote what has become the classic commentary on meteor scatter work.⁵ Among the topics covered were what to expect when attempting a contact and suggested sequencing lengths. Because CW was used as the mode of communication, there was support for five minute lengths of sequencing. However, today's almost exclusive use of SSB makes anything but fifteen second sequencing archaic. (Please see the sidebar for information on how to complete meteor contacts.)

Interest continued to grow as an increasing number of operators tried out the mode. Remember that in those days the operators didn't have sophisticated equipment with fancy filtering and digital readout. With analog readout, calibration as close as 5 kHz was considered excellent. Imagine trying to find a burst when you're not at all sure where to look. Also, the

primary mode of communication was high-speed CW (often around 40 wpm). It wasn't a method of propagation for the faint of heart. Nevertheless, encouraged by intensified publicity, more operators tried out the mode.

In November 1965, a number of operators experienced excellent results from the *Leonids* shower. Continued good press prompted more interest. Encouraged by predictions in the scientific journals,⁶ hams stood by for what they thought might possibly be a better than average night for the 1966 *Leonids* shower.

Indeed, it was! The headline for Sam Harris, W1FZJ's column⁷ was "November Leonids—Shower of a Lifetime." Sam recounted, "Hundreds of contacts were made by calling CQ, or by breaking stations when their skeds were completed, as most were in the first minute or two of prearranged calls."

Reports of visual observations were sent to *Sky and Telescope* from all over the country. Among the reports was one from Shelby Ennis, W4WNH (now W8WN). From his letter Shelby was quoted as saying, "For us in Kentucky, the 1966 Leonids will be rated much better as a 'radio' shower than as a 'visual' shower due at least in part to the very sharp peak coming after dawn." However, in areas where dawn hadn't come, particularly in the west, the display was awesome. Reports of 2,000 meteors per minute weren't uncommon. It was a night (or early morning) to remember for amateur radio operators and amateur astronomers, alike.

Meteor induced propagation today

Interest in meteor shower-induced propagation continues today. Because the *Perseids* is a relatively long running shower (meteors associated with the *Perseids* appear as early as two weeks before the peak) and it appears during the summer, it's the most popular of these type of events.

Intriguingly, interest in this shower has been increasing within the astronomical circles—in part because of reports of amateur radio operators' activities. In particular, Dr. Brian G. Marsden, who predicted the return of the Comet Swift-Tuttle (see **Photo A**), acknowledged the contributions made by hams as a swaying factor when he decided to commit to refining the predicted return date⁸ of the comet. And Mr. Joe Rao, an amateur astronomer working on estimating the potential for a storm and who's prediction is the basis for this article, told me in a telephone call that it was partly because of my article in August 1992 *Sky and Telescope* (on "listening to" the *Perseids*) that he continued his research into the possibility of a storm.

Indeed, since 1988, the growth of a visual observation of a new peak in meteor activity has been reported. This visual peak has been confirmed by amateurs' on-the-air reports for the last two showers. In my column,⁹ I reported that the 1991 peak, or spike, of the shower appeared to have occurred around 1600 UTC

on August 12th. Again, in my column,¹⁰ I reported that the 1992 peak seemed to have occurred around 1930 UTC on August 11th. Both these dates and times bear a close proximity to those of the peaks reported in astronomical journals (1622 UTC on August 12, 1991 and 1924 UTC on August 11, 1992, respectively). Similar reports of activity also appeared in the November 1991 and October 1992 issues of *QST*.¹¹

A meteor storm?

As I mentioned, Marsden was interested in these spikes because he felt they might be foretelling the return of the comet. As I also mentioned, Rao, an amateur astronomer who has observed the *Perseids* meteor shower regularly since 1966, felt that these spikes foretold something more exciting for both astronomers and amateur radio operators. For Rao, these spikes also seemed to confirm the return of the comet. Now, with the return of the comet, it has become a Rosetta stone to him—providing the key to predict the possibility of a meteor storm for this year's *Perseids* meteor shower!

In a forthcoming article,¹² Rao reveals why the return of the comet has become the key for making this prediction. He explains that there are two important facets to this return. First, the comet's return confirms that the *Perseids* meteor shower is caused by the comet. Second, the comet's return confirms the shift in the proximity of the orbits of the comet and the Earth.

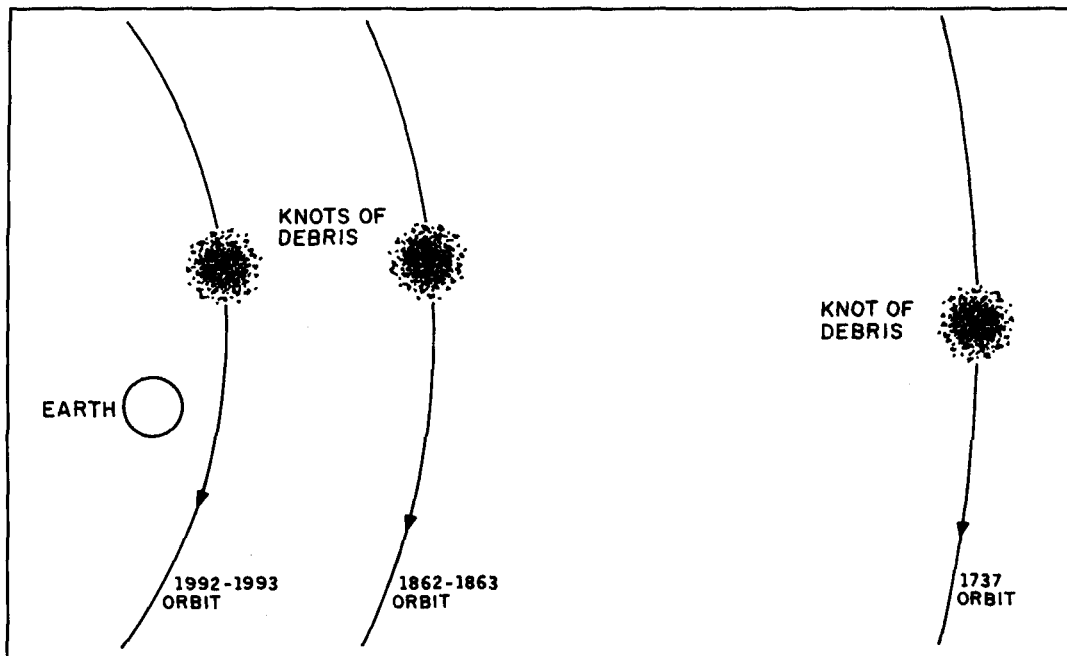


Figure 1. The orbits of the Earth and Comet Swift-Tuttle have closed to within 87,400 miles, increasing the chances that we may collide with debris expelled from the comet—possibly even with one of the knots of debris.

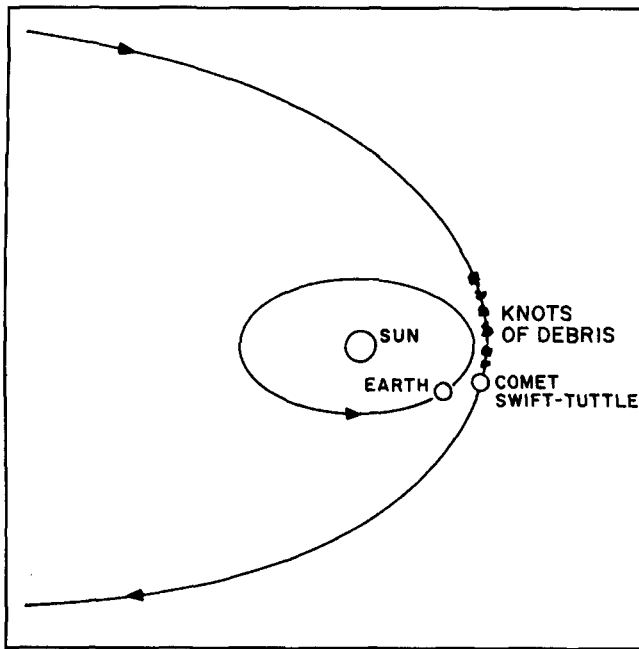


Figure 2. Knots of debris expelled from recent orbits of Comet Swift-Tuttle are circling with the comet. The Earth's collision with debris expelled from comets is what causes meteor showers. Are we on a collision course with one of these knots? If we are, several of the ham bands may experience spectacular short skip propagation during the resulting meteor storm.

Comet Swift-Tuttle Causes Perseids Meteor Showers. Even though G.V. Schiaparelli had mathematically proven that the Swift-Tuttle Comet was the originator of the Perseids meteor stream, a small doubt existed because no one had seen the comet after Schiaparelli published his findings.¹³ However, using these findings and the records of historical sightings of comets that could have been Swift-Tuttle (particularly the Comet Kegler of 1737), and assuming that the comet does cause the Perseids meteor shower, Marsden predicted the comet's next perihelion (the point where the comet passes nearest the Sun) to be November 25, 1992.¹⁴ Nevertheless, Marsden hedged his prediction by writing: "The point is, of course, that if the comet has not been found before late 1983, it would certainly be desirable to start thinking about searching...in 1992..."¹⁵ And, as he explained in the *Sky and Telescope* article,¹⁶ when the initial message of Shigemi Numazawa's September 26, 1992 sighting of the comet arrived, he wasn't particularly excited.

Comet's Return Confirms its Orbital Shift. Now that the comet had been recovered (a comet that has been presumed "lost," as in the case of the Swift-Tuttle Comet, and later "found" is considered to have been "recovered"), Rao next had to prove that its orbit had shifted closer to that of Earth. Using the orbital elements supplied in Marsden's 1973 article for the 1992 predicted return, Rao calculated that

the comet's projected perihelion would be one half million miles closer than the 1862 perihelion and 2 million miles closer than the 1737 perihelion. By extrapolation, he determined that the comet would come within 100,000 miles of Earth on its descending node. Rao took his figures to Marsden and he, in turn, calculated that the encounter with the Earth may be as close as 93,000 miles. Then, Rao took his findings to Dr. Donald K. Yeomans, who, as an employee of NASA's Jet Propulsion Laboratory, is interested in meteors from an outer space survey standpoint. Yeomans calculated that the rendezvous may be as close as 87,400 miles! (see **Figure 1**).

Because the comet has been recovered, Rao has been able to conclude that the increase in Perseids activity since 1988 can be attributed to debris expelled ahead of the comet. As the comet has proceeded through its orbit, the debris ahead of it has also traveled through the same orbit. Now that the comet has gone beyond the Earth's orbit, Rao concludes that what we will see in this year's shower is debris that has been expelled behind the comet (see **Figure 2**).

In order to conclude that this year's Perseids shower may be a storm, Rao looked at the Perseids displays for the years surrounding the previous return in 1862. Rao noted that the Far East reports indicated increases in Perseids displays for the years 1861 and 1862. He then examined a report made by William F. Denning, a tenacious meteor observer. Rao found that Denning had observed that the 1863 display produced a rate of 3 to 4 times the normal maximal Perseids rates (emphasis ours). Rao then examined the work of S. Herschel, another noted meteor observer. He found that Herschel, commenting on the Andromedid meteor storm of 1872, compared it favorably to the Leonids storm of November 1866 and the marked maximum (emphasis ours) of the Perseids shower of 1863.

Using the regularity of meteor storms (every 33, or so, years) associated with the Leonids shower, and the fact that this regularity closely follows the periodicity of the Temple-Tuttle Comet, the parent of the Leonids, most astronomers theorize that there is a knot (or knots) of debris in close proximity to the comet. Applying this theory to the Swift-Tuttle Comet, relating the intense increase in activity of the Perseids during the 1991 and 1992 showers, and noting that the comet's passage near the Earth's orbit is only 224 days for this orbit, as compared to 332 days for the 1862 orbit, Rao concludes that there exists a high probability for a storm during this year's Perseids display.

In analyzing the probability of a storm, one

other aspect Rao considered was the ability of the comet to continue to produce debris. An excellent measure of this ability is the comet's brilliance. Rao's research led him to discover contemporary comet expert John E. Bortle who has noted that the Swift-Tuttle Comet is "...4 to 10 times brighter than the average long-period comet." (Emphasis ours.) Rao concludes his research by saying: "It is quite possible to expect that, in view of the very small distance between the Earth's and comet's orbit, that we may very well encounter dust that was released as recently as 1737, and 'maybe' even 1862!"

Predicting the storm's arrival

So, when can we expect the potential storm? Rao notes that the Earth crossed the node for the 1992 orbit at 1926 UTC on August 11th and

the observed maximum for the *Perseids* shower occurred a couple of minutes earlier, at 1924 UTC. However, he observes that while the Earth crossed the node at 1331 UTC on August 12, 1991, the observed maximum was 1622 UTC—almost three hours *later*! This year the Earth will cross the node at 0115 UTC on August 12th. By extrapolation, based on the past two showers, Rao predicts that the peak could occur three hours *earlier*, or around 2222 UTC on August 11th. Rao also reports that, based on observations of past showers, the peak won't be much more than an hour long, centered on this predicted time.

Rao cautions that we may pass through a lull in debris. He feels that (however unlikely) we may actually be crossing in between two knots of debris that were expelled from the 1737 and 1862 orbits, respectively, and that these knots are still being held relatively intact.

Making Contacts Via Meteor Scatter

While the illustrations in the main article indicate that contacts made during previous storms were "nearly normal," in that the operators were able to copy each other for periods of 2 to 3 minutes at a time, contacts normally made via meteor scatter are structured much differently.

Random contacts are often possible during a meteor shower. One station calls a very brief CQ and listens for a response. A station hearing the first station returns the call, gives his callsign, and either his grid square or a signal report. The first station announces the calling station's callsign and gives the responding grid square or signal report. The second station responds by saying "Roger" several times. The contact is considered complete if both parties have all they need for the QSO. The entire contact may take as little as ten seconds to complete, if that.

However, for most meteor contacts a structured schedule is set between two stations who wish to contact each other. If you set a schedule, you'll probably run for half an hour. You'll transmit for fifteen seconds, then listen for fifteen seconds. The westward station transmits first. Some operators break at the end of seven seconds and listen briefly for the other station. Be sure to tell the other station you will be doing this before beginning your sked. The initial exchange includes the other station's callsign and your callsign, without either of you saying "his is."

For example, if I, in Oklahoma, grid square EM15, were running with Jack, AB4CR, in Kentucky, grid square EM78, I would start by saying "AB4CR N6CL" over and over again for fifteen seconds. I would then listen for Jack to say "N6CL AB4CR" over and over again during his fifteen seconds of transmission time.

After one of us has heard "complete callsigns," the receiving station begins transmitting a signal report. So, when I have heard both my call and Jack's call (in no particular order), I start repeating "S-2" over and over again during my fifteen second

segment, interspersing our callsigns—just in case Jack has yet to hear complete callsigns. (You would say the signal report is "S-2," rather than the traditional "59," as a way of telling the listener the length of the burns being heard. The letter "S" stands for the word "signal" and the number 1, 2, or 3, stands for the length of the burns. The number 1 stands for "pings;" the number 2 stands for burns long enough to make a contact; and the number 3 stands for very long burns, at least 15 to 30 seconds in length. Therefore, a signal report of "S-2" means that the sending station is hearing the receiving station on burns long enough to make a contact. As a matter of convenience, most operators stick with "S-2"—much like HF operators stick with "59.")

Assuming Jack has heard both calls and the signal report "S-2," he'll start saying "Roger, S-2," over and over again during his fifteen seconds. Once I have heard "Roger, S-2," I reply with "Roger," over and over again. Once Jack has heard my "Rogers," the QSO is considered complete. As an option, Jack can come back and say "Roger, 73," repeatedly during his sequence. However, it's not necessary to complete the contact.

Occasionally, the sequence can be broken. For example, when I ran with Ted Goldthorpe, WA4VCC, during the Perseids, I heard him give callsigns during the last 3 to 4 seconds of his fifteen-second segment. I immediately said, "WA4VCC N6CL. WA4VCC N6CL. S-2, S-2, break." Hearing me, Ted came back and said, "Roger, S-2. Roger, S-2, break." Continuing to hear him, I replied, "Roger, roger, roger, break." Hearing my "Rogers," Ted responded, "Roger, 73. Roger, 73. Break." I then replied "73, 73." At that point, the contact was considered more than complete by both of us after only three minutes into our half hour schedule.

For many, running the meteors is the most exiting part of operating on the VHF+ bands. Try it!

(Adapted, in part, from the "VHF Plus" column, *CQ* magazine, January 1993.)

Nevertheless, he has booked passage on a cruise ship that will be traversing the Mediterranean Sea, ostensibly to celebrate the birth of his new child who is due in late July, but more likely to be able to have the best advantage for observing the storm. Additionally, Rao has advised me that Marsden has made plans to travel to Italy at about the same time, also to be at an advantageous spot for viewing a potential storm.

Band conditions

Should there be a storm, what band conditions can we expect? Unfortunately, it's not entirely possible to predict with certainty, considering what propagation modes may be present at that time (sporadic E, tropo, etc.). However, generalizations can be made based on past experiences. Beginning with 12 meters, it will seem like the band is open everywhere (on short skip) during the hour or so long peak. On 10 meters, conditions will be much the same. If the storm is very intense, the same conditions that exist on 10 meters may also be present on 6 meters. On 2 meters, stations may have propagation over a given path for up to a minute, or so. On 135 cm, propagation may exist for up to five seconds, or so. On 70 cm, propagation may exist for a fraction of a second to a couple of seconds. If you want to try to complete a contact on 33 cm, this is the year to try it. If you have vacation time coming from your job, this is probably the time to schedule it!

Acknowledgements

I'd like to thank Mr. Joe Rao for graciously sharing a copy of his manuscript with me prior to its publication in *Sky and Telescope* and for taking my several phone calls at his place of employment, Compu-Weather Experts, Inc., before and after the weather "storm of the cen-

ture." I would also like to thank Drs. Brian G. Marsden and Donald K. Yeomans for spending time with me on the telephone, sharing their thoughts concerning the possibility of a meteor storm. Thanks also goes to Dennis di Cicco and *Sky and Telescope* magazine for supplying the accompanying photograph, which originally appeared in their January 1993 issue. ■

REFERENCES

1. Minor portions of this section originally appeared in the "VHF Plus" column, *CQ* magazine, August 1992, 90-91.
2. For a more in depth historical review of amateur radio operators' use of meteor scatter propagation, see *Beyond the Line of Sight*, edited by Emil Pocock, W3EP, pages 95-122, published by the ARRL. Many articles originally published in *QST* are contained within this publication.
3. Ed Tilton, "The World Above 50 Mc.," *QST*, December 1946, page 43.
4. John Stewart, et al., "Radar Observations of the Draconids," *Sky and Telescope*, Issue 65, pages 3-4. The shower became known as the Draconids because the radiant, the point of origin in the sky of the meteor shower, appeared to radiate from the head of the Draco star constellation.
5. Walter F. Bain, "VHF Meteor Scatter Propagation," *QST*, April 1957, pages 20-24, 140, 142, 144.
6. See November 1966 issues of both *Sky and Telescope* (page 251) and *Natural History* (pages 43-44).
7. Sam Harris, "The World Above 50 Mc.," *QST*, January 1967, Pages 83-85. By this time, Sam succeeded Ed as the editor of this column.
8. Brian G. Marsden, "Comet Swift-Tuttle: Does it Threaten Earth?," *Sky and Telescope*, January 1993, pages 16-19. This article principally deals with dispelling the fears that the comet will collide with earth in 2126. However, within this article Marsden acknowledges the reports generated by Shelby Ennis, W8WN, and exclaims, "Maybe the comet really was coming!"
9. Joe Lynch, N6CL, "VHF Plus," *CQ*, November 1991, page 83.
10. Joe Lynch, N6CL, "VHF Plus," *CQ*, November 1992, page 94, 96-98.
11. Bill Tynan, "The World above 50 MHz," *QST*, November 1991, page 84 and Emil Pocock, "Perseids Resurge in 1992," part of "The World Above 50 MHz," *QST*, October 1992, page 96.
12. This article has been accepted for publication by *Sky and Telescope*. It is scheduled for publication in the August 1993 edition.
13. G.V. Shiaparelli, 1867, *Astron. Nachr.* 68, 331.
14. B.G. Marsden, "The next return of the comet of the *Perseids* meteors," *The Astronomical Journal*, September 1973, pages 654-662.
15. B.G. Marsden, "The next return of the comet of the *Perseids* meteors," *The Astronomical Journal*, September 1973, page 662. During the course of his research, Dr. Marsden tried to make a comet sighting in 1748 "fit" with the comet sighting of 1862. Finding that there were flaws in this assumption, he went back to a comet sighting in 1737 (the Kegler Comet). In order to predict the 1992 sighting, he assumed that since the 1737 sighting was 11 years earlier, the next sighting would be 11 years later than a possible 1981 date. Yet, he would leave the door open for discussion by saying: "Our procedure for forcing the linkage of the 1737 and 1862 observations is certainly open to question, and these resulting observations is certainly open to question, and these resulting 1992 predicted elements are consequently somewhat uncertain."
16. Brian G. Marsden, "Comet Swift-Tuttle: Does it Threaten Earth," *Sky and Telescope*, January 1993, page 18.

PRODUCT INFORMATION

New Hamtronics® Digital Voice Recorder

The DVR-1 Digital Voice Recorder is a versatile pc board module, which is designed primarily as a voice ID'er for repeaters. But it also provides features so that it can be used as a contest CQ caller or as a "radio notepad" to record short parts of received transmissions for instant recall.

As a repeater ID'er, the DVR-1 module will record your voice, using either the built-in microphone or an external mic. It can be used with almost any repeater COR module, including the Hamtronics® COR-3, COR-4, and

COR-5 (used in the REP-200 Repeater).

The DVR-1 module is an economical pc board which can be purchased either in kit form at \$89 or as a wired & tested unit at \$139. It includes a small electret microphone and pushbuttons for record and playback.

For more details, write to Hamtronics, Inc., 65-Q Moul Rd, Hilton, NY 14468-9535 or call 716-392-9430 (fax: -9420). Ask for a complete catalog, which also includes all their VHF/UHF transmitters, receivers, repeaters, converters, preamps, and accessories.

Reviewed by: Rick Littlefield, K1BQT
109A McDaniel Shore Drive
Barrington, New Hampshire 03825

RADIO FREQUENCY TRANSISTORS: PRINCIPLES AND PRACTICAL APPLICATIONS

*A practical guide for the roll-up-your
sleeves builder by Norm Dye and
Helge Granberg*

Not all radio designers are formally trained for what they do. When I went to school, communication technology as we know it simply didn't exist. Even today, digitally oriented under-graduate EE programs rarely offer comprehensive RF-design courses. Many of us learn our craft from a mosaic of conferences and workshops, application notes, journal articles, data sheets, CAD programs, and the ever-present school of hard knocks. And, when these avenues fail us, we pick up the phone and call a factory RF application engineer for help. But wouldn't it be great to have an application engineer's knowledge available on your bookshelf?

Design course in a book

As its introduction states, *Radio Frequency Transistors: Principles and Practical Applications*, isn't a theoretical text about the physics of semiconductors. Rather, it's a practical guide for people who roll up their sleeves

and build things. Norm Dye and Helge Granberg are both well-known Motorola application engineers who routinely field questions over the phone, teach workshops, write articles and application notes, and design products. They are the people you and I call when we run out of rope and nothing seems to work. What's more, they have been doing this for over 20 years. Consequently, both bring a particularly keen sense of what the practical designer needs to conquer the "Murphys" that creep into RF amplifier designs. In addition to technical expertise, both authors have mastered the fine art of presenting technical material simply, without diluting its integrity or meaning. In short, they know a lot of useful stuff, and you probably won't have a difficult time understanding it.

Overall, I found *Radio Frequency Transistors* thorough and well organized. Like a well-planned series of workshops, each chapter takes on a single aspect of amplifier design, and covers it from start to finish (see the list of chapters). Because everything is well organized, the

book is easy to study and especially useful as a reference. Say, for example, you wish to cookbook a SSB bias network. The Index or Table of Contents will quickly lead you to a range of proven circuits—cataloged in order of complexity, and explained in enough detail so you can implement them easily. It then becomes a simple matter of selecting the best solution for your particular application. No thumbing through obscure texts or piles of application notes in search of something that “looks” workable.

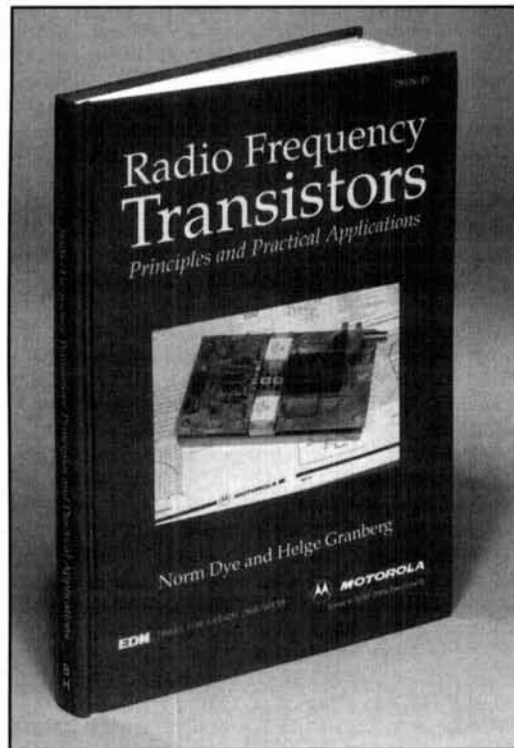
In addition to being practical, the book takes a forward-looking approach to technology. Dye and Granberg are quick to acknowledge the utility of PC-oriented CAD packages for designing impedance-matching networks. They cover traditional “pencil-and-paper” calculations for conjugate networks (an obligatory inclusion), but then devote a full chapter to powerful CAD alternatives that provide far more exciting opportunities than a hand calculator and a piece of Smith Chart paper could ever yield. Not surprisingly, Dye and Granberg focus special attention on MIMP (Motorola Impedance Matching Program), a special in-house CAD package developed by Motorola that features Smith Chart visualization. MIMP is a very useful tool, and it’s free (available upon request from Motorola)!

The inside information

The content of this book is well focused, providing the basics you’ll need to get the job done. The text covers a lot of ground in 235 pages, so you won’t learn everything there is to know about each topic. On the other hand, you’ll learn what you need to know to implement a workable design. And, if you do require more information, each chapter concludes with an extensive list of references.

In the Table of Contents, each chapter listing is followed by several sub-headings—reading something like a course syllabus. Chapter titles provide a pretty good clue as to the range of topics covered in the book. They are:

1. Understanding RF Data Sheet Parameters
2. RF Transistor Fundamentals
3. FETs and BJTs: Comparison of Parameters and Circuitry
4. Other Factors Affecting Amplifier Design
5. Reliability Considerations (die temperatures, etc.)
6. Construction Techniques (pc layout, mounting, etc.)
7. Power Amplifier Design
8. Computer Aided Design Programs
9. After the Power Amplifier Output (filters, SWR protection)



10. Wideband Impedance Matching
11. Power Splitting and Combining
12. Frequency Compensation and Negative Feedback
13. Small Signal Amplifier Design

As with most good technical reference books, a detailed topical index will help you find specifics quickly.

A valuable resource

Over the years, Motorola Semiconductors has distinguished itself with a strong commitment to customer support and product education. This, in turn, has enabled career designer/educators like Norm Dye and Helge Granberg to acquire a tremendous store of “real world” technical and instructional savvy. *Radio Frequency Transistors: Principles and Practical Applications* is a compilation that represents the best of what this partnership in learning has produced, and it is a book well worth having on your shelf.

Radio Frequency Transistors: Principles and Practical Applications, by Norm Dye and Helge Granberg is published by Butterworth-Heinemann, 80 Montvale Avenue, Stoneham, Massachusetts 02180. The list price is \$39.95. To order by phone, call (800)366-2665 or FAX (617)438-1479; a \$3.50 handling fee will be added to all credit card orders. The book is also available in technical bookstores nationwide, and from the Motorola Literature Distribution Center as publication TB326. ■

1.5:1 and 2:1 Baluns

A look at the many versions of these two baluns

There are many applications for broadband baluns with impedance transformation ratios close to 1.5:1 and 2:1. Two applications involve matching 50-ohm cable to balanced loads of 75 or 100 ohms, which are the input impedances of a half-wave dipole at heights of 0.22 or 0.34 wavelengths above ground. Another is the matching of 50-ohm cable to the 100-ohm input impedance of a quad antenna. An interesting and somewhat unexpected application concerns the matching of 50-ohm cable directly to the input impedance of the driven element of a Yagi beam antenna of 33 or 25 ohms. This would eliminate the common hairpin matching network presently used to raise their input impedances to 50 ohms.

There are many versions of these two baluns.

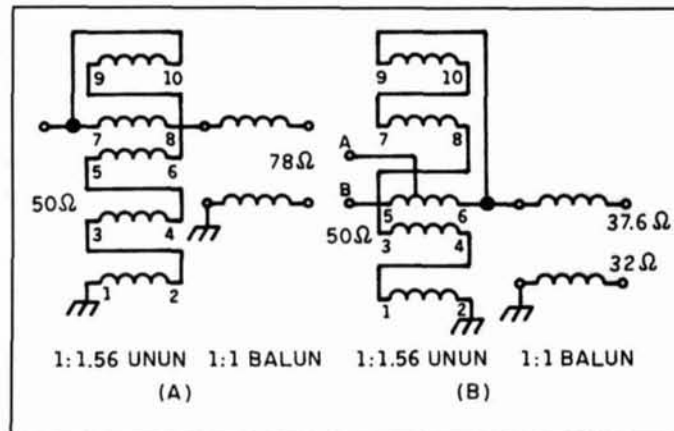


Figure 1. Schematic diagrams of two 1.56:1 baluns. (A) step-up, 50:78 ohms; (B) step-down 50:37.6 ohms—connection A, 50:32 ohms—connection B.

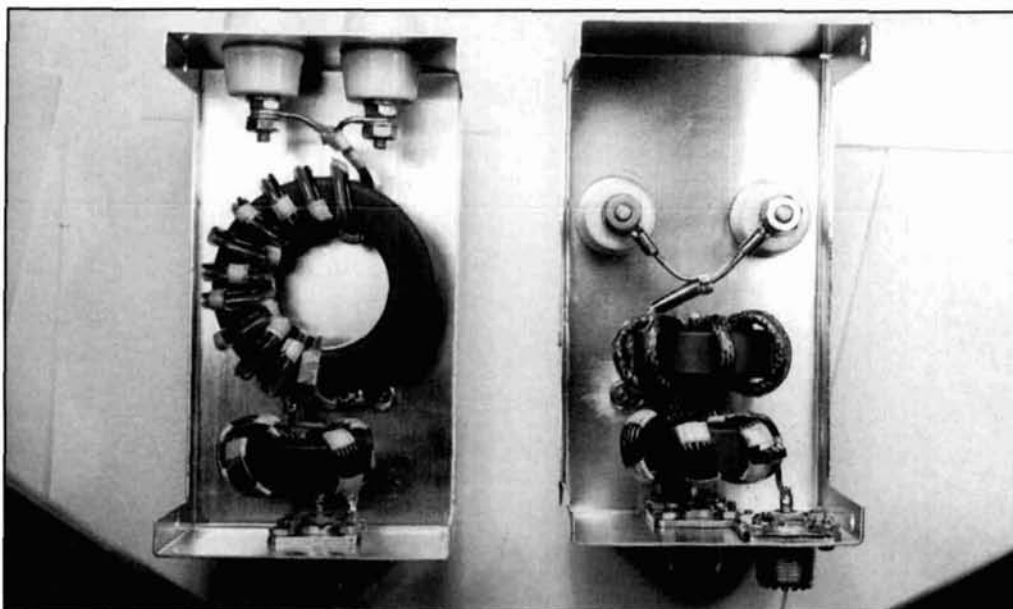


Photo A. Baluns using the schematic diagram of Figure 1A. Balun on the left matches 50-ohm cable to a balanced load of 78 ohms. Balun on the right matches 50-ohm cable to balanced loads of 37.6, or 32 ohms..

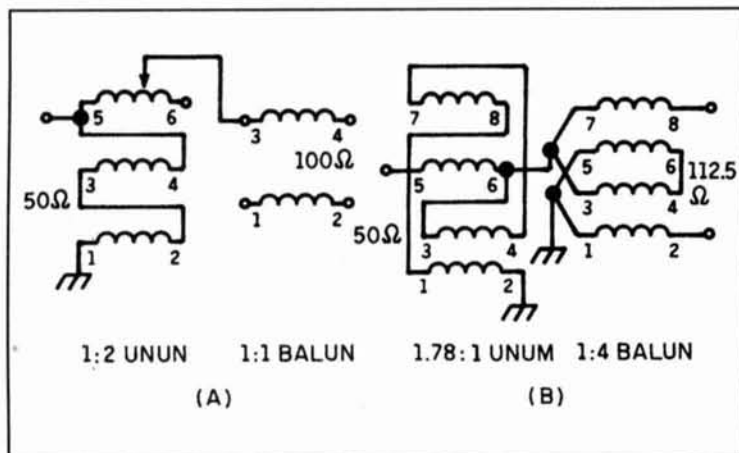


Figure 2. Schematic diagrams of two versions of the series-type balun. (A) 1:2 (50:100 ohms) balun; (B) 1:2.25(50:112.5 ohms) balun.

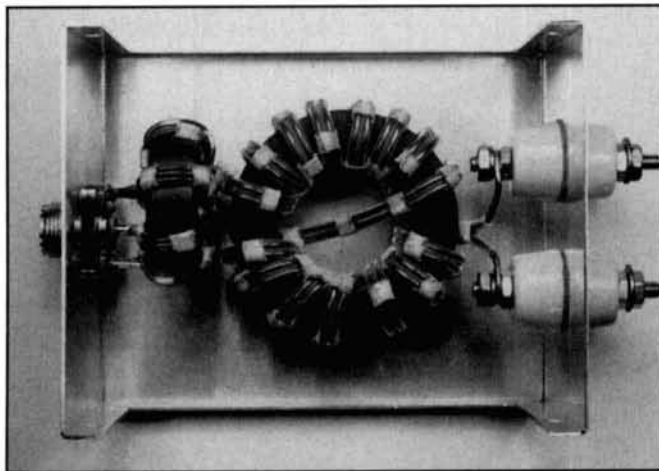


Photo B. A series-type balun using the schematic diagram of Figure 2A designed to match 50-ohm cable to a balanced load of 100 ohms.

They include: 1) high and low-power designs, 2) designs matching 50-ohm cable to higher or lower impedances, 3) series or parallel-type designs, 4) single or dual-core designs, 5) dual-ratio designs, and 6) HF and VHF designs. In this article, you'll read about many high-power designs* capable of handling the full legal limit of amateur radio power. They are optimized for sufficient margins in choking reactance at their low frequency ends and in efficiency throughout their passbands. Two of the 2:1 baluns are specifically designed for 2-meter operation.

1.5:1 baluns

In this section, I'll present two series-type 1.5:1 baluns (actually 1.56:1, which should be close enough). They both use 1.56:1 ununs¹ in

*Available from Amidon Associates, Inc., 2216 East Gladwick Street, Dominguez Hills, California 90220.

series with Guanella² baluns. **Figures 1A and B** show their schematic diagrams. **Figure 1B** has an extra input (to a tap) that provides another ratio of 1.33:1.

The left-hand side of **Photo A** shows a design using **Figure 1A** mounted in a CU 3006 minibox 5.25 inches long by 3 inches wide by 2.25 inches high. (Radio Shack carries a similar enclosure.) The 1:1.56 unun has 4 quintufilar turns on a 1.5-inch OD ferrite toroid with a permeability of 250. Winding 7-8 is no. 14 H Thermaleze wire and the other four are no. 16 H Thermaleze wire.

The 1:1 Guanella balun has 11 bifilar turns of no. 14 H Thermaleze wire on a 2.4-inch OD ferrite toroid with a permeability of 250. One wire is covered with Teflon™ tubing resulting in a characteristic impedance very close to 78 ohms (the optimum value).

In matching 50-ohm cable to a balanced load of 78 ohms, the impedance transformation ratio is literally flat (within a percent or two) from 1.5 to 40 MHz! Note that (separately) the 1:1 (75:75 ohm) balun would make an excellent isolation transformer for 75-ohm hardline and the 1.56:1 (78:50 ohm) unun would provide an excellent match between 75-ohm hardline and 50-ohm cable.

The right-hand side of **Photo A** shows a design using **Figure 1B** mounted in a similar enclosure. The 1.56:1 unun has 5 quintufilar turns on a 1.5-inch OD ferrite toroid with a permeability of 250. Winding 5-6 is no. 14 H Thermaleze wire and is tapped at one turn from terminal 5. The other four wires are no. 16 H Thermaleze wire.

The 1:1 Guanella balun has 7 turns of home-made coaxial cable on a 1.5-inch OD ferrite toroid with a permeability of 250. The inner conductor is no. 14 H Thermaleze wire and is covered with Teflon tubing. The outer braid, which is from a small coaxial cable or from 1/8-inch tubular braid, is also tightly wrapped with Scotch no. 92 tape to preserve the low characteristic impedance.

In matching 50-ohm cable to a balanced load of 37.6 ohms (connection A) or to a balanced load of 32 ohms (connection B), the response is essentially flat (within a percent or two) from 1.5 to 30 MHz.

2:1 baluns

The 2:1 balun lends itself to more choices in design than the 1.56:1 balun. This is especially true because the parallel-type design, which provides a 2.25:1 balun with the widest possible bandwidth, can easily be employed. The 1.56:1 balun is at a disadvantage here. This section contains many baluns using both series and parallel-type designs. Analysis of the parallel-

type 2.25:1 balun is probably presented here for the first time.

Series-type baluns

Figure 2 shows circuit diagrams for two versions of the series-type balun. **Photo B** shows a design using **Figure 2A** mounted in a CU 3005-A minibox 5 inches long by 4 inches wide by 3 inches high. The 1:2 unun³ has 7 trifilar turns on a 1.5-inch OD ferrite toroid with a permeability of 250. The output tap is located 6 turns from terminal 5. Winding 5-6 is no. 14 H Thermaleze wire and the other two are no. 16 H Thermaleze wire.

The 1:1 Guanella balun has 14 bifilar turns of no. 14 H Thermaleze wire on a 2.4-inch OD ferrite toroid with a permeability of 250. Both wires are covered with Teflon tubing, resulting in a characteristic impedance of 100 ohms (the optimum value). A crossover, placing 7 turns on one side of the toroid and 7 turns on the other, is used so that the output and input are on opposite sides of the toroid. **Figure 3** is a drawing of the crossover. Although this technique has no electrical advantage at HF, the mechanical advantage is obvious.

When matching 50-ohm cable to a balanced load of 100 ohms, the response is literally flat (within 2 to 3 percent) from 1.5 to 30 MHz. By connecting the output of the unun to terminal 6 instead of to the tap, the balun would match 50-ohm cable to a balanced load of 112.5 ohms with about the same response.

Photo C shows two slightly different versions of series-type 2.25:1 baluns using the circuit of **Figure 2B**. Both have the same 1.78:1 step-down unun which has 5 quadrifilar turns on a 1.5-inch OD ferrite toroid with a permeability of 250. Winding 5-6 is no. 14 H Thermaleze wire and the other three are no. 16 H Thermaleze wire. Each version also has 8 bifilar turns of no. 14 H Thermaleze wire on both of the 1.5-inch OD ferrite toroids with a permeability of 250.

The differences are: 1) the balun on the left in **Photo C** has one layer of Scotch no. 92 tape on one of the wires in each bifilar winding and a crossover after the fourth turn, and 2) the balun on the right has two layers of Scotch no. 92 tape on one of the wires on one toroid and no extra insulation on the wires of the other toroid. Therefore, one of the windings in the 1:4 Guanella balun has a characteristic impedance a little less than 50 ohms resulting in a canceling effect. Furthermore, the crossover isn't used in this design. The balun on the left is mounted in a CU 3006 minibox 5.25 inches long by 3 inches wide by 2.25 inches high. The balun on the right is mounted in a CU 3015-A minibox 4 inches long by 2 inches wide by 2.75 inches high.

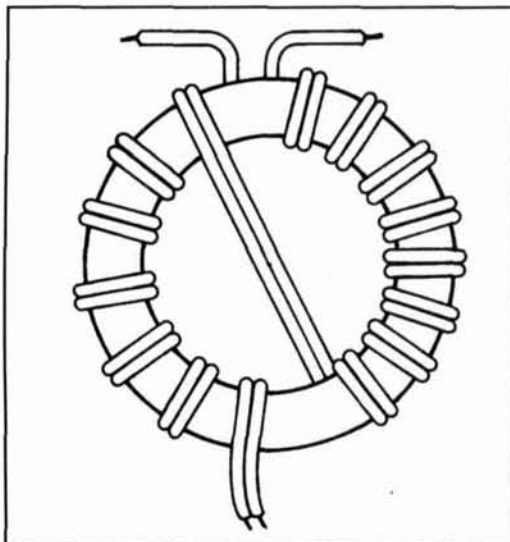


Figure 3. Construction of a 1:1 Guanella balun with a crossover placing the input and output terminals on opposite sides of the toroid.

The performance of these two baluns is essentially the same. When matching 50-ohm cable to balanced loads of 112.5 ohms, the responses are essentially flat (within 2 to 3 percent) from 1.5 to 30 MHz.

From preliminary measurements on series-type 2:1 baluns, the balun in **Photo B** is the one I'd recommend for matching 50-ohm cable to balanced loads of 100 ohms, while the baluns in **Photo C** would be best for matching to balanced loads of 112.5 ohms. Also by replacing the 1.78:1 unun in **Figure 2B** with a 2.25:1 unun and not adding any extra insulation to the windings of the 1:4 balun, it's possible to

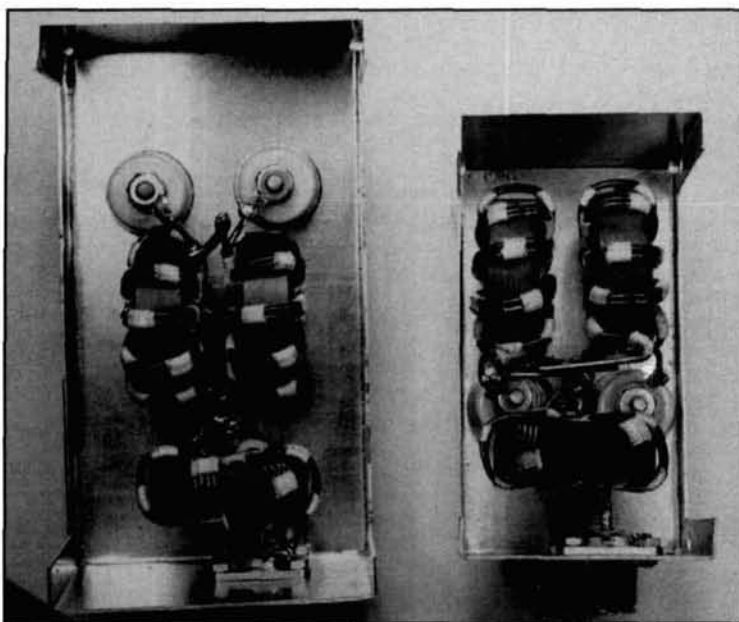


Photo C. Two series-type baluns using the schematic diagram of **Figure 2B** designed to match 50-ohm cable to balanced loads of 112.5 ohms.

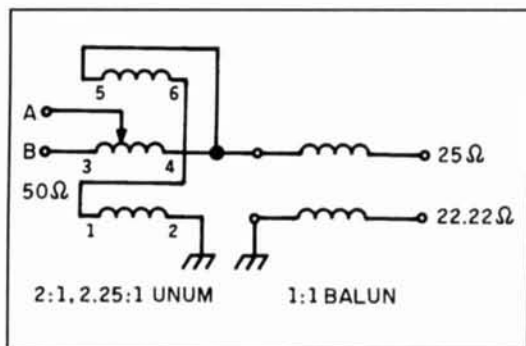


Figure 4. Schematic diagram of a dual-ratio series-type balun. Connection A matches 50-ohm cable to a balanced load of 25 ohms. Connection B matches 50-ohm cable to a balanced load of 22.22 ohms.

obtain an excellent balun matching 50-ohm cable to a balanced load of 89 ohms.

Figure 4 shows the schematic diagram of a series-type balun designed to match 50-ohm cable to balanced loads of 25 or 22.22 ohms.

Photo D shows two versions of this dual-ratio balun. The balun on the left, for its unun, has 6 trifilar turns of no. 14 H Thermaleze wire on a 1.5-inch OD ferrite toroid with a permeability of 250. Winding 3-4 is tapped one turn from terminal 3, yielding the 2:1 ratio. The 1:1 Guanella balun has 6 turns of homemade coaxial cable on a similar toroid. The inner conductor is no. 12 H Thermaleze wire and is covered with Teflon tubing. The outer braid, from a small coax or 1/8-inch tubular braid, is tightly wrapped with Scotch no. 92 tape to preserve its low characteristic impedance.

In matching 50-ohm cable to balanced loads

of 25 ohms (connection A) or to 22.22 ohms (connection B), the response is essentially flat (within a percent or two) from 1.5 to 30 MHz.

The balun on the right in Photo D has similar windings on a single 2.4-inch OD ferrite toroid with a permeability of 250. Its performance is quite comparable.

Parallel-type baluns

As you saw in the previous section, the series-type baluns in this article are combinations of ununs with ratios of 1.33:1 to 2.25:1 in series with Guanella 1:1 or 4:1 baluns. The ununs, which are really an extension of Ruthroff's⁴ bootstrap technique for obtaining a 4:1 unun, sum a direct voltage with a delayed voltage that traverses a single transmission line. Therefore, the unun eventually limits the high-frequency response of the series-type balun.

On the other hand, the parallel-type balun is an extension of Guanella's technique of summing voltages of equal delays. Instead of connecting only transmission line in parallel-series, the parallel-type balun connects Guanella baluns in parallel-series. As I noted in my article in the Winter 1992 issue of *Communications Quarterly*,⁵ two 4:1 Guanella baluns can be connected in parallel-series, yielding very broadband ratios of 6.26:1. This section shows how a 1:1 Guanella balun can be connected with a 4:1 Guanella balun in parallel-series, yielding a very broadband ratio of 2.25:1.

Figure 5 is the schematic diagram of the high-frequency model of a 2.25:1 unun used for analysis purposes. Since the current through the

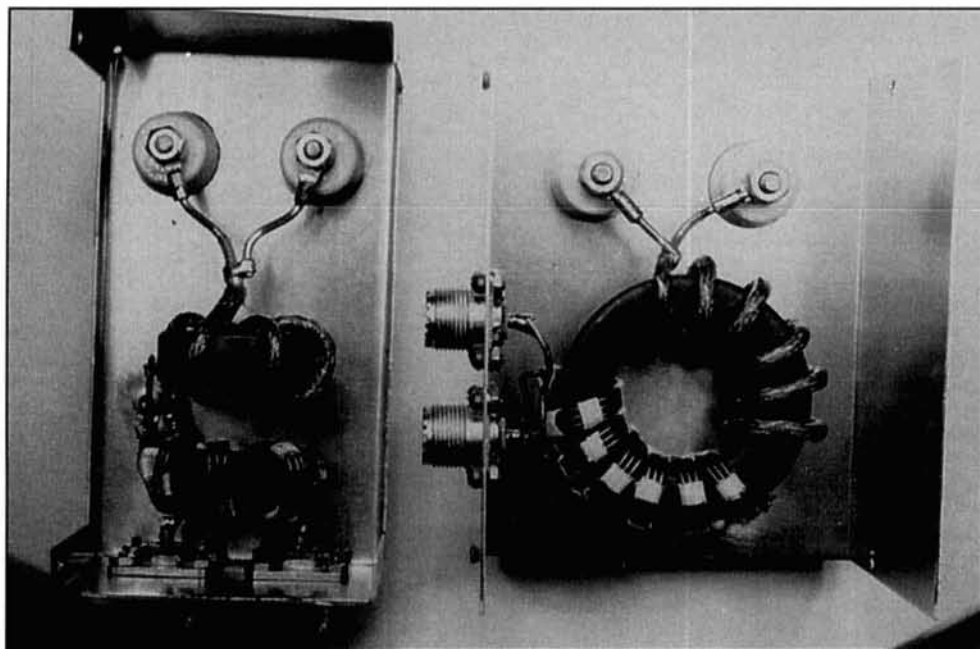


Photo D. Two series-type baluns using the schematic diagram of Figure 4 designed to match 50-ohm cable to balanced loads of 25 ohms or 22.22 ohms.

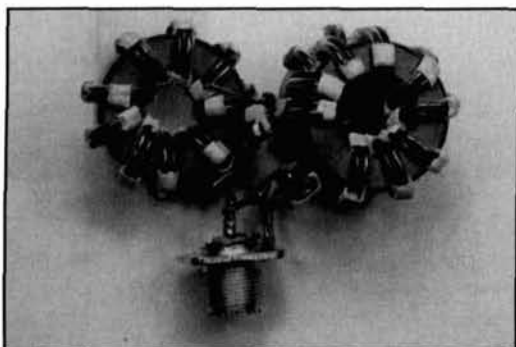


Photo E. Parallel-type 2.25:1 balun matching 50-ohm cable to a balanced load of 112.5 ohms.

load is $3/2 I_1$, the transformation ratio is $(3/2)^2$, or 2.25:1. Therefore, if the impedance seen on the left side is 50 ohms, a matched impedance on the right side is 2.22 ohms. Since 2/3rds of the 50 ohms appears across the input of the Guanella 1:1 balun, its optimum characteristic impedance is 33.33 ohms. This is also the value of the optimum characteristic impedance for the windings of the 4:1 balun. Because the 1:1 balun wants to see 33.33 ohms on its output (a matched condition) and the 4:1 balun 66.66 ohms, putting these two values in parallel results in the confirming value of 22.22 ohms.

If the 50-ohm generator is placed on the right side in Figure 5, the circuit becomes a step-up unun matching 50 ohms to 112.5 ohms (on the left). If the ground is removed on the left side, the transformer becomes a balun. A similar analysis as above, shows that the optimum characteristic impedance of the three bifilar windings now becomes 75 ohms.

Photo E shows a parallel-type 2.25:1 balun designed to match 50-ohm cable to a balanced load of 112.5 ohms. It has 9 bifilar turns of no. 14 H Thermaleze wire on each of the three toroids that have a 1.5-inch OD and a permeability of 250. Also, one of the wires on each toroid is covered with Teflon tubing resulting in a characteristic impedance of 75 ohms (the optimum value). When operating as a balun, the response is essentially flat from 7 to over 45 MHz. As an unun, the flat response is broadened to 1.5 to over 45 MHz.

Because the coiled wire, parallel-type balun didn't provide any real advantage over the series-type balun (in fact the low-frequency response was poorer), I investigated the beaded transmission line version for possible use in the VHF band. Figure 6 shows a schematic diagram of one using coaxial cable. Obviously, twin lead could be substituted for the coaxial cable. Photo F shows both versions.

The top design in Photo F has 4 inches of 3/8-inch OD ferrite beads with a permeability of 125 on each of the three 75-ohm transmis-

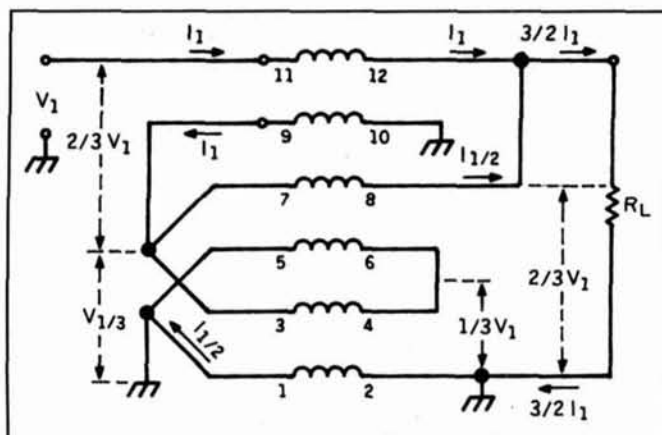


Figure 5. High-frequency model of the parallel-type 2.25:1 transformer. Connections shown are for unun operation.

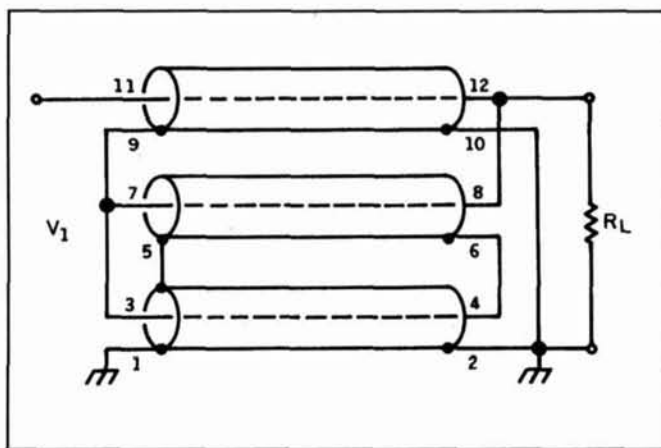


Figure 6. The coaxial-cable version of the parallel-type 2.25:1 transformer of Figure 5.

sion lines. It is designed to match 50-ohm cable on the right to a balanced load of 112.5 ohms on the left (with the ground on terminal 1 removed). The transmission lines consist of two no. 14 H Thermaleze wires separated by the Teflon tubing covering one of the wires. When matching 50-ohm cable to a balanced load of 112.5 ohms, the response is essentially flat from 30 to over 100 MHz (the limit of my bridge).

The bottom design in Photo F also has 4 inches of 3/8-inch OD ferrite beads with a permeability of 250. But they are now threaded by homemade coaxial cable with a characteristic impedance of 33.33 ohms. It is designed to match 50-ohm cable on the left to a balanced load of 22.22 ohms (with the ground on terminal 2 removed). The inner conductor of the coax is no. 14 H Thermaleze wire and is covered with Teflon tubing. The braid is from small coaxial cable or from 1/8-inch tubular braid. The braid is also tightly wrapped with Scotch no. 92 tape in order to preserve the low

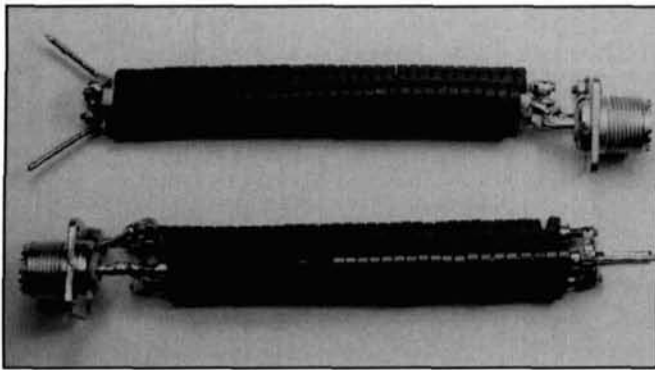


Photo F. Beaded transmission line versions of the 2.25:1 parallel-type balun for operation in the VHF band. The top balun matches 50-ohm cable to a balanced load of 112.5 ohms. The bottom balun matches it to a balanced load of 22.22 ohms.

characteristic impedance. When matching 50-ohm cable to a balanced load of 22.22 ohms, the response is essentially flat from 14 to over 100 MHz (again, the limit of my bridge).

Closing comments

In closing, I'd like to make a couple of comments regarding parallel-type baluns.

First, if you are interested in a 1.78:1 ratio, replace the 1:4 balun in **Figures 5 and 6** with a 1:9 Guanella balun (three transmission lines connected in parallel-series). This will yield an

output current of $4/3I_1$ and a ratio of $(4/3)^2$, or 1.78:1. If you replace the 1:4 balun with a 1:16 Guanella balun (four transmission lines connected in parallel-series), the output current will be $5/4I_1$ with a ratio of $(5/4)^2$, or 1.56:1.

Second, because the parallel-type balun is really an extension of Guanella's technique of summing voltages of equal delays^{6,7} the high-frequency response is mainly limited by the parasitics in the interconnections. Therefore, beaded transmission lines offer the best opportunity for successful operation in the VHF band. I also recommend that the ferrite beads have permeabilities of 300 or less⁸ in order to achieve the very high efficiencies of which these transformers are capable. ■

REFERENCES

1. Jerry Sevick, W2FMI, "The 1.5:1 and 1.33:1 Ununs," *CQ*, November 1992, pages 26-30.
2. G. Guanella, "Novel Matching Systems for High Frequencies," *Brown-Boveri Review*, Vol. 31, September 1944, pages 327-329.
3. Jerry Sevick, W2FMI, "The 2:1 Unun Matching Transformer," *CQ*, August 1992, pages 13-18.
4. C.L. Ruthroff, "Some Broad-band Transformers," *Proceedings of the IRE*, Vol. 47, August 1959, pages 1337-1342.
5. Jerry Sevick, W2FMI, "6:1 and 9:1 Baluns," *Communications Quarterly*, Winter 1992, pages 43-51.
6. D. Meyer, "Equal-Delay Networks Match Impedances Over Wide Bandwidths," *Microwaves & RF*, April 1990, pages 179-188.
7. S.E. London and S.V. Tomeshevich, "Line Transformers with Fractional Transformation Factor," *Telecommunications and Radio Engineering*, Vol. 28, 29, April 1974.
8. Jerry Sevick, W2FMI, *Transmission Line Transformers*, 2nd Edition, Amateur Radio Relay League, Newington, Connecticut, 1990, Chapters 7, 9, 11.

PRODUCT INFORMATION

New Catalog of Quartz Crystals Published

A new catalog of quartz crystals and holders has just been published by JAN Crystals. It is an illustrated eight page booklet containing descriptions, specifications, and prices on crystals for frequency control, microprocessors, data transmission, telemetry, and telecommunication voice.

To receive a free copy of the catalog, call JAN Crystals at 1-800-JAN-XTAL.

Three-Dimensional Radiation Patterns

Brian Beezley, K6STI, announces a new feature for the AO 5.0 Antenna Optimizer program. AO will include 3-D, radiation-pattern display. The display shows an antenna's entire radiation pattern at once. The pattern may be rotated, translated, scaled, and sliced by the user. The display uses a dual-illumination technique (not reproducible on printed patterns). 3-D objects float against a black background and appear to be illuminated from the

front by a bright white light and from the rear by a soft blue glow. Colors are user-selectable, as are optional 3-D axes and labels. Elevation-plane resolution may be set from 2 to 6 degrees. Azimuth-plane resolution varies automatically with elevation angle.

The new 3-D display lets an antenna designer easily evaluate results of AO's three-dimensional radiation-pattern optimization. The display can reveal radiation-pattern problems not otherwise apparent, like unexpected high-angle backlobes away from the principle axes.

AO-Professional 5.0, which includes AO 5.0, the MN-Professional 4.5 Antenna Analysis package, and a commercial-use license, costs \$500. AO 5.0 is discounted to \$100 for amateur use. AO 5.0 requires an IBM PC compatible computer with math coprocessor. For more information contact: Brian Beezley, K6STI, 507 1/2 Taylor, Vista, CA 92084; or call 619-945-9824 (0700-1800 Pacific Time).

Claton Cadmus, KA0GKC
3985 Alabama Avenue South,
St. Louis Park, Minnesota 55416
and

Bruce L. Meyer,
WOHZR/NNN0BQH
9410 Blaisdell Avenue South,
Bloomington, MN 55420-4323

FSK SIGNAL MONITOR

*"X-display" circuit makes HF signal
tuning easier*

Radio amateurs and MARS operators who use the digital modes packet, AMTOR, and RTTY, and have struggled to properly tune in a noisy HF signal while watching a TNC bargraph tuning indicator, will appreciate the ease of tuning afforded by the "X" display on a CRT screen.

The "X"-display circuit gets its input directly from the radio receiver audio rather than from filters following limiters inside a terminal unit, modem, or TNC. As a result, it provides more information to the user than a bargraph or crossed-ellipse type indicator. The display shows the amount of frequency shift, the direction of mistuning,

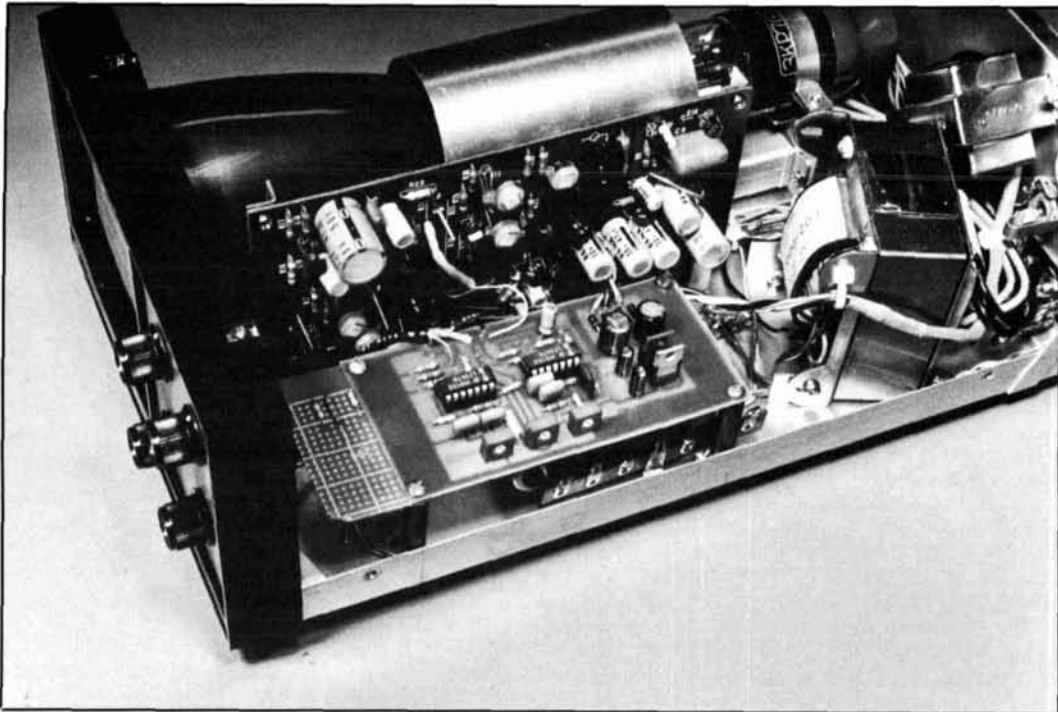


Photo A. Typical oscilloscope conversion.

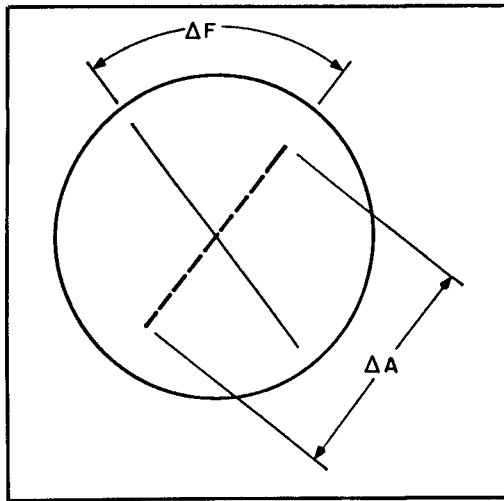


Figure 1. The CRT display.

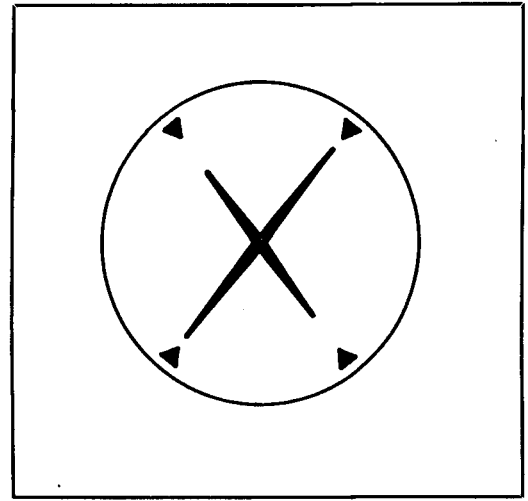


Figure 2. Typical pattern for correct tuning.

selective fading characteristics of the signal, and some keying irregularities. Yet, it's easy to build and calibrate.

Nearly any oscilloscope can be used as the foundation of the FSK monitor, as long as it contains electrostatic deflection circuits. You should be prepared to dedicate the 'scope to the FSK monitoring function, because the gain must be reproducible, and because a bright beam will eventually produce a small dark spot in the center of the CRT screen.

We have recently installed the FSK monitor circuit in four medical monitor 'scopes, three "kit" 'scopes, and two old service 'scopes containing vacuum tubes. To make installation easier, we removed the sweep circuits—creating additional space. **Photo A** illustrates one such modification; the monitor conversion board is in the foreground. A small perfboard holds three biasing resistors needed for trace centering.*

Background

Hams in several countries, who were aware of early work on a 850-Hz "phase shift" RTTY monitor,** have sent requests and technical inquiries about a 170 or 200-Hz shift monitor to meet the needs of AMTOR and HF packet users. This article

*WOHZR offers free assistance to any reader planning an oscilloscope conversion. Simply send the 'scope schematic to him at the address found at the beginning of the article, along with an SASE.

**Meyer described the original vacuum-tube version in *RTTY Magazine* in November 1954. Additional articles appeared in *CQ* magazine of May 1956, the *W2JTP RTTY Handbook*, and the *RSGB Communications Handbook*. Meyer's solid-state FSK monitor appeared in *Ham Radio* magazine in September 1988. Recent improvements to this design are offered here.

Claton Cadmus, KA0GKC, gave invaluable assistance in the preparation of this latest article and also provided the printed circuit board layout and the component placement diagram.

offers recent improvements that were made to the original design.

The display

The tuning indicator provides a line on the face of the cathode ray tube (see **Figure 1**). This line passes through the center of the screen and rotates as the audio frequency, F , changes. The line leans to one side for an increase in frequency and to the other side for a decrease.

About midway between the marking and spacing frequencies, the line is vertical. The rotation rate is greatest near the center frequency and slower above and below the FSK spectrum of interest. The length of the line is proportional to signal amplitude, A .

Two rapidly alternating frequencies appear as an X on the monitor screen, as shown in **Figure 2**. Calibration marks on the CRT screen permit the operator to evaluate the FSK signal for correct shift. The angle formed by the X increases with increased shift. A tilted X indicates that the radio receiver is incorrectly tuned. The presence of more than two straight lines is an indication that the sender's keying contacts are bouncing, or that a frequency-division multiplex signal is being received. Selective fading characteristics of the radio signal appear as lines of unequal length. Often one line fades completely. It comes as a shock to many operators to learn that two frequencies only 170 Hz apart can differ in amplitude by 20 dB or more.

Noise signals and sidebands appear as curved lines or Lissajous figures on the screen. Faint "retrace" lines related to filter Q appear inside the X. In 300-baud packet operation, sideband energy is clearly ap-

parent beyond the legs of the X, which appear brighter than the sidebands.

Theory of operation

The block diagram in **Figure 3** shows the FSK monitor's simplicity. The audio output of the radio receiver is applied to the input of the monitor circuit board, either directly or by way of a gain control potentiometer. An input buffer amplifier provides a constant source impedance to the second stage—a bandpass filter that includes the marking and spacing frequencies. This filter circuit emphasizes the frequencies of interest, while allowing the operator to recognize and acquire off-frequency signals. The third stage is a series-resonant discriminator circuit that is tuned to the central frequency of the bandpass filter. This stage separates the vertical and horizontal deflection components of the signal for the CRT display. The vertical component provides a reference level and phase as the receiver is tuned. The horizontal component is the vector sum of the voltages across the capacitor and the inductor. This voltage goes to zero at resonance because the voltages across C and L become equal and opposite in phase. When they add up to a null, the CRT displays a vertical line. On either side of resonance, the voltages across the C and L become unequal, causing the line on the CRT face to slope toward one side for the marking frequency and to the other side for the spacing frequency.

Separate amplifiers for the marking and spacing outputs of the discriminator allow adjustment of line slope at the calibration marks. They also provide isolation to prevent loading of the C and L. All inputs and outputs of the monitor circuit are at ground potential. If DC isolation from the oscilloscope circuits is required, blocking capaci-

tors or transformers may be inserted in place of the jumper wires.

If you look at the schematic diagram in **Figure 4**, you'll note that there are no inductor symbols. The inductor function is provided by a circuit called a gyrator.¹ Operational amplifiers U1B and U1C, with their associated resistors and capacitors, comprise a synthetic inductor whose value is adjusted by R8. A second gyrator is used in the discriminator circuit.

Construction

You can install the discriminator board and its power transformer in a box outside the oscilloscope if there's not enough room inside the 'scope, or if you don't want to remove parts. When installing the circuit board, you *must* provide access to the potentiometer adjustments.

If the power transformer won't fit inside the 'scope cabinet, there may be room on the back of the cabinet. Transformer terminals must be covered for safety. If the 'scope already provides the required stable voltages, (± 9 to 15 volts DC), it won't be necessary to install power supply components for the X-display.

Signal leads should be shielded to prevent hum pickup.

When soldering components to the circuit board (see **Figure 5**), be careful not to overheat semiconductors and filter capacitors. The regulators don't require heat sinks. We recommend that you use 14-pin DIP sockets. After soldering, inspect your work carefully for solder bridging between adjacent pads and leads.

Calibration

Place symmetrical calibration marks on the edges of the CRT screen, identifying the ends of the legs of the X. Select a conveni-

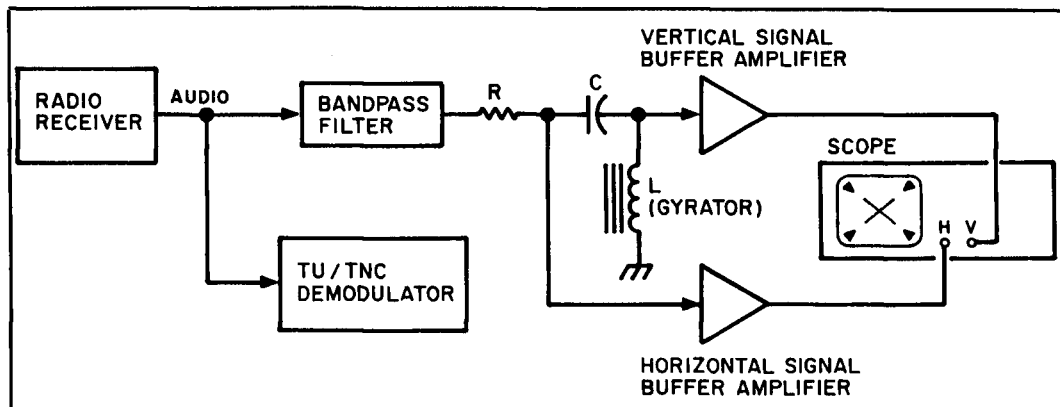
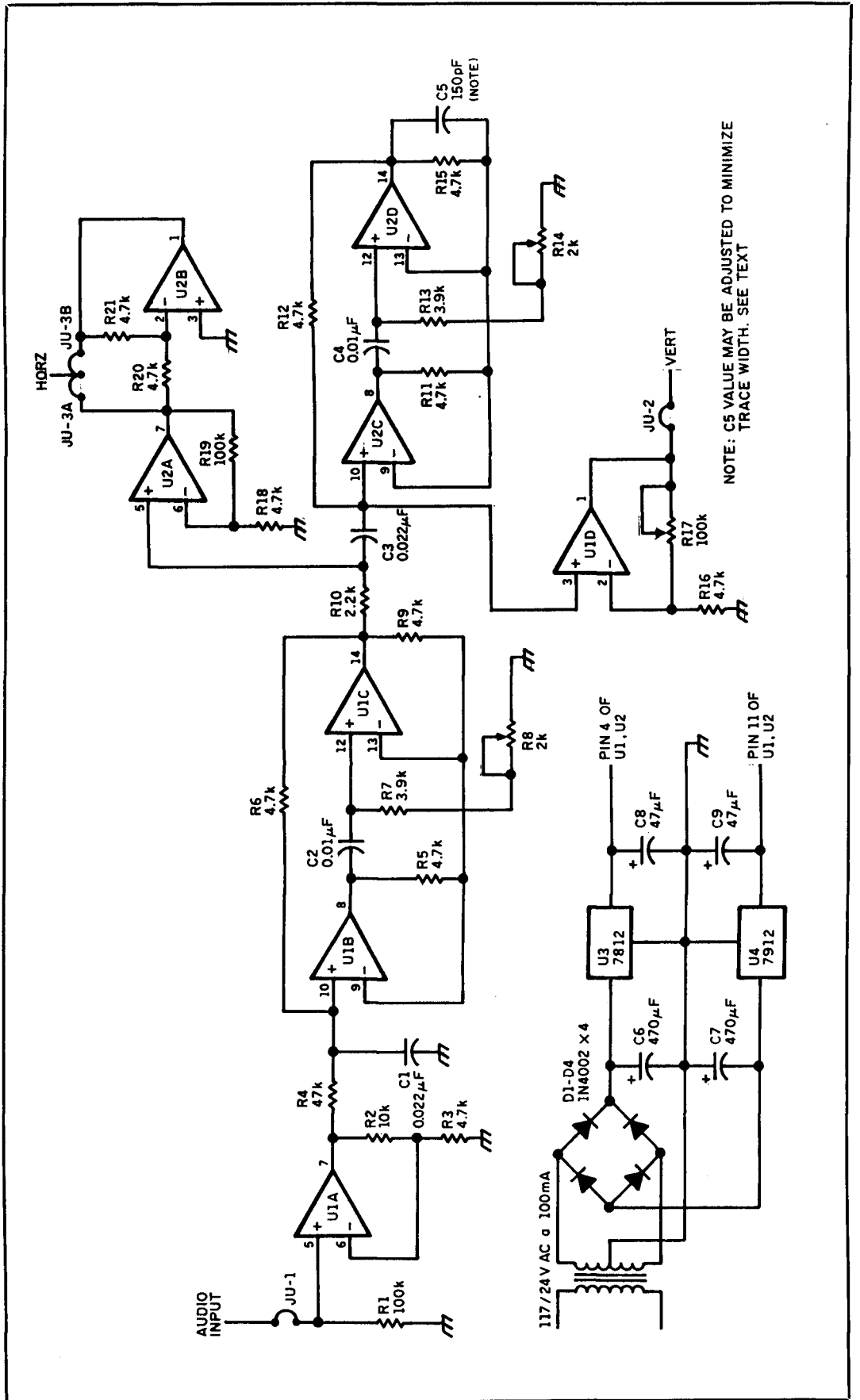


Figure 3. FSK monitor system block diagram.



NOTE: C5 VALUE MAY BE ADJUSTED TO MINIMIZE TRACE WIDTH. SEE TEXT

Figure 4. Schematic diagram.

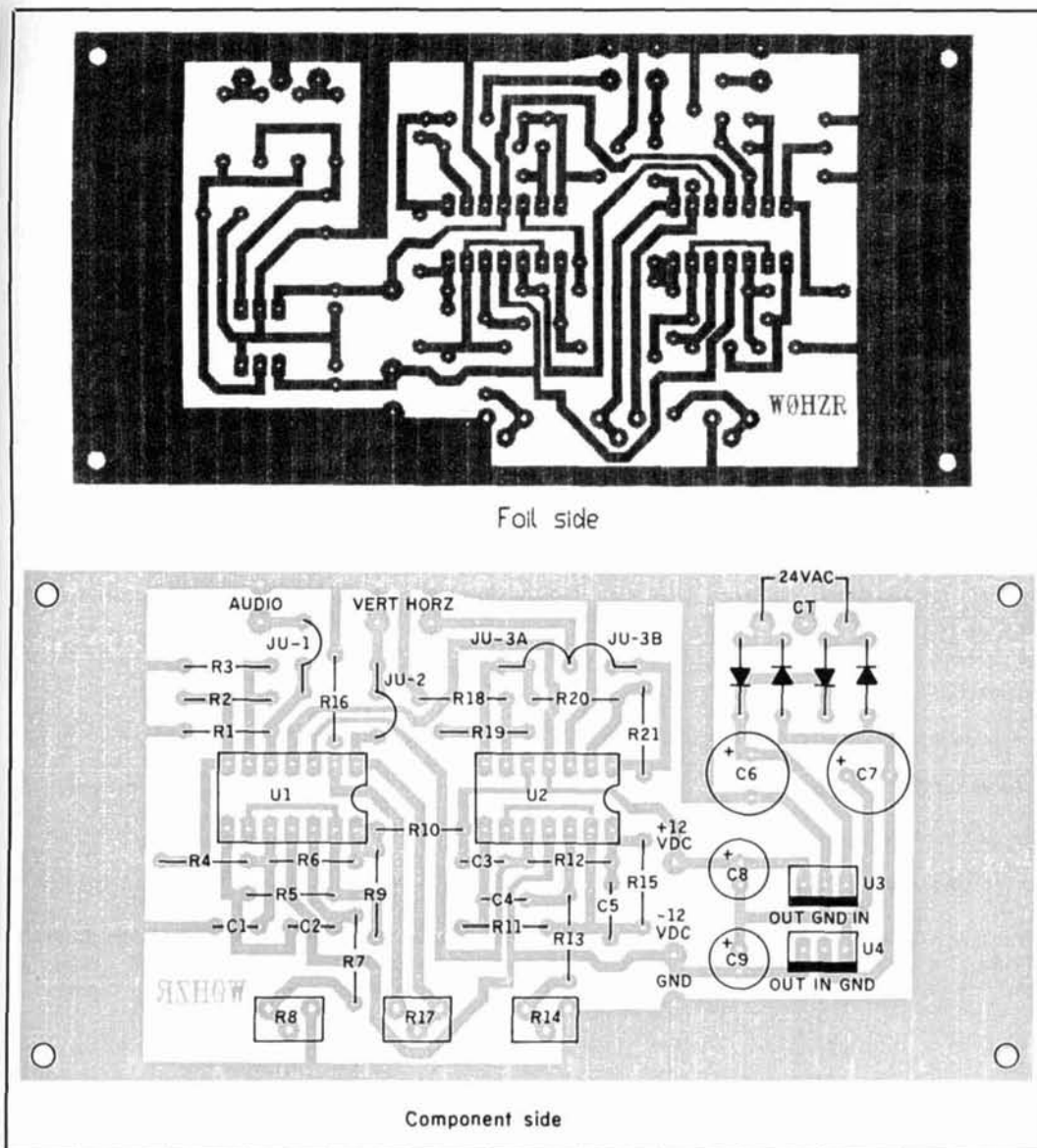


Figure 5. PC board layout and component placement.

ent angle like 60 or 90 degrees for the X. Pieces of electrical tape cut in the shape of small arrow heads are suitable.

After connecting the FSK monitor circuit to the vertical and horizontal inputs of the selected oscilloscope or foundation unit, apply power and check for proper voltages. Next, provide an audio input at a frequency midway between the marking and spacing frequencies selected. *Make sure the audio does not overload the monitor circuit.*

While observing the signal as a line on the CRT screen, adjust R8 for maximum line length and R14 for a vertical trace.

Finally, apply the marking and spacing frequencies alternately to the input of the monitor and align their respective traces with the calibration marks on the CRT

screen. The easiest way to do this is by running the horizontal gain "wide open" and adjusting R17 (vertical gain) and R14 (null frequency) until the marking and spacing signal traces line up with the calibration marks. (Note that R17 controls the angle between the traces while R14 rotates both traces in the same direction).

If one leg of the X appears elliptical rather than linear, adjust the value of C5 to collapse the ellipse.

If there is insufficient horizontal gain to fill the screen without distortion, you may need to swap the oscilloscope's horizontal and vertical amplifiers. This will rotate the traces 90 degrees. To correct the undesired rotation, you may either rotate the CRT 90 degrees in its mounting or exchange the ver-

tical with the horizontal deflection plate connections.

Operation

When tuning a radio to a digital mode signal, an operator usually listens for a characteristic pitch or pattern of sound, and then refers to a tuning indicator. Using the X display, you'll note that the X is compressed and leaning until it's almost horizontal until you tune the receiver to bring the X upright. It takes only two or three seconds to tune the radio correctly, even if you're not listening to the signal.

To evaluate the amount of shift observed, first compare the angle of the FSK signal traces with the angle of the calibration marks. Assuming that you used 170-Hz shift calibration, a signal that slightly exceeds the angle of the calibration marks is probably 200-Hz shift. A signal that greatly exceeds that angle is likely to be either 425 or 850-Hz shift. To measure any shift pre-

cisely, you need a receiver with precise tuning dial calibration. After aligning one leg of the X formed by the incoming signal with the closest calibration mark, note the frequency of the received signal. Then retune the receiver until the other leg of the X is aligned with the same calibration mark, and note the receiver frequency again. The shift is the difference between the two dial readings.

Assistance

If you find it difficult to obtain components or to get your X-display conversion to operate, feel free to contact W0HZR by telephone (no collect calls) at (612)881-2909. Please send a self-addressed stamped or equivalent with all written inquiries.

REFERENCE

I. A. Antoniou, "Realization of Gyration using Operational Amplifiers, and Their Use in RC-Active Network Synthesis," *Proceedings of the IEEE*, 1969, Volume 116, pages 1838-1850.

Parts List

Component	Value	Digi-Key Part #	Description
C1,C3	0.022 μ F	P4517	Panasonic V-Series
C2,C4	0.01 μ F	P4513	Panasonic V-Series
C5	150 pF	P4026	Panasonic Ceramic
C6,C7	470 μ F 35WV	P6255	Panasonic SU-Series
C8,C9	47 μ F 25WV	P6238	Panasonic SU-Series
D1-D4	1N4002, 100V	1N4002	(1N4001, 50 V okay)
R1,R19	100k	100KQ	
R21	10k	10KQ	
R3,R5,R6,R9,R11,R12,			
R15,R16,R18,R20,R21	4.7k	4.7KQ	
R4	47k	47KQ	
R7,R13	3.9k	3.9KQ	
R8,R14	2k pot	3FA23	Panasonic 3/8 square cermet pot.
R10	2.2k	2.2KQ	
R17	100k pot	3FA15	Panasonic 3/8 square cermet pot.
U1,U2	TL084, LF347,	LM348	
U3	LM7812	LM7812	National Semiconductor
U4	LM7912	LM7912	National Semiconductor

All resistors 1/4 watt.

Capacitors 50 WV unless noted.

Mounting holes 11/64 inch for no. 6 machine screw.

Holes for diodes, pots, U3,U4, and wires are 0.040 inch.

All other holes 0.030 inch.

Note: The use of the LM324 operational amplifier must be avoided because it exhibits crossover distortion. Many people may assume that the LM324 is an acceptable alternative to those op-amps listed; however, it is not.

THE SOLAR SPECTRUM

Update on the VLF receiver

This issue I thought a change of pace from our usual subject matter might be in order. Because so many of you have expressed an interest in our efforts to collect data on sudden ionospheric disturbances (SIDs), I've asked our Technical Coordinator Arthur Stokes, N8BN, to describe his new, special-purpose VLF receiver design. We use these devices in our program to detect solar flares, virtually as they occur, by monitoring the effects such events have on the ionosphere (**Photo A**)*.

An improved VLF receiver for solar flare monitoring

For quite some time, I've considered various ways to improve the selectivity of the VLF receiver circuit we've used during the past few years. Although this receiver has proven to be very effective and practical for many of our contributors, it does have a fairly wide band-pass that can result in some degradation of the record when the data are displayed or plotted.

A few months ago, I ran across an article written by Eugene Pearson, W3QY¹, in which he described a vacuum tube-based receiver with sharp tuning characteristics. Pearson's intriguing piece prompted me to modify our current receiver using the resonant circuits described in his article.

In the past, our circuit has employed two Miller 6319 or 9006 slug-tuned inductors shunted by 0.001 μF capacitors for the tuning selector. The Pearson circuit uses 88-mH toroid inductors shunted by air variable tuning capacitors. For our application, we chose three of these units in separate stages of amplification. Our conventional three-transistor circuit was

easily modified to accommodate these tuned sections. The schematic for the new design is shown in **Figure 1**. A parts list is provided.

Those of you who aren't "old timers" in the amateur radio game may wonder what these toroid inductors are. It's my understanding that these devices (**Photo B**) were originally used as loading coils on telephone transmission lines. For many years amateur radio operators used them for tuned circuits in radio teletype communication. In that relationship, toroid inductors were used to select the mark and space tones that operate teletype printers. Of course, with the development and ready availability of small personal computers, this technology is no longer practical.

My modified receiver is housed in a 3 x 6 x 8-inch metal cabinet from Radio Shack. The three tuning capacitors are mounted, equally spaced, across the front panel. The gain control is also attached to this panel.

The remaining components are installed on a section of perfboard trimmed to fit inside the metal cabinet (**Photo C**) and attached to the cabinet base with $\frac{1}{2}$ -inch stand-off spacers. I found it convenient to run two pieces of no. 14 bare copper wire, spaced about $1\frac{1}{4}$ inches apart, along the underside of the perfboard. These become the +12-volt and ground bus that connect all the components.

The three transistors are placed toward the rear of the perfboard. They are separated from each other as much as possible to reduce the likelihood of feedback from the output transistor back to the antenna input. While placement of most of the components isn't critical, the output portion of the circuit should be located as far away from the antenna input as possible. The four resistors associated with each transistor are mounted close to the transistor in question. The resistors are all $\frac{1}{2}$ watt.

The power transformer should be mounted on the back panel. The antenna BNC connector

*For additional information concerning this interesting and scientifically productive project, see "Recording Solar Flares Indirectly," in the Summer 1991 issue of Communications Quarterly.

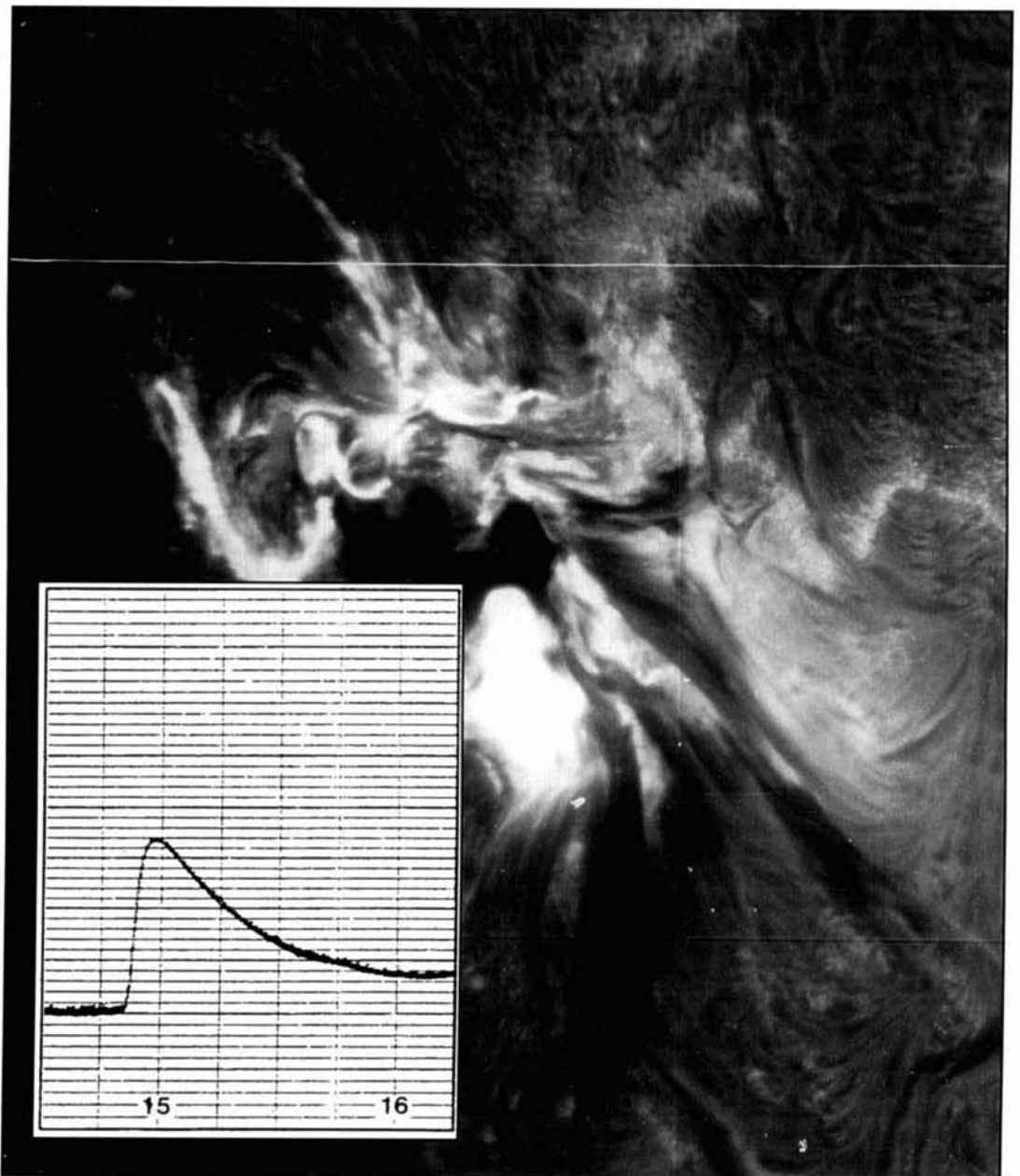


Photo A. Solar flares, such as this event which occurred during March 1989, strongly influence the radio propagation capabilities of the ionosphere. When very low frequency radio signal variations are recorded and displayed graphically, these atmospheric anomalies—known as sudden ionospheric disturbances—produce the distinctive signature shown in the inset.

is also installed on the back panel close to the input transistor. All leads should be kept as short as possible. The toroids are placed on the perfboard, adjacent to each of the tuning capacitors.

This particular circuit was designed to drive a 100 μ A Rustrak strip-chart recorder, rather than output its data to a small computer as described in our original article.² However, if a 5-k resistor is placed across its output terminals, the receiver can also be connected to an A-to-D converter and data recorded with a computer.

Obtaining the tuning capacitors and the

toroid coils will probably pose the most difficult task for anyone who wishes to build this receiver. These are not off-the-shelf items from Radio Shack! Amateur radio “flea-markets” might be a good source, although in my case the tuning capacitors were salvaged from several old broadcast radios. Fortunately, many of us have radio amateur friends that maintain a large store of “junk-box” parts, and they may be the best source of all.

I would not advise those who lack experience building electronic equipment to attempt this project. Because the tuning capacitors and the coils may well vary, it will almost certainly

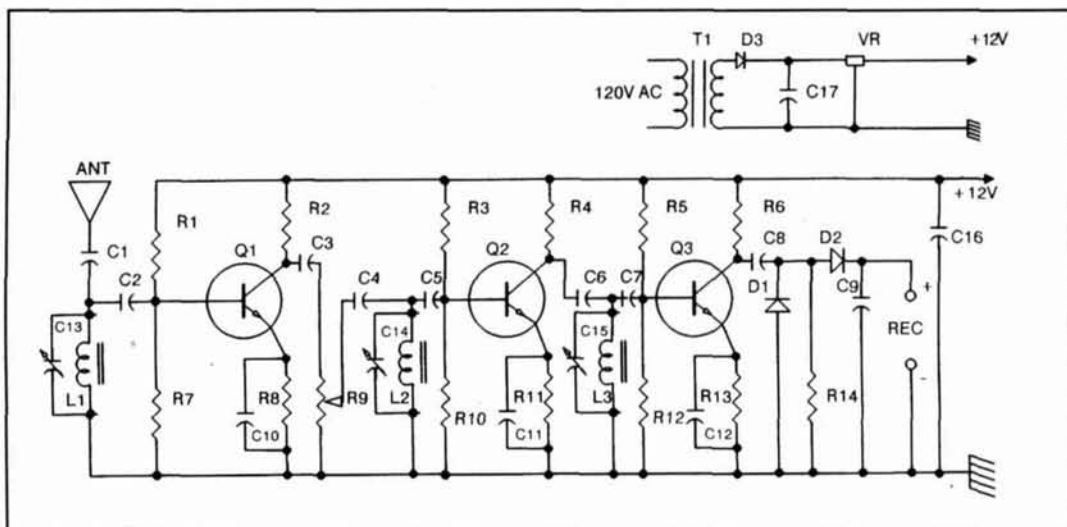


Figure 1. Schematic drawing for the new version of the VLF receiver.

be necessary to use a signal generator and an oscilloscope when adjusting the tuned circuits to the proper frequency range. The final calibration of the dials should also be done with a signal generator. This circuit tunes very sharp, making it somewhat difficult to locate a station if the three tuned circuits aren't tracked reasonably close together.

When adjusting the tuning range of this receiver, you may find it necessary to add 200 pF or so in parallel with each of the tuning capacitors. Another way to achieve the same result would be to include toggle switches with each capacitor unit—allowing one to switch a capacitor in or out of the circuit. Such a technique could provide both a high and low range of available frequencies—an additional benefit for some observers.

The final step in the tuning process involves calibrating the three tuning dials. This is done by placing a known frequency from a signal generator into the antenna connector through a 100-k resistor. An oscilloscope is then attached to the collector of the third transistor, each of the capacitors is adjusted for maximum output signal, and the dials are marked with the corresponding frequency.

As we have indicated, this receiver tunes quite narrowly. For example, 100 Hz off center frequency drops the recorder output to near zero. The tuning range of my system begins at approximately 20 kHz and ends at a bit over 30 kHz—the prime range of frequencies for recording solar flare-induced atmospheric anomalies.

I've had excellent results with these systems using several different types of antennas. One that has worked very well for me is a 40-meter inverted dipole with about 50-feet of RG-8U coaxial cable leading to the receiver. In this sit-

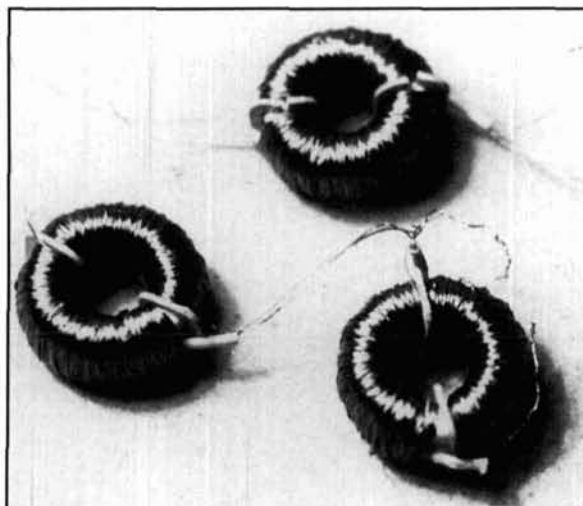


Photo B. Toroid inductors may have originally been intended to be used as loading coils on telephone transmission lines.

uation, however, I didn't use the standard method of coax connection. Instead, I tied the coax sheath and center wire together and attached the pair to the center pin of the receiver's BNC antenna connector. The receiver chassis was then solidly grounded. I've also obtained excellent results using a 24-foot section of 1-inch diameter aluminum tubing, erected vertically, as an antenna. The tubing is insulated from the ground using suitable nonconducting spacers, and connected to the center conductor of coax. The coax sheath is grounded in the usual manner.

We have also successfully employed small, tuned-loop antennas for this project. This is an especially attractive solution for those of us who are unable to construct an exterior unit. However (as might be expected) exterior anten-

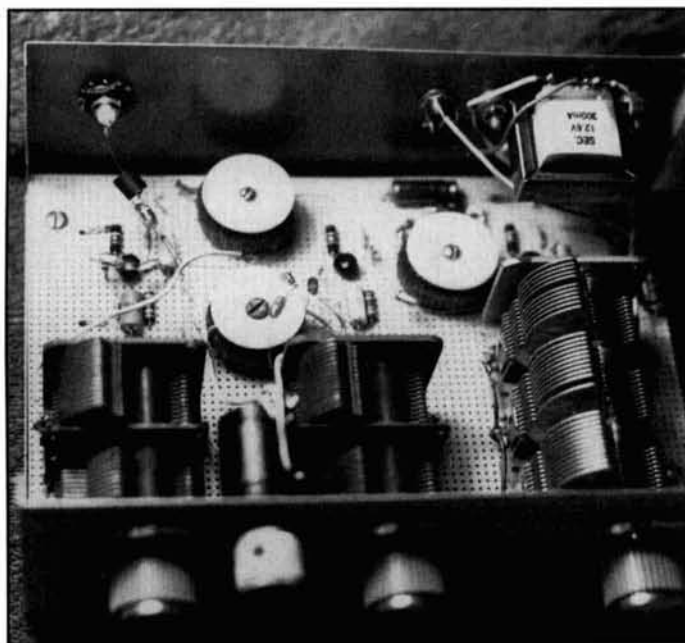


Photo C. The suggested parts arrangement for the Stokes receiver.

nas like those I've described will generally provide the most satisfactory results.

(Note: Mr. Stokes would be happy to hear from anyone who assembles a flare-detection system that uses this circuit, or who knows of a

*Arthur Stokes, N8BN, P.O. Box 398, Hudson, Ohio 44236.

good source for the toroid coils and/or tuning capacitors.)*

Recent solar activity and short-term outlook³

The Sun was relatively quiet in December and January, becoming more active during February. Just six class M solar flares were recorded during the first two months of the period, half the number that were detected in the first half of February. One of the more interesting of these events was the occurrence of the first major flare since early November—a complex M9.6/2B Tenflare.

The geomagnetic field also became more active during February, with occasional reports of severe storm conditions at some high-latitude stations. These disturbances have been linked to several sources—primarily solar flare activity and coronal hole effects.

During the next three months activity is expected to decline, although scattered intervals of high activity will almost certainly occur. Geomagnetic activity should intensify over the short-term in accordance with the arrival of the vernal equinox. ■

REFERENCES

1. Eugene Pearson, W3QY, "A WWVL Receiver," *QST*, November 1971.
2. Peter Taylor and Arthur Stokes, "Recording Solar Flares Indirectly", *Communications Quarterly*, Summer 1991, page 29.
3. A portion of this information was obtained from the *SELDADS* data base.

Parts List: VLF Receiver

C1, C2, C4, C5, C6, C7	100-pF capacitors
C3, C8, C11, C12	2.2- μ F capacitors
C9	200- μ F capacitor
C10	0.1- μ F capacitor
C13, C14, C15	360-pF variable capacitors
C16, C17	100- μ F capacitors
D1, D2	1N34 diodes
D3	1N4001 diode
L1, L2, L3	88-mH toroid inductors
Q1, Q2, Q3	2N4401 transistors
R1, R3, R5	100-k resistors
R2, R4, R6, R7, R10, R12, R14	10-k resistors
R8, R11, R13	470-ohm resistors
R9	10-k potentiometer
T1	12-volt transformer
VR7812	12-volt regulator

BUILDING THE PERFECT NOISE BRIDGE

Construction tips for series and parallel R-C bridges

The noise bridge is a very useful piece of test equipment for the radio amateur.

One of the reasons for its rather limited use by the amateur community may be its lack of accuracy—especially that of commercially available units. A good noise bridge can serve as an impedance bridge, a capacitance bridge, and an inductance bridge.

There have been several excellent articles about noise bridges^{1,2,3} that were mostly concerned with the manufacture of a balanced transformer for the parallel R-C noise bridge. Although the transformer is the most critical part of a noise bridge, it's also necessary to pay attention to the influence of the connector capacitance and the distributed parameters of the variable resistor. This is especially impor-

tant for the series R-C noise bridge, but these factors have never been considered in the past. In the parallel bridge, only the variable resistor influences accuracy—assuming a perfect transformer.

If you consider the large frequency range of 1.8 to 28 MHz and the relatively large impedance range (zero to several hundred ohms), making a good noise bridge may appear to be a difficult task, but with the recommendations given here, you can build an excellent and sufficiently accurate instrument. Some of the improvements can be applied to existing bridges, thereby increasing their usefulness.

In the past, the main problem has been the calibration of the resistance and reactance dials together with the lack of knowledge of the per-

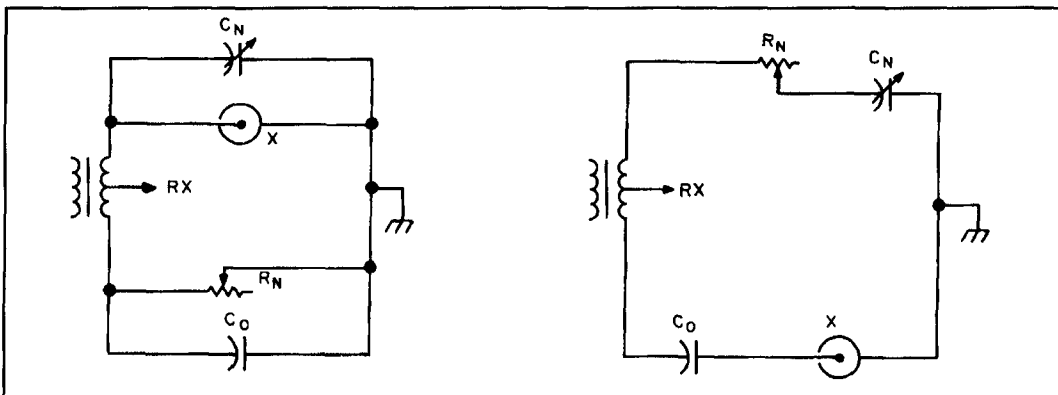


Figure 1. The parallel R-C and series R-C noise bridge configurations.

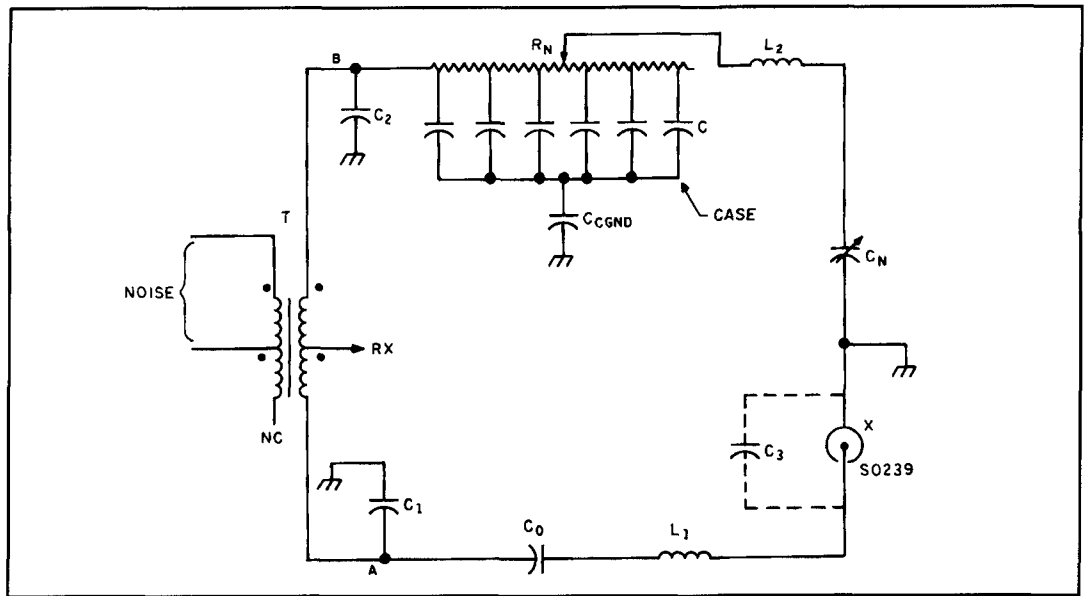


Figure 2. Typical schematic of a series R-C noise bridge.

formance limits of the noise bridge. I will point out and analyze the error sources of the two noise bridge types and calculate their intrinsic errors. This will enable you to build a series R-C noise bridge that reaches the theoretically possible accuracy and a parallel R-C bridge whose total error is below 1 percent—a fraction of the series-bridge error. As a result, the parallel R-C bridge becomes a precision bridge with its error determined only by mechanical factors like mechanical backlash.

The two types of noise bridges

There are two configurations, the parallel R-C and the series R-C bridge (see Figure 1).

Figure 1 shows that the R-measuring range is the same in both cases. The series bridge can measure capacitances up to infinity, while the parallel circuit can only handle values smaller than C_0 , the reference capacitor. However, the latter can measure small capacitances more accurately. The main drawback of the series circuit is that the capacitance of the S0239 connector lies in parallel to the unknown impedance. This causes errors, especially when the unknown impedance is high and has a reactive component. In the parallel bridge, this capacitance lies in parallel to the variable capacitor and has no effect. One advantage of the series circuit is that its results are presented in series notation, which are conveniently applied to transmission line formulas and the Smith Chart.

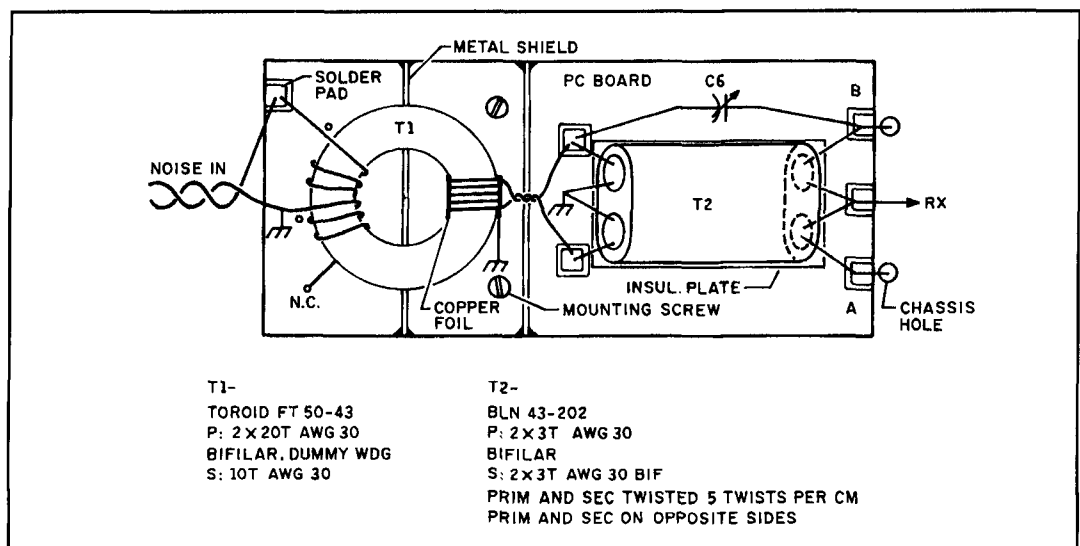


Figure 3. Transformer assembly.

Despite their lesser accuracy, all commercial bridges are of the series type. For this reason, it is desirable to improve their performance.

Actual series R-C noise bridge schematic

Figure 2 shows the typical schematic of a series R-C noise bridge. The transformer, T, serves as a balanced noise source for the bridge. The center tap of its secondary winding is connected to an RF receiver that serves as a null indicator. The tuning of the receiver determines the measuring frequency. The variable bridge elements are adjusted for noise minimum in the loudspeaker.

To analyze the noise bridge circuit, consider the unavoidable wiring capacitances C_1 and C_2 , wiring inductances L_1 and L_2 , the inherent capacitive characteristic of the variable resistor R_N , and the capacitance, C_3 , of the S0239 connector.

If you assume an ideal transformer, T, you can only expect error-free operation of the noise bridge if there is complete electrical symmetry between upper and lower bridge legs, as seen from points A and B. Cursory inspection of **Figure 2** shows that this perfect symmetry can never be obtained with the series R-C configuration; however you can improve it far beyond its present standard, to such a degree that even a critical user will be satisfied. Consequently, the two main tasks are to provide a transformer that is symmetrical and free of capacitive cross talk between primary and secondary, and to establish as much circuit symmetry as possible.

The transformer

The transformer is the most critical part of a noise bridge. The ideal transformer should have only magnetic coupling between primary and secondary, without capacitive cross talk between the windings—either directly or via the magnetic material of the iron core that carries

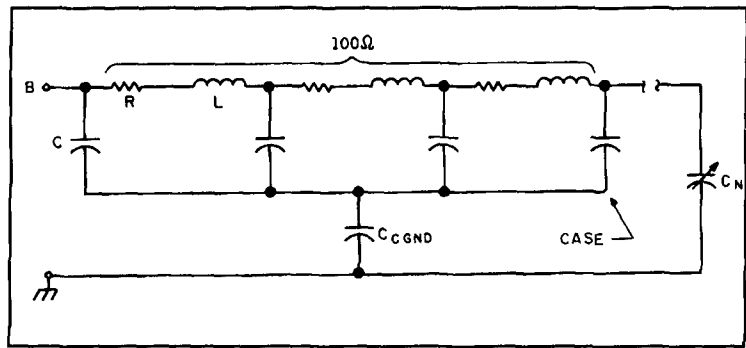


Figure 4. If you measure the capacitance between the wafer and potentiometer case, you will encounter values between 6 and 30 pF. There is also inductance present.

noise potential. A single transformer can never fulfill this requirement; a second transformer, T1, is needed to provide a ground-symmetrical noise source for the actual bridge transformer, T2. Much work has been done in this area and one of the best solutions seems to be the transmission line-type transformer with quadfililar windings and without an iron core. However, the transformer is difficult to wind and can, because of its low impedance, induce noise currents in the metal chassis that cause a shifting of the null settings of the variable elements.

The acid test of a noise bridge is that the reversal of the noise input leads should not affect the bridge readings at any frequency. If there is no change, this proves not only that the transformer is nearly perfect, but also that there are no noise currents flowing in the metal chassis. Here, all noise bridges are deficient to a large degree. It is useless to try to build or improve a bridge if this requirement is not met. I have observed differences in readings of 10 percent when noise leads were reversed. However, this does not mean that the bridge has a 10 percent error, because the dial calibration was made individually and certainly in the presence of other imperfections. The dual transformer assembly, shown in **Figure 3**, is easy to build and renders excellent results.

The capacitive cross talk between the windings can be reduced with bifilar windings and

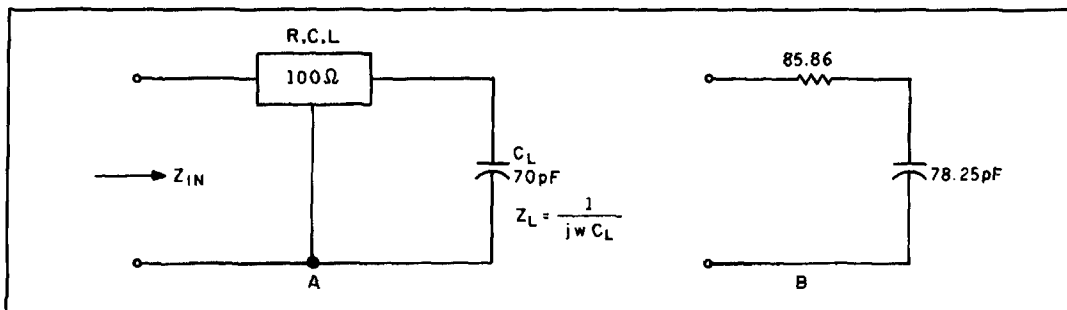


Figure 5. (A) "Transmission line" terminated with a load impedance of $Z_L = 1/j\omega C_L$, the variable capacitor C_N . (B) An 85.86-ohm resistor is added in series with a 78.25 pF capacitor.

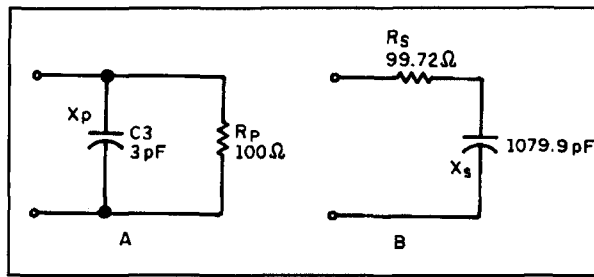


Figure 6. Direct connection to the S0329 connector for (A) the parallel R-C circuit and (B) its series equivalent.

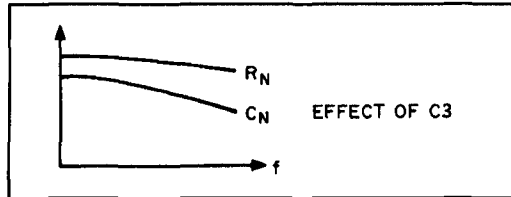


Figure 7. The effect of C3 on the settings of R_N and C_N .

an unused dummy winding at the primary side, but this measure alone is insufficient. Transformer T2 uses a two-hole ferrite core with the two terminals on opposite sides, as recommended by Grebenkemper. Noise balancing transformer T1 uses a toroidal core. Its primary winding is fed from the noise generator. The secondary is placed on a one-turn, non-shorting copper foil which functions as an electrostatic shield that prevents capacitive noise pickup from the "hot" core material. It is located on the opposite side of the primary. A non-shorting metal shield is placed between primary and secondary. The copper foil is grounded at the small pc board that serves as a mounting base for the transformer assembly. The ferrite core of T2 should not be grounded and should be glued to the pc board via a pad of insulating material. A small pc board between the transformers serves as an electrostatic shield. A hole is provided for the twisted-wire connection between the transformers. Points A and B are the connections to the bridge elements and are preferably mounted on the other side of the chassis to prevent capacitive coupling.

The variable resistor in the series R-C bridge

A potentiometer is not a pure resistance. If you measure the capacitance between the wafer and potentiometer case, you will encounter values between 6 and 30 pF, depending on the model. It is obvious that this must have a considerable effect at 28 MHz. In addition, there is also inductance present (see Figure 4).

Because all three parameters, R, C, and L, are distributed, you may consider the potentiometer a transmission line and apply the transmission line formulas. In our example, the carbon layer of the 100-ohm potentiometer has a length of 0.0523 m and the resistance parameter is therefore:

$$R = \frac{100}{0.0523} = 1912.046$$

ohms per meter. Assuming a capacitance of 12 pF between wafer and case, the capacitance parameter is

$$C = \frac{12}{0.0523} = 229.44 \text{ pF}$$

per meter. The L parameter is 240×10^{-9} H per meter. G can be assumed to be zero. Because the potentiometer lies in series with the variable capacitor, C_N , you must consider the whole upper part of the bridge, neglecting the effect of inductance L2 for the moment (see Figure 5).

Figure 5A shows the "transmission line" terminated with a load impedance of $Z_L = 1/j\omega C_L$, the variable capacitor C_N . The frequency is 28 MHz. The formula for the input impedance of a transmission line terminated with Z_L is shown in Equation 1 below.

The result is shown in Figure 5B, a 85.86-ohm resistor in series with a 78.25-pF capacitor. If the variable resistor is set to 100 ohms and the variable capacitor to 70 pF, the actual values that determine bridge balance are 85.86 ohms and 78.25 pF. In other words, the potentiometer causes R_N readings to be too high and

$$Z_{IN} = R + jX_{IN} = \frac{Z_L + Z_0 \tanh(\alpha + j\beta) l}{1 + \frac{Z_L}{Z_0} \tanh(\alpha + j\beta) l} \quad (1)$$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \quad G = 0 ; l = 0.0523$$

C_N values to be low. Calculations show that the effect of the inductance is very small; the wafer capacitance is the dominating element. If one potentiometer terminal is grounded, as with the parallel R-C bridge, inductance does play a role, as will be shown later.

Mount the potentiometer using a phenolic shaft extension rather than attaching it directly to the front panel. (You want to avoid grounding the potentiometer housing.) With the isolated mounting, you can obtain a capacitance C_{CGND} of 1.5 pF. This reduces the error to a negligible value. A good potentiometer would be one with about 10 pF capacitance from wafer to case.

The S0239 connector

As stated before, the main accuracy-limiting element of the series R-C noise bridge is the connector. Replace the connector with a capacitance to obtain a good approximation for load resistors above a few ohms. A suitable connector may have about 3 pF capacitance. If you want to measure, for instance, a noninductive 100-ohm resistor, you may connect it directly to the S0239 connector, or via a PL259 connector. In this case, the total capacitance, C_3 , is 5.7 pF (3 pF for the S0239 and 2.7 pF for the PL259 connector). In **Figure 6**, I assume direct connection to the S0239 connector. The frequency is 28 MHz.

If you convert the parallel R-C circuit into its series equivalent with the formulas:

$$R_s = \frac{R_p X_p^2}{R_p^2 + X_p^2} \quad (2)$$

$$X_s = \frac{X_p R_p^2}{R_p^2 + X_p^2} \quad (3)$$

the result is 99.72 ohms in series with 1079.9 pF, as shown in **Figure 6B**. Since this capacitor lies in series with the reference capacitor $C_0 = 70$ pF, the total capacitance in the lower bridge leg is 65.74 pF. In other words, the variable capacitor in the upper leg must be set to 65.74 pF instead of 70 pF and the variable resistor to 99.72 ohms instead of 100 ohms. Both R_N and C_N settings are too low due to C_3 . **Figure 7** shows that this error increases with a rise in frequency.

If you connect the 100-ohm resistor via a PL259 connector, the error increases. If you want a more accurate evaluation of the S0239 connector, think of it as a short piece of transmission line, as with the potentiometer. Calculations show that at very small load resistance, the S0239 connector becomes inductive—the dominating effect of the center conductor.

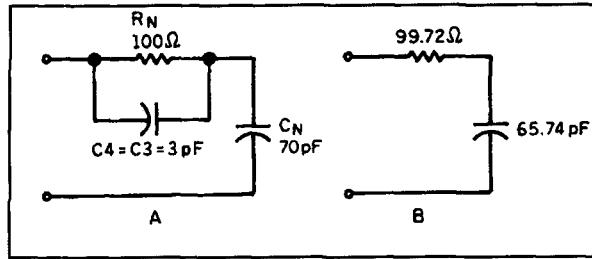


Figure 8. To have electrical symmetry between the two branches of the bridge, place capacitor C_4 , equal to capacitor C_3 , in parallel to the variable resistor, R_N , as in (A). The equivalent circuit of (B) shows that both values are lower than the correct ones of 100 ohms and 70 pF.

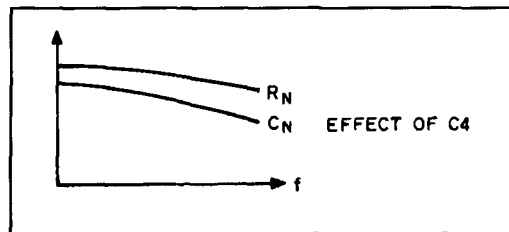


Figure 9. The effect of C_4 . The error of **Figure 8** increases with frequency.

lations show that at very small load resistance, the S0239 connector becomes inductive—the dominating effect of the center conductor.

Error correction for the series R-C bridge

As I have pointed out before, it is mandatory to have electrical symmetry between the two branches of the bridge. An excellent way to accomplish this is to place a capacitor, C_4 , equal to capacitor C_3 , in parallel with the variable resistor R_N , as shown in **Figure 8A**. The frequency is 28 MHz.

The equivalent circuit of **Figure 8B** shows that both values are lower than the correct ones of 100 ohms and 70 pF. This error also increases

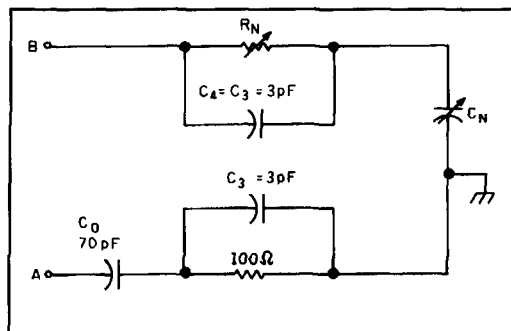


Figure 10. The two sections of the error-compensated noise bridge, assuming you want to measure a 100-ohm resistor.

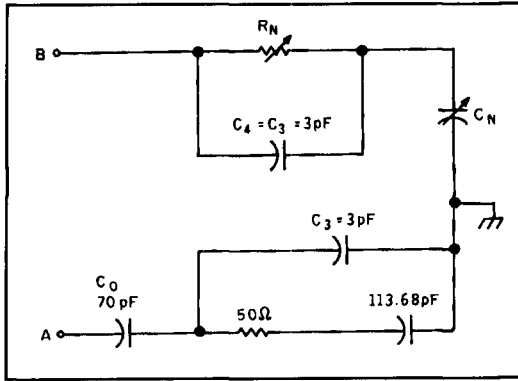


Figure 11. Series connection of a 50-ohm resistor with a capacitive reactance of 50 ohms, measured at 28 MHz.

es with frequency, as shown in Figure 9. Because both circuits are on opposite sides of the bridge, an error canceling effect takes place.

Now you have perfect symmetry, except for the distributed capacitance of the potentiometer. Calculations and measurements show that the R_N error caused by the potentiometer tends to cancel the connector-caused R_N error. However, the C_N errors of both bridge elements do add together. Therefore, the potentiometer cannot be used as an error-reducing device. It is better to make the potentiometer-related errors as small as possible and to rely on the compensating capacitor, C_4 , entirely for error compen-

sation. By making C_4 variable, you can significantly decrease the bridge error over the entire frequency range.

Figure 10 shows the two sections of the error-compensated noise bridge, assuming that you want to measure a 100-ohm resistor. The compensating capacitor, C_4 , equals the connector capacitance, C_3 . Because of electrical symmetry, there is perfect agreement between R_N and the unknown resistor at all frequencies. R_N must be set to 100 ohms and C_N to 70 pF for bridge null.

However, this is *only* true if $C_3 = C_4$, and if the unknown impedance is a pure resistance. If $C_3 = C_4$, but there is reactance, you'll have an error that increases with the increasing value of the total impedance.

A practical case may be the series connection of a 50-ohm resistor with a *capacitive* reactance of 50 ohms, measured at 28 MHz, and shown in Figure 11.

A 113.68 pF capacitor provides a capacitive reactance of 50 ohms at 28 MHz. If you convert the parallel connection in the lower leg into its series equivalent and add to it the series capacitor, C_0 , the residual resistance is 47.43 ohms and the total capacitance is 43.34 pF. Neglecting capacitance C_4 for the moment, R_N must be 47.43 ohms and C_N must be 43.34 pF for bridge balance. If you consider C_4 , the final values are $R_N = 47.46$ ohms and $C_N = 43.74$

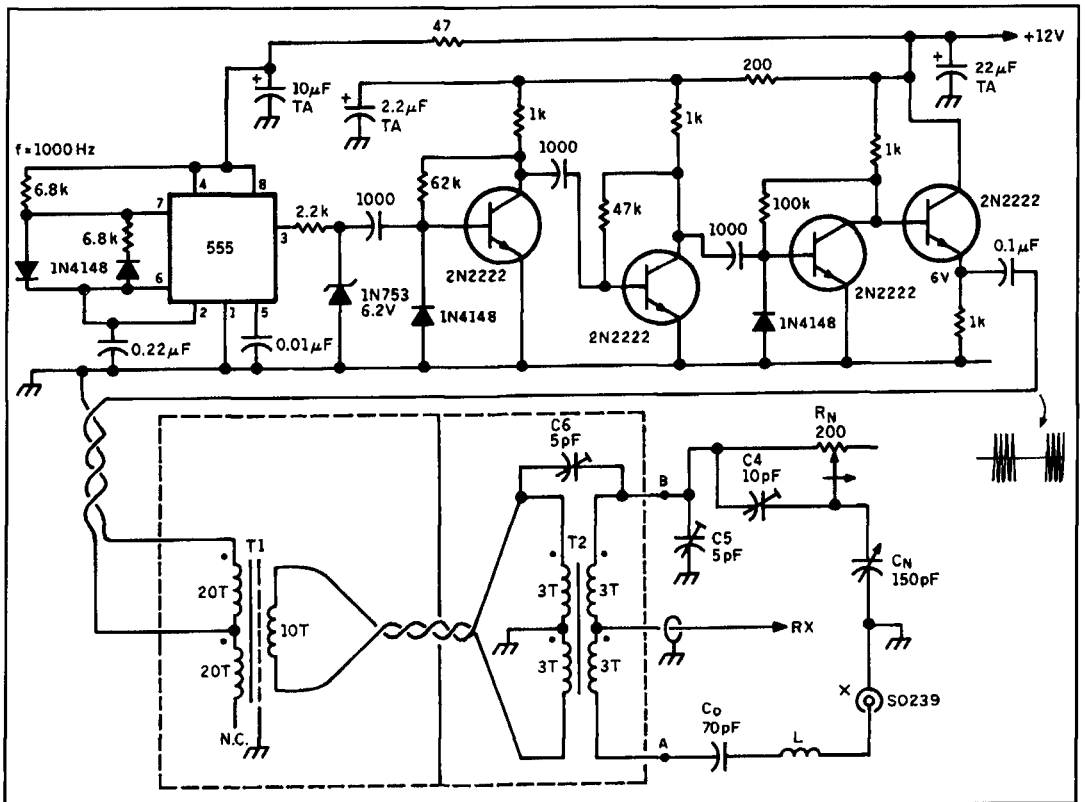


Figure 12. The final schematic of the compensated series R-C noise bridge.

pF. Without the connector capacitance, the correct values would be $R_N = 50$ ohms and $C_N = 43.32$ pF—the sum of the 113.68 and 70 pF capacitors. This corresponds to errors of 5.3 and 0.96 percent, respectively.

If the unknown impedance consists of a 50-ohm resistor in series with an *inductive* reactance of 50 ohms (as opposed to the capacitive reactance of **Figure 11**), the R_N error is about the same, but the C_N error is considerably larger, about 4.7 percent. The reason for this increase is that the electrical balance is degraded more because of the inductance in the circuit. It is interesting to note, and measurements confirm, that the noise bridge readings are frequency independent if the reactive part is capacitive, but vary considerably if the reactive part is inductive.

At lower frequencies, the errors are much smaller, since the error increases nonlinearly with frequency. For example, at 1.8 MHz and for a 50-ohm inductive reactance in series with 50 ohms, the errors are only 3.5 percent for R_N and 0.07 percent for C_N .

There is nothing that can be done to the series bridge to reduce this error, but it does go down as the reactive part decreases, and it is zero for a pure resistive load. As a result, when you try to improve an antenna system by reducing the reactive part of the antenna impedance, the bridge accuracy increases—a desirable effect.

For pure resistive loads, the bridge accuracy is limited by the precision of the dial markers, the lack of repeatability of resistor settings, or the possible backlash of reduction gears. I recommend that you align the bridge with three noninductive resistors of 0, 50, and 180 ohms. The bridge error will then be smallest at all other loads.

If you want to make a low-inductance resistor, solder a SMD resistor to a small strip of pc board material which is then soldered between the center conductor and the rim of a short PL259 connector of low capacitance.

The final schematic of the compensated series R-C noise bridge is shown in **Figure 12**. This figure shows the transformer assembly with transformers T1 and T2, balancing trimmers C5 and C6, and balancing inductor L (whose locations must all be found experimentally), along with compensating trimmer C4. The noise source is amplitude modulated; the frequency is about 1000 Hz. The operating point of the amplifier output stage should be adjusted for a symmetrical waveform at the input to transformer T1.

Noise bridge calibration

To calibrate your noise bridge, first reverse

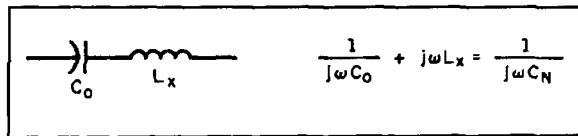


Figure 13. The schematic of the lower bridge leg where the unknown inductance, L_x , lies in series with the reference capacitor, C_0 .

the noise input leads to transformer T1. This should not affect the settings of the null position of R_N and C_N . If you have built the transformer assembly using **Figure 3**, there is no need for this test.

Next correct the capacitive imbalance of transformer T2. Disconnect the noise bridge elements from the secondary of T2 at points A and B, but leave the receiver connected to the center tap of the secondary winding. In most cases you will hear a strong noise signal in the loudspeaker. This is an indication of capacitive cross talk between the primary and secondary of T2. Depending on construction and wire position inside the ferrite core, you must connect a small trimmer capacitor (C6) between one secondary and one primary terminal. You must find its location experimentally, but if you make the correction, it is possible to completely eliminate the noise in the receiver. This compensation is also frequency independent.

Now measure the potentiometer resistance with a DVM and mark the resistance dial accordingly. You can also mark the C_N dial (reactance dial) in picofarads, using an accurate capacitance meter.

Next balance the parasitic inductances L1 and L2. Following Grebenkemper's recommendations, place a short circuit at the X terminal and set the R_N dial to zero. I recommend shorting the potentiometer terminals with a piece of wire because the zero position does not always have zero ohms. Tune the receiver to 1.8 MHz. Adjust the C_N dial for noise minimum. This is the null position and should be marked on the C_N dial. If the frequency is increased, the capacitance reading should stay constant. If the reading decreases with increasing frequency, the inductance in the variable leg is too high and you must add a small compensating inductance in series to the S0239 connector. This inductance may be just a small loop of wire.

Continue by balancing the parasitic capacitances of the bridge. It is preferable to do this by placing a 180-ohm resistor at the X terminal; the frequency is 28 MHz. If the reading is higher than 180 ohms, add to capacitance C1. If the reading is lower, add to C2. I recommend that you use a small trimmer capacitor for this purpose.

Finally, adjust frequency-compensating trimmer C4. Connect a noninductive 100-ohm

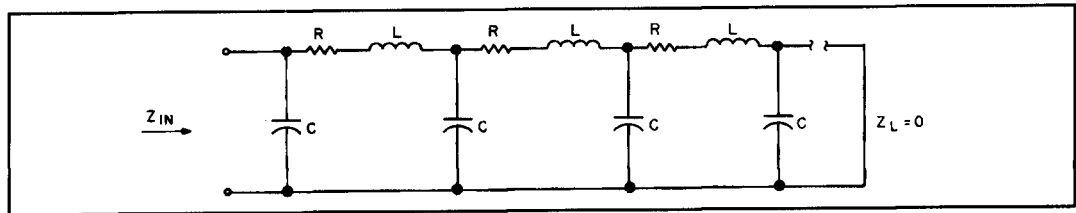


Figure 14. The equivalent circuit of the 218-ohm potentiometer.

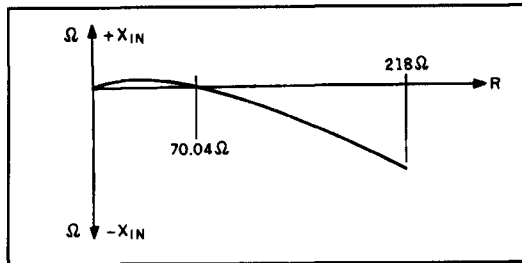


Figure 15. Plot of the reactive part of the input impedance versus the potentiometer value.

resistor at the X terminal. The frequency is 1.8 MHz. Null the bridge and increase the frequency to 28 MHz. If the capacitance reading decreases for bridge null, increase capacitance C4 until the dial reading stays frequency independent. You may repeat this test with other resistors.

How to measure inductances with the series bridge

Because the C_N dial is calibrated in picofarads, the inductive reactance $X_L = \omega L_x$ can easily be calculated from the dial readings. **Figure 13** is the schematic of the lower bridge leg where the unknown inductance, L_x , lies in series with the reference capacitor, C_O . Looking at **Equation 4**:

$$\frac{1}{j\omega C_O} + j\omega L_x = \frac{1}{j\omega C_N} \quad (4)$$

where C_N is the capacitance of the variable capacitor in the upper leg. If C_N is larger than C_O , the unknown impedance is an inductor. Its inductance reactance ωL_x is:

$$X_L = \omega L_x = \frac{1}{\omega C_O} - \frac{1}{\omega C_N} \quad (5)$$

If the unknown reactance is capacitive, its reactance is:

$$X_c = \frac{1}{\omega C_x} = \frac{1}{\omega C_N} - \frac{1}{\omega C_O} \quad (6)$$

$$X_L = \frac{10^{12}}{2\pi \times 70 \times 7 \times 10^6} - \frac{10^{12}}{2\pi \times 130 \times 7 \times 10^6} = 149.9\Omega$$

By way of a practical example, assume $C_O = 70$ pF, $C_N = 130$ pF, and $f = 7$ MHz.

Construction guidelines, series R-C bridge

1. Improve transformer symmetry with a pre-balancing transformer. Avoid chassis currents by using single-point grounding.
2. Minimize the effect of the distributed potentiometer capacitance by isolating it from the metal chassis, mounting it on a plate of insulating material, and using a phenolic rod as a shaft extension. Select a type with 10 pF or less capacitance between wafer and case.
3. Select a S0239 connector with a small capacitance of about 3 pF.
4. Simulate this connector capacitance with a variable capacitor (C_4) to establish complete circuit symmetry for resistive loads.
5. Cancel the existing capacitive imbalance between primary and secondary windings of transformer T2 with a trimmer capacitor (C_6) and a dummy winding on the primary side of transformer T1.
6. Balance the stray capacitances at points A and B with a variable capacitor (C_5).
7. Balance the stray inductances on the bridge with an added inductor (L).

Test results, series R-C noise bridge

If you follow the steps in the preceding section, the series R-C noise bridge will perform exactly as the calculations predict. Pure resistances are measured error free. For complex loads, there is a moderate, but predictable error in R_N and C_N that is larger if the reactive part of the unknown impedance is inductive. The worst-case error at 28 MHz for a resistor of 50 ohm in series with a capacitive reactance of 50

ohms is 5.3 percent for R_N and 0.96 percent for C_N . If the unknown impedance is 50 ohms in series with an inductive reactance of 50 ohms, the errors are 5.3 and 4.7 percent, respectively.

Worst-case errors, that cannot be corrected due to lack of bridge symmetry, show clearly the limitations of the series R-C noise bridge.

The parallel R-C noise bridge

As stated before, the series configuration of the noise bridge has two error sources: the connector and the variable resistor. The parallel circuit has only one, the variable resistor.

However, since one potentiometer terminal is grounded, its error contribution is much smaller, as will be shown in the next section.

The potentiometer in the parallel R-C noise bridge

The effect of the distributed capacitance is much smaller and the inductance becomes more important, whereas the latter played no role in the series bridge. The equivalent circuit of a 218-ohm potentiometer is illustrated in **Figure 14**.

To calculate the effect of the distributed parameters, use the transmission line formulas for the shorted line:

$$Z_{IN} = R_{IN} + X_{IN} = Z_0 \tan h n l (\alpha + j\beta)$$

$$l = 0.0523m$$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (7)$$

$$\alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$$

where l is the length of the carbon layer, n describes the wiper position (if $n = 1$, $R = 218$ ohms). The R parameter is now

$$R = \frac{218}{0.0523} = 4168 \text{ ohm/m.}$$

With a wiper-to-case capacitance of 7.8 pF, the C parameter is

$$C = \frac{7.8}{0.0523} = 149 \text{ pF/m.}$$

$L = 240$ nH/m and $G = 0$. If you plot the reactive part of the input impedance versus the potentiometer values, you have **Figure 15**. The frequency is 28 MHz.

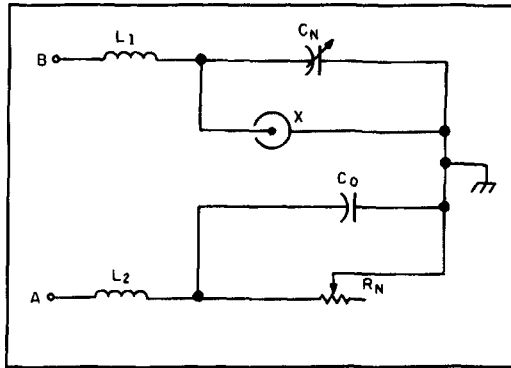


Figure 16. The bridge elements as they are connected to points A and B of **Figure 2**.

It is interesting to note that the input impedance is inductive up to a resistance of 70.04 ohms. Above this point, it is capacitive. At 70.04 ohms, it is purely resistive. Because the potentiometer lies in parallel to the reference capacitance, C_0 , a parallel inductance makes the apparent reference capacitance smaller than the correct value below 70.04 ohms and larger above 70.04 ohms. In our example, the effective parallel inductance at $R = 30$ ohms is 20.3 μH ; at $R = 100$ ohms, the additional capacitance is 0.6 pF. Assuming that the reference capacitor $C_0 = 160$ pF, the respective values for variable capacitor C_N are 158.4 and 160.6 pF. This corresponds to errors of 0.99 and 0.37 percent, respectively. The R_N errors pertaining to these values are 0.16 and 0.16 percent, respectively. The C_N error is practically frequency independent, but the R_N error decreases rapidly with decreasing frequency.

This calculation indicates that the ideal potentiometer for the parallel bridge is one with capacitance wafer-to-case and that it is not advantageous to mount the potentiometer isolated from ground, as with the series connection. A good choice would be a 220-ohm potentiometer of about 1-inch diameter—a capacitance of 10 pF wafer-to-case.

Schematic of the parallel R-C noise bridge

Figure 16 shows the bridge elements as they are connected to points A and B of **Figure 2**.

It is important to position the S0239 connector and the potentiometer so the lead lengths are as short as possible. The parasitic inductances, $L1$ and $L2$, are also shown.

Calibration of the parallel bridge

1. Determine the value and position of the compensating capacitor ($C6$) as described earlier.

2. Reconnect the bridge elements carefully.
3. Connect a noninductive 50-ohm resistor at the X terminal. The frequency is 1.8 MHz. Null the bridge. If the frequency is increased to 28 MHz and the zero readings decrease with increasing frequency, inductance L_2 in series with the variable resistor is too low, and you must increase its value. This inductance may be a short piece of wire.
4. Repeat this test with a 22-ohm and 150-ohm resistor. The R_N and C_N "rotation" must be zero between 1.8 and 28 MHz. If it is not, you must readjust either L_1 or L_2 . To find out which inductance you should increase, use a small piece of ferrite glued to a Plexiglas™ strip to use as a tuning wand next to each coil.

Test results for the parallel R-C bridge

If you use a 220-ohm potentiometer with a wafer capacitance of about 10 pF, the impedance crossover point will be around 100 ohms and the sum of the R_N and C_N errors will be less than 1 percent for resistors under 100 ohms, at all frequencies.

Summary

The proper functioning of a noise bridge requires a truly balanced noise source for the bridge and good electrical tracking of the two bridge halves over the large frequency range.

The balanced noise source can be realized with a prebalancing transformer that also reduces capacitive cross talk between noise source and bridge elements.

For the series bridge, tracking can be improved by adding a capacitor in parallel to the potentiometer. But this measure does not fully correct the situation if the load impedance is capacitive, and is even less effective if it is inductive. The two error sources of the series bridge are the capacitance of the connector and

the wafer-to-case capacitance of the variable resistor. The first can be corrected, but only for pure resistances. The second can be reduced by the isolated mounting of the potentiometer and by selecting one with a small capacitance wafer-to-case.

For the parallel connection, the potentiometer can cause a small error that can be minimized by selecting a unit whose capacitance and inductance are matched properly. Isolated mounting is not desirable. Thus the remaining error is a fraction of the series-bridge error, and the reason for the superiority of the parallel R-C bridge. For sake of comparison, if the unknown impedance at 28 MHz is a 50-ohm resistor in series with a 50-ohm inductive resistance, the errors for the series bridge are 5.3 percent for R_N and 4.7 percent for C_N , whereas the sum of both errors is less than 1 percent with the parallel bridge.

For proper noise bridge operation, it is important to avoid chassis currents that can cause a shifting of the null position of the variable controls. Shielding and single-point grounding of the noise generator and a short, twisted-wire connection between noise generator and transformer assembly are mandatory. The noise generator and transformer assembly should be on opposite sides of the bridge elements on the chassis. If possible, use a reduction gear for both dials to take advantage of the high accuracy. I recommend an audio-frequency modulated noise source which makes it easier to detect the deep null. A great advantage is that both dials can be precalibrated with a DVM. Any possible R_N and C_N dial "rotations" can easily be corrected easily with compensating capacitors or balancing inductances.

REFERENCES

1. R.A. Hubbs and A.F. Dotting, "Improvements to the RX Noise Bridge," *Ham Radio*, February 1977, page 10.
2. Forrest Gehrke, K2BT, "A Precision Noise Bridge," *Ham Radio*, March 1983, page 50.
2. J. Grebenkemper, "Improving and Using Noise Bridges," *QST*, August 1989, page 27.

PRODUCT INFORMATION

VHF FM Monitor Receiver Fits In Shirt Pocket

The AR-33B is a solid-state, fully synthesized, portable VHF FM monitor receiver covering the communication band of 140–170 MHz in 5 kHz increments. Frequencies are selected from the entire band with a thumb-wheel switch, plus two channels may be recalled from memory via a single switch. The receiver employs 4 varactor tuned stages which track with the selected frequency for optimum performance over the entire band. Sensitivity is less than .22 μ V and power con-

sumption is 45mA when squelched. The unit is 5.12" x 2.45" x 1.02" and weighs 7 oz. with two AA alkaline batteries (included).

The AR-33B is available from Communications Specialists, Inc. and comes complete with a rubber flex antenna, earphone, instruction manual with detailed schematic diagram, BNC to SMA adapter, and 2 AA batteries. Contact Communications Specialists, Inc. at 426 West Taft Avenue, Orange, CA 92665-4296; or call 800-854-0547 (USA); or 714-998-3021 (Int'l).

QUARTERLY DEVICES

Radios without knobs

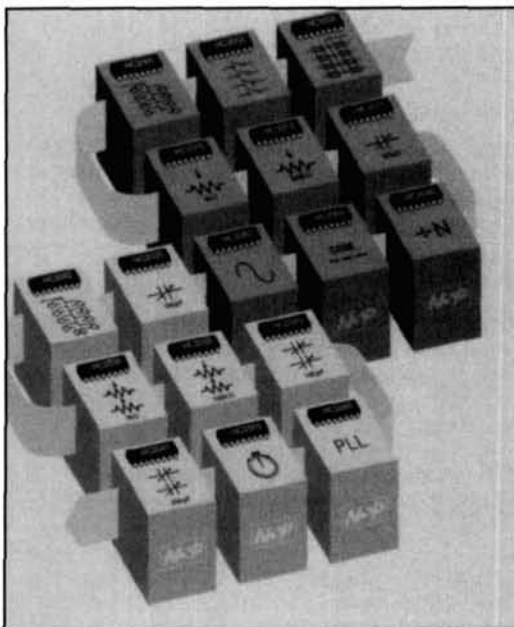
Imagine unpacking your new HF SSB transceiver to find a blank front panel—the faceplate totally devoid of switches, knobs, or displays of any kind. This notion may sound unappealing and perhaps even unthinkable to many of us, but with the emergent popularity of virtual instrumentation, such a radio is well within the realm of possibility. Indeed, most transceivers already accept limited RS-232 control, and it's probably human resistance more than technical feasibility that inhibits the marketing of a totally software-driven radio.

In reality, the notion of computer-driven radios and test equipment makes sense. The display, memory, and control possibilities are vast. And, the elimination of costly harnesses, operator control modules, and elaborate faceplates makes virtual equipment less expensive to manufacture. These factors could ensure a healthy future for products of this type.

To build a virtual instrument, the traditional operator-manipulated controls like pots and switches must be replaced by specialized circuits that change state or value in response to digital commands. In this edition of "Quarterly Devices," we'll explore a new family of digital products from Hughes Semiconductor that are especially designed for this task.

What is an NSP and what does it do?

The familiar potentiometers, trimmers, and switches found on radio and test-equipment panels have some important traits that we may take for granted. First, they hold their value or state when power is removed—meaning they exhibit a primitive form of nonvolatile (NV) memory. Second, their status is easy to read by glancing at the knob pointer or switch handle. Finally, settings are easily changed with a twist or flip of the wrist. It follows that any "virtual" substitute for these components should have



similarly user-friendly characteristics in the digital realm.

The NSP, or nonvolatile serial-programmable device, represents a new family of integrated circuits that offer the equivalent characteristics of traditional panel-type controls for software-driven equipment. In order to do this, NSP devices are designed to provide the following features:

1. A programmable circuit element such as a potentiometer, trim capacitor, switch matrix, or other adjustable circuit (some NSP devices contain more than one adjustable element).
2. A nonvolatile memory to store the desired switch state, component value, or operating function even when power is removed.
3. A serial interface to conveniently access the device's memory for the purpose of reading or changing it.

Serial access allows several devices to be cascaded with little or no added overhead, and requires a minimum number of control lines (3

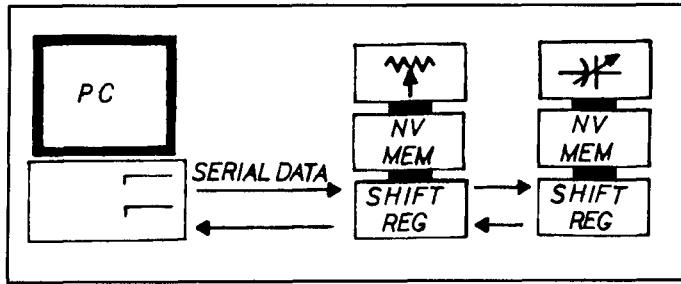


Figure 1. The NSP concept. A PC, using simple NSP software, programs and reads the chip's memory via serial interface. The memory then controls the device's variable element.

for write only, 5 for read/write).

In addition to the three basic functions outlined above, each NSP chip self-generates a regulated high-voltage source for reprogramming purposes, allowing the device to run from a single low-voltage source. Using CMOS technology, the NSP operates over a 3 to 10 volt supply range and draws less than 1 μ A in stand-by mode. Redundant memory cells back up the primary memory to ensure long-term reliability. NSPs provide up to 1,000,000 write cycles, and have a data retention life of 10 years.

Controlling the NSP

The conceptual diagram shown in Figure 1 provides a thumbnail sketch of how the NSP chip works. An IBM-compatible PC—or other microprocessor-based system—communicates

with the chip's serial port to either program or read information stored in the device's non-volatile memory. This is done by means of a relatively simple software program. The chip's memory, in turn, controls the switch state, component value, or adjustable function provided by the NSP device.

For a closer look at the architecture of the NSP control system, refer to Figure 2. The serial data line has two tandem shift registers. The first, an "N" bit register, holds adjustable-element commands and provides a parallel data path to/from the chip's NV memory. The second, an 8-bit register, holds ID codes and provides a parallel data path to/from the chip's ID code register.

Each type of NSP device has a unique code stored in its ID register. When programming, the ID code accompanying new NV-memory data must match the chip's internal code for any update to take place. This prevents accidental programming, and lets several devices share the same serial control line. In read mode, the ID code is retrieved with the NV-memory data and is used to identify which type of device is being interrogated.

Data stored in the N-bit non-volatile memory controls the NSP's adjustable element. Between 4 and 32 internal control lines may be used—depending upon the number of bits needed to set up the device (8 is typical for a single-gang pot or trimcap). These control lines become valid any time power is applied and remain valid until power is removed.

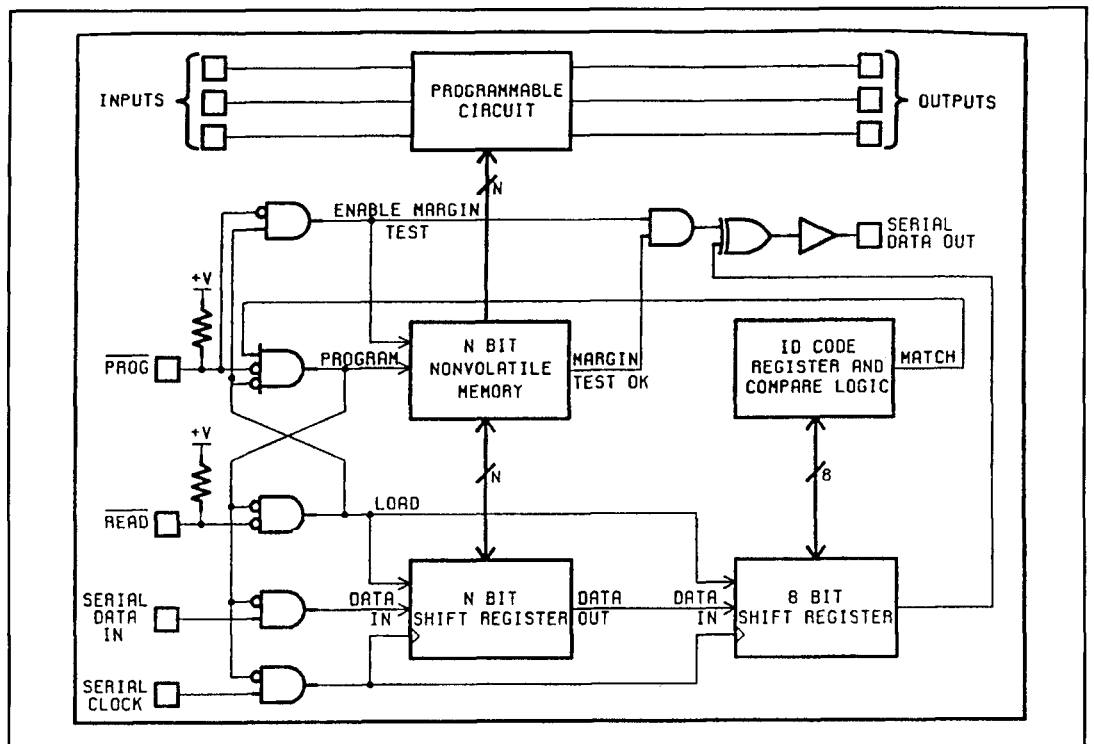


Figure 2. NSP control system block diagram.




MODE	READ	PROG	SERIAL CLOCK	SERIAL DATA IN	SERIAL DATA OUT	NV DATA
STANDBY	1	1	0	X	X	VALID
READ	0	1	0	X	LSB of ID CODE	VALID
SHIFT DATA	1	1		VALID DATA IN	VALID DATA OUT	VALID
PROGRAM	1	0	X	ID CODE & DATA	X	NEW DATA
MARGIN TEST	0		0	X		VALID

Table 1. Truth table for NSP operation modes.

Operating modes are set by the state of the READ and PROGRAM input lines (see Figure 2). The NSP device has 5 operating modes, as outlined below:

1. STANDBY: Places serial interface and NV memory in low-current standby.
2. READ: Initiates asynchronous parallel loading of both registers (NV data into the N-bit register, ID code into the 8-bit register).
3. SHIFT MODE: Invoked prior to each PROGRAM cycle to load new NV memory data and access coding into the shift registers. Also used after a read pulse to shift NV data and ID coding out of the device.
4. PROGRAM: Used to transfer data from the N-bit shift register to NV memory.
5. MARGIN TEST: A check to see if all memory cells have been sufficiently programmed during the transfer.

A truth table showing the input conditions required to initiate each mode is shown in Table 1.

Switches, components, and functions provided by the NSP

The Hughes NSP family of devices offers a wide range of "drop in" adjustable components. NSPs are being made to replace panel switches, dip switches, jumpers, mechanical potentiometers, and trim-caps, plus many other circuit building-blocks like timers, filters, PLLs, and waveform generators. Table 2 provides a complete listing of all NSP devices either currently available or scheduled for delivery in the near future. However, more are in development, so you may want to contact Hughes for an update.

A look inside the adjustable element, or "business end" of the NSP is revealing. To provide sufficient resolution, the value range of the device's adjustable element must be split into small incremental steps. To accomplish this, NSP trim capacitors use a simple binary addi-

tion scheme like the one illustrated in Figure 3A. This particular 250-pF trimmer has 256 equally spaced steps derived from a parallel network of 8 switchable fixed-value capacitors (resolution is 0.4 percent with 256 steps). Potentiometers use an 8-stage R-2R resistor ladder network like the one in Figure 3B. This R-2R ladder configuration also provides 256 steps with 0.4 percent resolution. Switch matrixes, like the quad analog switch shown in Figure 3C, use the same basic control system as the analog parts to establish the required two-way signal paths through the device.

One note for RF designers: NSP trimmer-capacitors do not exhibit sufficiently high Q at HF for direct application in RF oscillator, amplifier, or filter circuits. Their use is primarily limited to trimming lower-frequency circuits such as active filters and oscillators operating below 1 MHz. Also, NSP resistors and switches have frequency response limitations that could restrict their use in some applications. Response curves for some NSP devices are included in the data sheets.

Learning to use NSPs

If you want to put NSP devices to work in your next project, it may be a lot easier than you think. Hughes Semiconductor offers a special NSP development kit for designers that promises to have you up and running in a matter of minutes. The development kit provides the following items:

- PC interface board
- DB-25 cable
- 5-1/4" floppy disk
- HC2062 sine/cosine generator (sample chip)
- Documentation

You'll need to supply an IBM-compatible PC with a free parallel port and an external +5 volts DC power supply. The kit lets you test a single device or access and program a ring of

PART NUMBER	PART TYPE	DESCRIPTION	VALUE	RESOLUTION	NV MEMORY SIZE	ID CODE
HC2001	DIP SWITCH	8 Three State Outputs with a Common Enable	-	-	8	D1
HC2002	DIP SWITCH	8 Open Drain Outputs with a Common Enable	-	-	8	D2
HC2003	ANALOG SWITCH	4 Independent Analog Switches	-	-	4	D8
HC2004	DATA SELECTOR	8:1 Analog Multiplexer with Independent Switch Controls	-	-	8	A9
HC2010	TRIM POTENTIOMETER	R-2R Ladder with Unbuffered Output	1K Ω	0.4%	8	39
HC2012	TRIM POTENTIOMETER	R-2R Ladder with Unbuffered Output	100K Ω	0.4%	8	3A
HC2021	TRIM CAPACITOR	Digitally Controlled Trim Capacitor	250 pF	1 pF	8	99
HC2030	DUAL TRIM POTENTIOMETER	Dual R-2R Ladders with Unbuffered Outputs	1K Ω	0.4%	16	33
HC2032	DUAL TRIM POTENTIOMETER	Dual R-2R Ladders with Unbuffered Outputs	100K Ω	0.4%	16	36
HC2040	DUAL TRIM CAPACITOR	Two Digitally Controlled Trim Capacitors	100 pF	0.4 pF	16	93
HC2041	DUAL TRIM CAPACITOR	Two Digitally Controlled Trim Capacitors	250 pF	1 pF	16	96
HC2055	TIMER	Long Range Programmable Timer 2 ²⁶ Maximum Count	-	-	16	C5
HC2062	WAVEFORM GENERATOR	Digitally Controlled Sine & Cosine Generator with Quadrature Outputs	-	16 Levels, 32 Steps/Cycle	12	71
HC2063	SECURITY CODE	16 Bit Code Generator/Detector with Binary and Manchester Outputs	-	2 ¹⁶ Codes	16	16A
HC2090	DIVIDE BY N COUNTER	Single 16 Bit Counter or Dual 8 Bit Counters	-	16 Bit	16	CA
HC2020	TRIM CAPACITOR	Digitally Controlled Trim Capacitor	100 pF	0.4 pF	8	9A
HC2027	TRIM CAPACITOR	Digitally Controlled Trim Capacitor	500 pF	0.5 pF	10	95
HC2095	RATE MULTIPLIER	Programmable 2 Decade Rate Multiplier	-	0.01	8	CC
HC2061	PULSE WIDTH MODULATOR	Programmable 2 Decade Rate	-	0.1%	10	56
HC2065	PHASE LOCKED LOOP	PLL with 8 Bit Input and Feedback Dividers	-	-	16	59
HC2070	FILTERS	6 Pole Low Pass Switched Capacitor Filter	-	-	16	F0

Table 2. Listing of NSP devices.

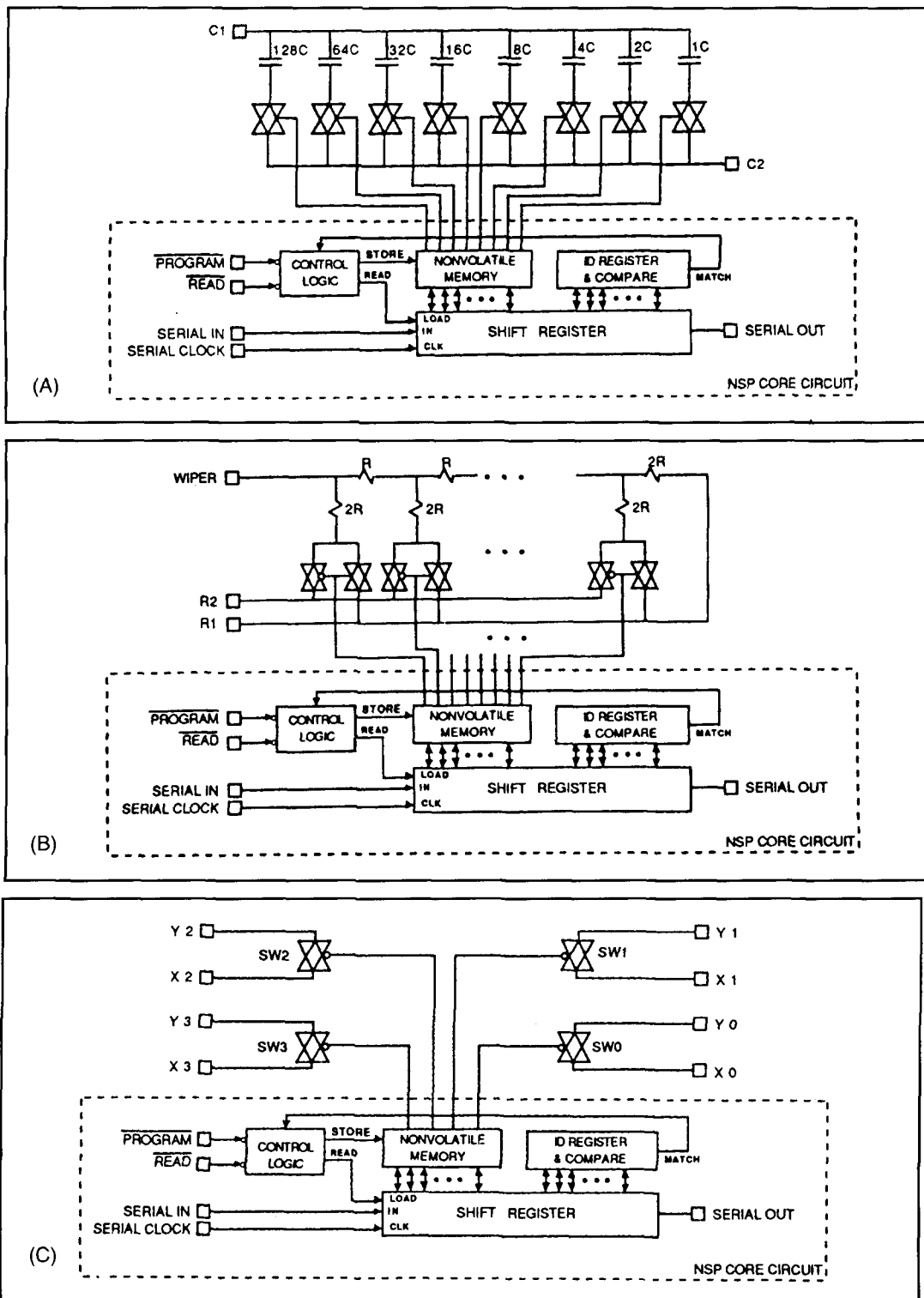


Figure 3. Typical programmable circuits (A) trim capacitor, (B) potentiometer, (C) bi-directional quad switch.

up to 15 NSP devices. Three operating modes—direct entry, increment, and ramp—are supported by the software. Complete hardware and software documentation, quick-start instructions, and a request form for additional samples are included.

The bottom line

NSPs are relatively inexpensive—approximately \$4 per unit in small quantities, falling to under \$2 in larger lots (prices vary depending on the package). A comprehensive packet of

literature on NSPs and how they operate is available free of charge from the Hughes Aircraft Company Semiconductor Products Center, P.O. Box H, Newport Beach, California 92658-8903. The Center's NSP HOTLINE number is (714) 759-6589, or you can FAX them at (714) 759-2913. The introductory price for the NSP development kit is \$50. To order a development kit, or to obtain product, sample, and dealer information, call the NSP HOTLINE.

Conclusion

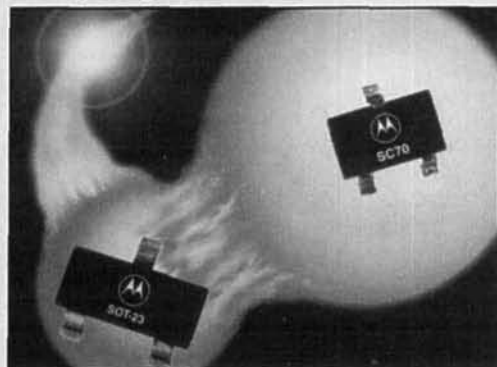
Virtual instrumentation—and specialized

devices like NSPs—represent a growing trend in hardware design. Hopefully, this brief introduction to NSPs will start your wheels turning and make the virtual instrumentation concept "real" enough to apply in a forthcoming project. In closing, I invite your suggestions for future columns. Last issue's torsional filter article was suggested by Scott Prather, K9BY, and NSPs were brought to my attention by John Pivnichny, N2DCH—living proof that I depend upon creative scouting by authors and readers to keep the column relevant and interesting. Thank you for your help. ■

PRODUCT INFORMATION

Motorola's Lowest Noise Amplifier Transistors Now Available In An Ultra-small Surface Mount Package

Motorola has introduced its lowest noise small signal amplifier transistors in a miniature surface-mount package. The MRF947 and MRF957 small signal transistors are state-of-the-art low noise NPN bipolar devices with the same die as Motorola's MRF941 and MRF951 transistors, mounted in the SC-70 surface mount package.



Applications for these transistors include low noise front ends in receivers and low power linear amplifiers, particularly where size, weight and low cost are prime design goals and the method of manufacture is automated surface mount assembly.

Typical performance characteristics of both devices at 1.5 GHz are gains of 10dB and noise figures of 2dB. The MRF947 transistor has a maximum current rating of 50mA. The MRF957 has a maximum current rating of 100mA.

Samples and small quantities are available from stock, and production quantities have a 16 week lead time. For further information, contact your local Motorola Semiconductor Sales Office or Authorized Motorola Distributor.

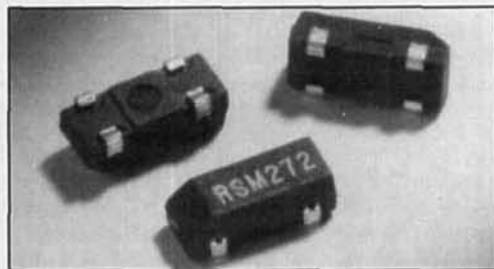
New Low-Frequency SMT Crystal Uses Tiny Tuning Fork Inside

Raltron Electronics Corp. announced a new surface mount crystal for use in clock radios, wrist watches, timers, computers, telecomm systems and other applications which generate, use and display real time in seconds, minutes, hours and days.

Quartz crystals ordinarily generate much higher frequencies. In order to generate the standard 32.768 kHz for convenient divide-by-60 countdowns, Raltron has shaped the internal crystal into a little tuning fork. This new tuning-fork crystal is hermetically sealed in an SMT package and designated Model RSM-200.

New Model RSM-200 Tuning Fork Crystal features an accuracy of ± 30 ppm at 25°C. It has an equivalent series resistance of 35 K ohms max., a load capacitance of 12.5 pF max., shunt capacitance of 2.0 pF max., and a drive level of 10 μ W maximum. Operating temperature range is spec'ed from -10°C to +60°C.

Surface mount pinout is on 5.6 mm centers. Profile height is 3.3 mm; footprint on the board is 9.1 mm x 4.0 mm maximum. For automatic insertion, units are tape-and-reel packaged in lots of 1,000.



For more information contact Raltron Electronics Corporation, 2315 NW 107th Avenue, Miami, FL 33172; 305-593-6033.

SMALL LOOP ANTENNAS: PART 2

Loop preamplifiers and loop projects

In part 1 of this two-part series on loop antennas, I discussed the basic theory of loop antennas. I gave the equations for calculating the loop inductance for square loops, and the equation for determining resonating capacitance. In this final installment, I'll look at some loop preamplifiers needed to overcome the low signal levels of loop antennas, and some actual loops I built while researching this article.

The loop projects may or may not be optimal, but were designed with simplicity of construction in mind. I'd be interested in seeing any variations that readers constructed either in the past or from this article.

The output signal levels available from loop antennas are very low. Because of this, a loop preamplifier is indicated for all but the most sensitive receivers. The preamplifier can be mounted either at the receiver or the antenna, but is most effective when mounted at the antenna (unless the coax to the receiver is short).

Figure 1 shows a typical loop preamplifier based on common NPN silicon transistors.* The transistor for this application is the 2N5179, but it is sometimes a bit hard to obtain from electronic parts distributors that cater to hobbyists.** There are, however, suitable substitutes available from service replacement transistor manufacturers like ECG and NTE.

Direct crossovers are the NTE-316 and ECG-316 devices. I had to order these from my local distributor, so in the meantime I successfully used the NTE-108 (similar to the ECG-108) device—which they did have in stock. The amplifier produced an average gain of about 15 dB across the AM broadcast band, although it had higher gain at the upper end than at the lower end.

The circuit for the preamplifier (Figure 2) is relatively straightforward, except for the matching transformer, T1. This transformer is a balun device you wind yourself. Use Amidon TF-37-61 or TF-50-61 ferrite toroidal cores. The winding is bifilar, no. 26 enameled wire. I used twelve turns on a TF-37-61 form, which seemed to work well. You might want to experiment with this circuit by varying the number of turns with changes of frequency.

Balun transformers are wound in the bifilar manner, as I just mentioned. This means that the two wires used for the primary and secondary windings of T1 are twisted together, about eight twists to the inch, and then wound over the form as if there were a single wire. I make twisted enameled wire for baluns by taking a length of No. 26 enamel insulated wire, folding it back on itself, and chucking it up in a hand drill. Anchor the far end in a vise, then turn the crank on the drill to twist the wire. This operation can be done using an electric drill, if the drill has a gentle trigger and variable speed control, but it can be dangerous. If you chose to twist the wires using an electric drill, wear safety goggles or glasses to protect your eye-

*This circuit has appeared in ARRL and Heys (see "Small Loop Antennas: Part 1," *Communications Quarterly*, Winter 1993), and is a derivation of a W7ZOI design.

**These transistors are available from any Motorola distributor. Ed.

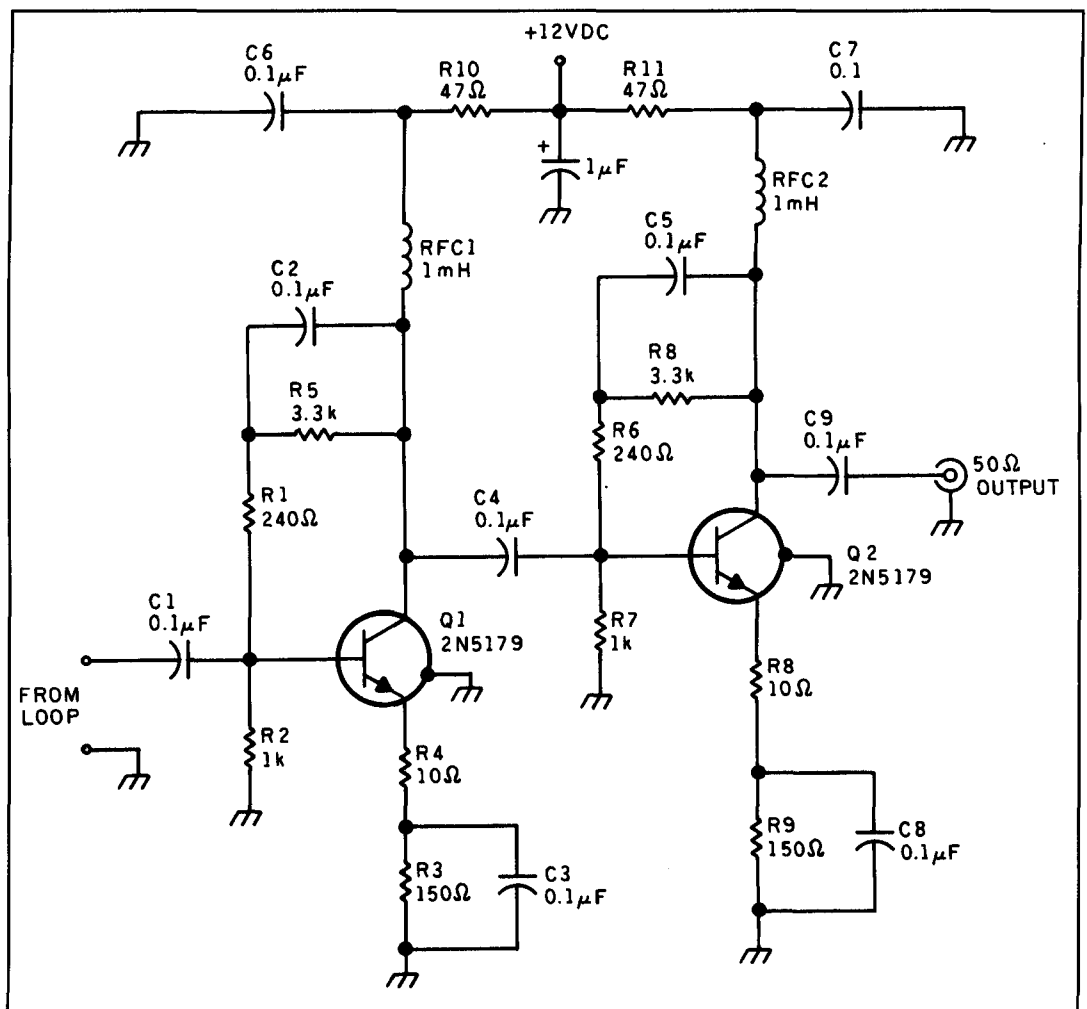


Figure 1. A typical loop preamplifier based on common NPN silicon transistors.

sight. If that wire comes loose and starts whipping around, it could damage your eyes.

A related amplifier circuit is shown in **Figure 2**. This circuit also uses the 2N5179 transistor, or its service-grade replacements. The amplifier consists of two identical stages, each providing 10 dB of gain. Up to four stages can be cascaded for 40 dB of gain, although you must be careful to observe good layout practices in such a case (inadvertent feedback can cause oscillation when gains are high).

The amplifier of **Figure 2** is useful at full specification gain from about 1 MHz throughout the HF shortwave bands. At lower frequencies gain will drop off somewhat, but that doesn't necessarily mean that it becomes useless. Besides, you can make certain modifications that will coax a little more gain out of the circuit at VLF frequencies. For example, all of the coupling and decoupling capacitors (that is, every one by C2 and C5) can be increased in value to accommodate lower frequencies. I obtained a bit more gain by using 0.47 μF instead of 0.1 μF for those capacitors. You can

also increase the values of the RF chokes (RFC1 and RFC2) from 1 to 2.5 mH.

The circuits of **Figure 3 and 4** are based on the junction field effect transistor (JFET); the circuit of **Figure 3** is a common gate design, while that of **Figure 4** is a common drain design (that is, a "source follower"). Both are broadband amplifier circuits, so they can be used throughout the HF spectrum, as well as VLF and MW. If gain is too low at some very low frequencies, the same advice regarding the value of capacitors stands as given for the circuit in **Figure 2**.

The particular JFET specified for these circuits is the MPF-102—a device that works well into the VHF region. Suitable replacements from the service-shop lines of devices include the ECG-312 and the NTE-312. Because the application isn't terribly critical, and the devices aren't expensive, it's reasonable to experiment with other JFET devices if they are more readily available than those I selected.

All of the preamplifier circuits in this section can be used with either tuned or untuned loops,

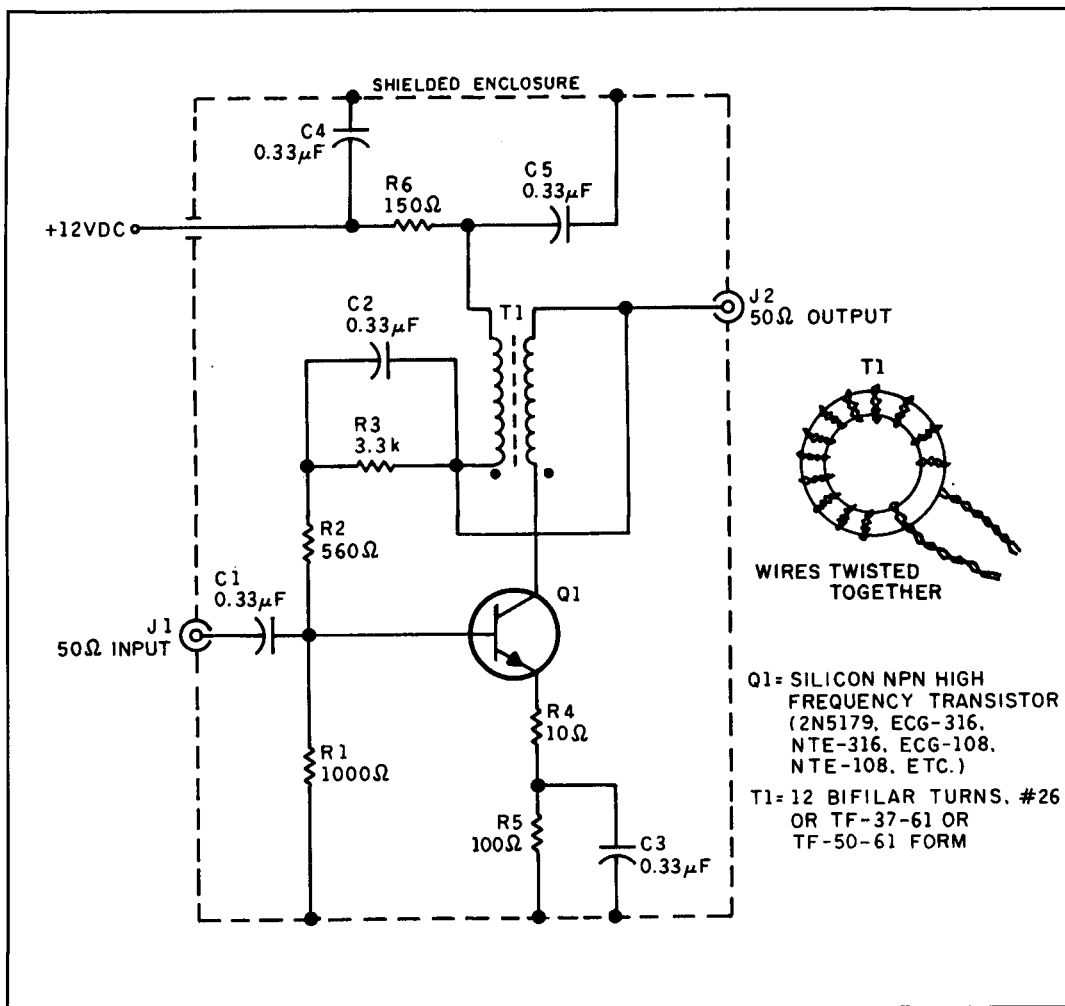


Figure 2. Another version of the transistor preamp.

although neither the loop nor the tuning capacitor are shown in all cases.

The preamplifier in **Figure 5** is based on a product called a monolithic microwave integrated circuit (MMIC), which can work at all frequencies from near DC to more than 1,000 MHz (microwave). The principal limitation on operating frequency is the capacitors used for C1 and C2. They should have a value of 0.1 μF for VLF, the AM broadcast band and the medium wave bands (<3 MHz); 0.01 μF for the low end of the HF region; 0.001 from the upper HF to about 100 MHz; and 100 pF for >100 MHz operation. They are shown with 0.1 μF values in **Figure 5** because this article is on loop antennas, and these antennas work best on lower frequencies.

There are several MMIC chips available on the commercial market, but the one that's most easily available to hobby purchasers is the Mini-Circuit Laboratories, Inc., type MAR-1 used in **Figure 5**. Note that only four connections are used on the MAR-1 device: input, two grounds, and output (the output also serves for

DC V+ power supply connection). The value of the resistor used for R1 depends on the value of V+. The goal is to obtain a voltage (V) of 7 volts DC or a little less at the output terminal of the MAR-1 when 0.015 amperes of current are drawn. Therefore, from Ohm's law, you may deduce very cleverly:

$$R1 \text{ (ohms)} = \frac{V+ - 7 \text{ volts}}{0.015 \text{ amps}}$$

For a +12 volt value for V+, then, use a 300-ohm resistor for R1, and for a +9 volt battery, use 120 ohms. A typical homebrew MAR-1 preamplifier is shown in **Photo A**.

Sharpening the loop

Many years ago the *Q multiplier* was a popular add-on accessory for communications receivers. These devices were sold as Heathkits, and many construction projects

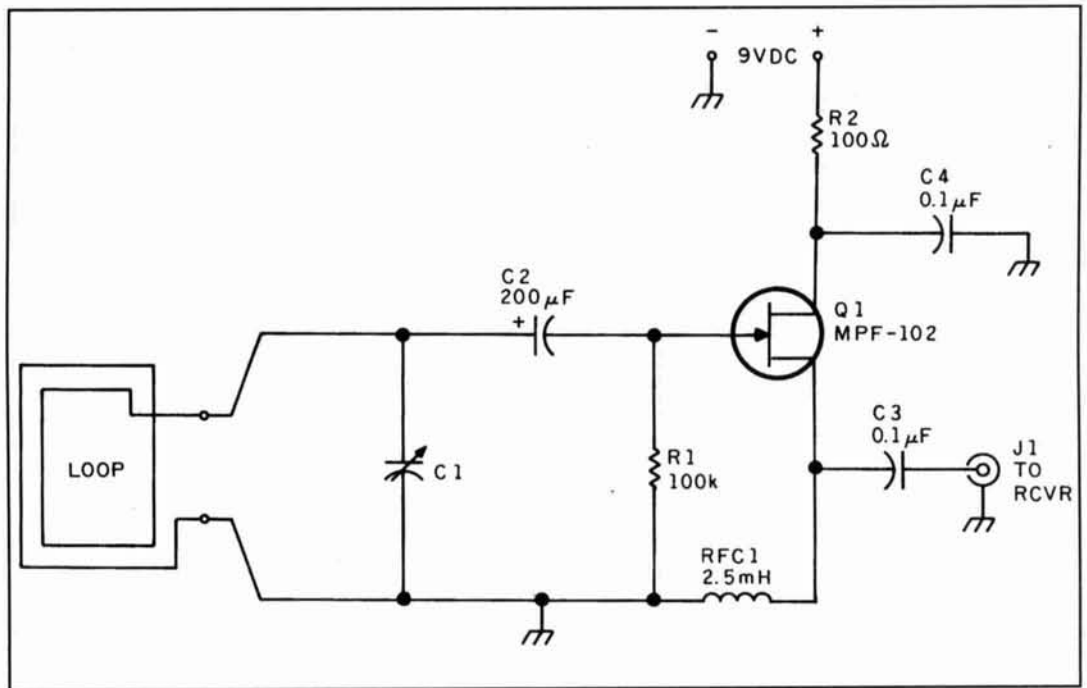


Figure 3. Common gate design circuit based on the JFET.

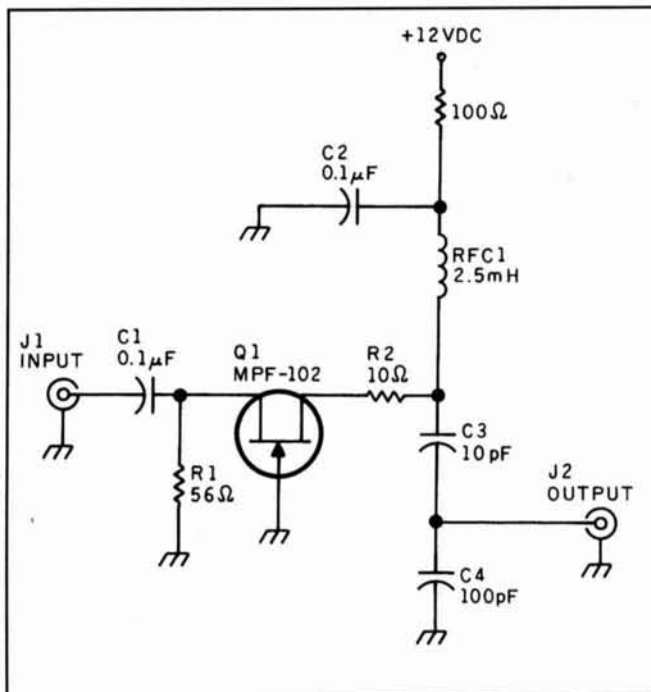


Figure 4. Common drain design circuit based on the JFET.

appeared in magazines and amateur radio books. The Q multiplier has the effect of seeming to greatly increase the sensitivity of a receiver, as well as greatly reducing the bandwidth of the front end. Thus, the device will allow better reception of some stations because of increased sensitivity and narrowed band-

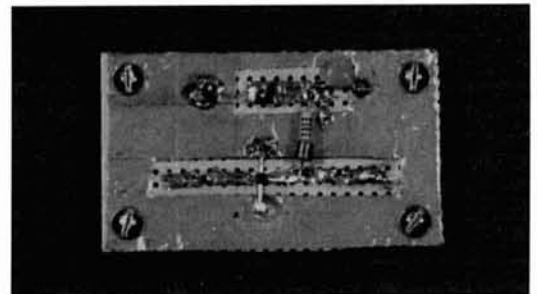


Photo A. A typical homebrew MAR-1 preamplifier.

width. Cornell described a loop antenna Q-multiplier in 1975.¹

A Q multiplier is an active electronic circuit placed at the antenna input of a receiver. It is essentially an Armstrong oscillator (Figure 6) that doesn't quite oscillate. These circuits have a tuned circuit (L1/C1) at the input of an amplifier stage, and a feedback coupling loop (L3). The degree of feedback is controlled by the coupling between L1 and L3. The coupling is altered both by varying how close the two coils are, and their relative orientation with respect to each other. Certain other circuits use a series potentiometer in the L3 side that controls the amount of feedback.

The Q multiplier is adjusted to the point that the circuit is just on the verge of oscillating, but not quite. As the feedback is backed away from the threshold of oscillation, but not too far, the narrowing of bandwidth occurs—as does the increase in sensitivity. It takes some skill to

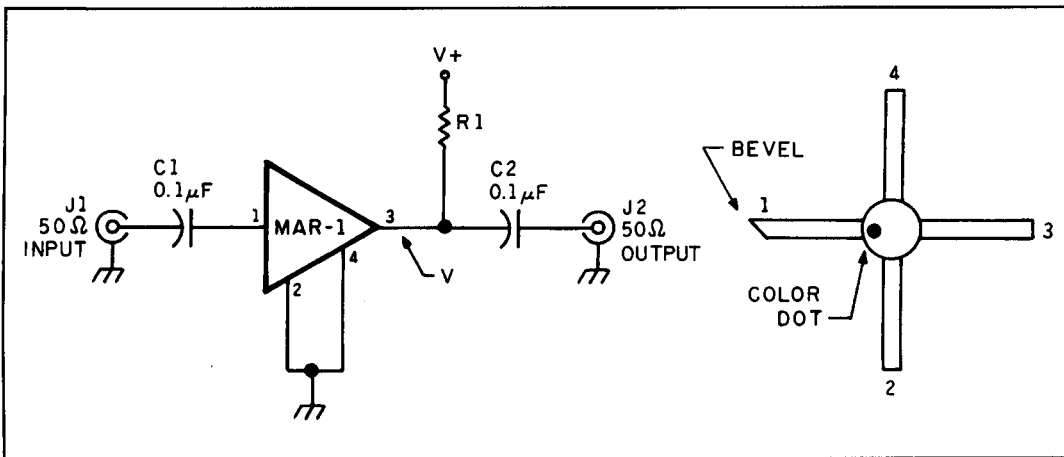


Figure 5. Preamplifier based on the MMIC. The MAR-1 by Mini-Circuits is used in this circuit.

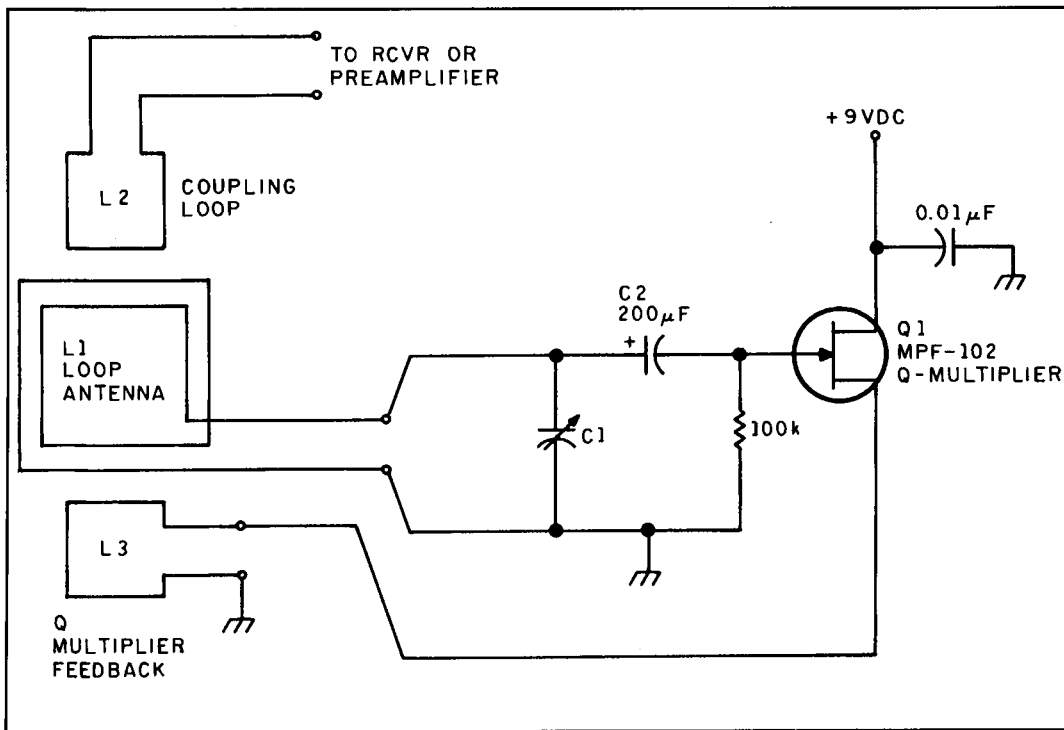


Figure 6. Armstrong oscillator.

operate the Q multiplier, but once the skill is gained it's easy to use.

One of the better applications of Q multipliers is in the reception of Morse code (A1 radiotelegraphy). Adjust the Q multiplier to a point where it is unstable and nearly ready to oscillate *when no signal is present*. When a CW signal is tuned in, the extra signal trips the circuit into oscillation. This tends to "rejuvenate" the signal, and make it easier to recover.

The "Embroider loop"

Fashioning the loop is usually a bit of a chore, and winding even a dozen or so turns on

a large form can be daunting—especially since the turns keep falling off the form. While working on my antenna column in *Popular Communications*, I hit upon an idea that allowed me to make a multi-turn transformer loop in which the outer loop (L1) is a 14-turn loop, and the receiver coupling loop is 2 turns. The loop is made of 16-conductor colored parallel ribbon cable of the sort used in the computer industry. Two wires in the center of the bundle (grey and white) are used for L2, while all the rest are used for L1. The form is a 14-inch embroidery or needlepoint hoop, which I bought from a craft store for \$1.49. It consists of two concentric, close-fitting wooden hoops. The inner hoop is continuous, while the outer

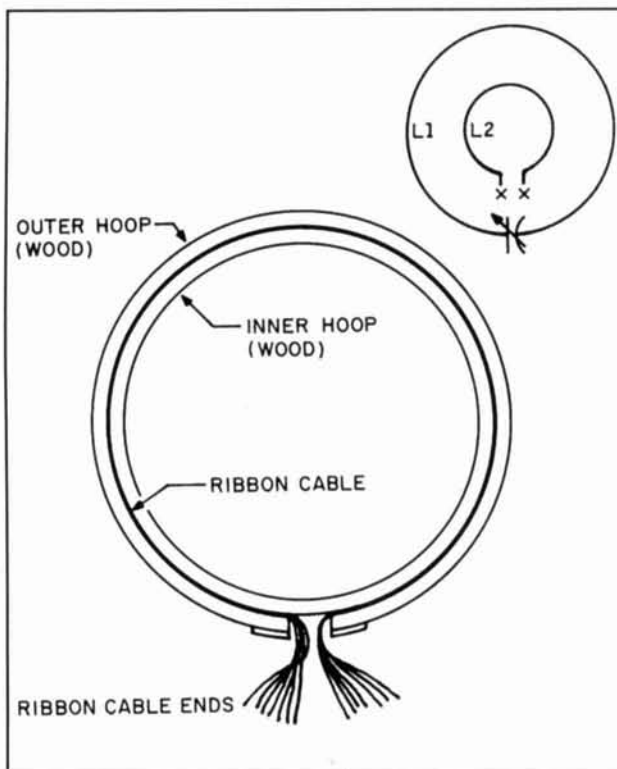


Figure 7. (A) Sandwich the ribbon cable into the space between the inner and outer hoops.



Photo B. The Embroider loop antenna.

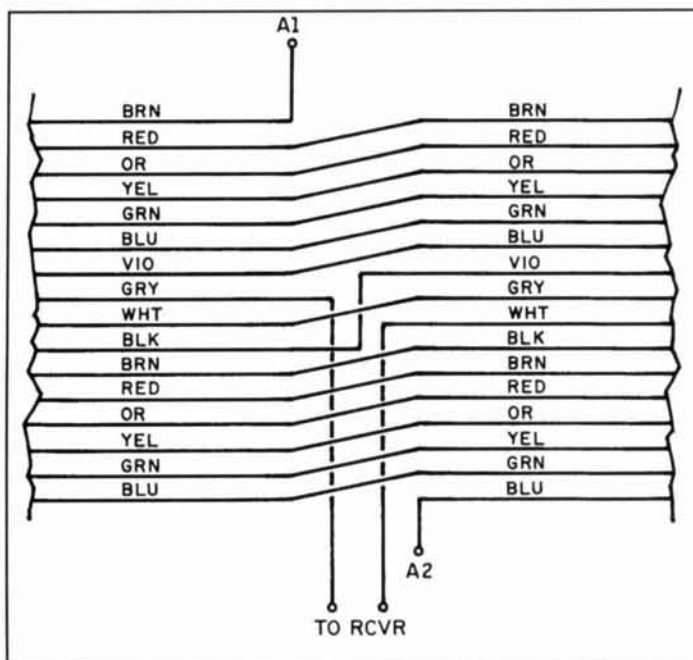


Figure 7. (B) The conductors of the cable are cross connected to form one continuous coil.

hoop is broken at one point. A screw and two tabs hold the ends together, and lets you adjust for size. When embroidering, the fabric is placed between the two hoops and secured with the screw; in my loop antenna the 16-conductor

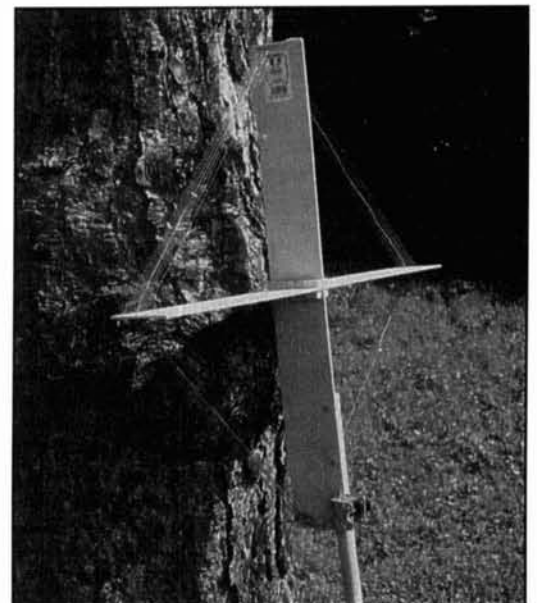


Photo C. The square loop hobby board antenna.

ribbon cable is sandwiched into the space between inner and outer hoops (Figure 7A). The conductors of the cable are cross-connected in the manner shown in Figure 7B to form one continuous coil.

The inner loop is connected to a length of coaxial cable, which carries signal to the radio receiver. The outer loop is connected to a variable tuning capacitor. I found that a 365 pF "broadcast band variable" capacitor was suffi-



Photo D. The picture frame loop antenna.

cient to tune the loop in all sections of the AM band. But some people tell me that two or three sections of 365 pF may be needed in some cases where fewer turns are used. **Photo B** shows the completed loop.

The square "hobby board" loop

A very common way to build a square loop antenna is to take two pieces of thin lumber, place them in a cross shape, and wind the wire around the ends of the wooden arms. This type of antenna is shown in **Figure 8A**. The wooden supports can be made of 1x2 lumber, or some other stock. While researching this article, I made a test loop with two 1/8 x 3 x 24-inch *Bass wood* "hobby board" stock acquired from a local hobby shop. Model builders use this wood as a stronger alternative to balsa.

The electrical circuit of the hobby board loop (also shown in **Figure 8A**) is a transformer loop design. The main tuned loop is on the outside, and consists of ten turns of no. 26 enameled wire. It's tuned by a 365-pF capacitor. The inner loop is used for coupling to the receiver, and consists of a single turn of no. 22 insulated solid hook-up wire.

The mechanical fit between the vertical and horizontal wooden pieces is shown in **Figure 8B**. Note that four 1/2-inch wooden blocks (also obtained at the hobby shop) are glued to the junction of the two cross pieces to stiffen the joint. Details for the boards are shown in **Figure 8C**. Each board is 24 inches long. At the midpoint (12 inches), there's a 1/8-inch

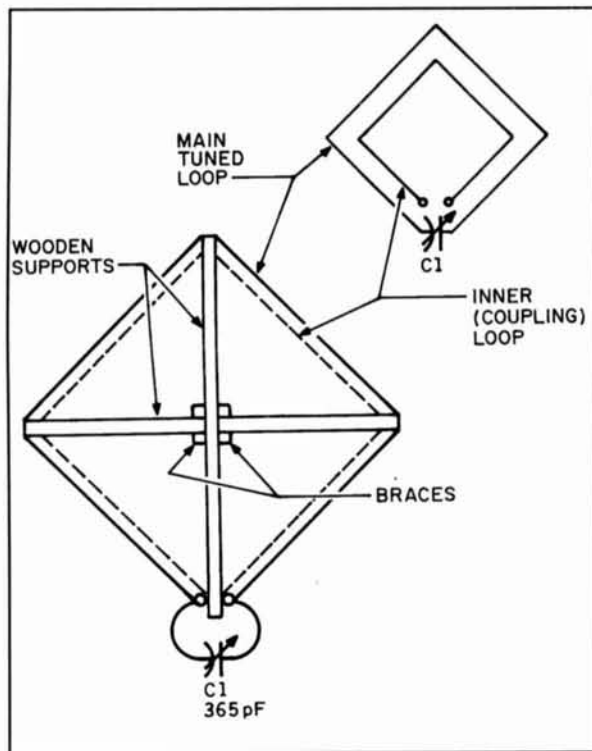


Figure 8. (A) The design of and electrical circuit for the hobby board loop.

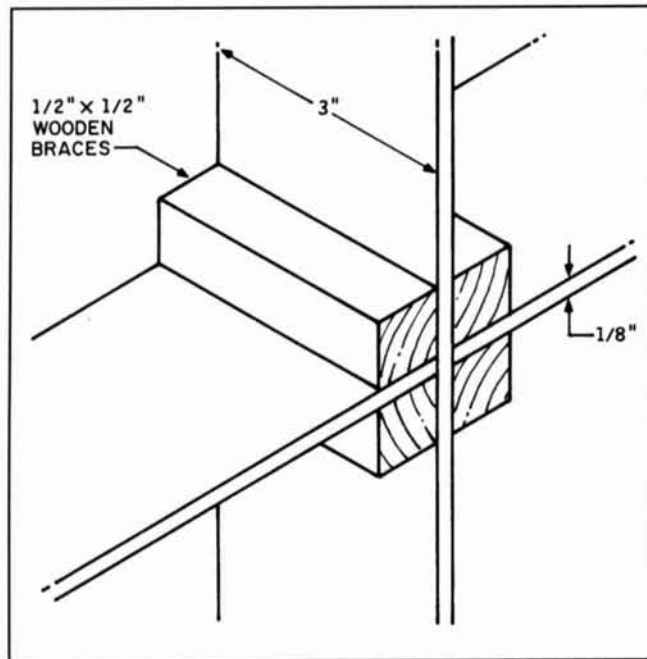


Figure 8. (B) The mechanical fit between the vertical and horizontal wooden pieces.

wide, 1.5-inch long slot cut. These slots are used to mate the two boards together.

At each end there are ten tiny slits made by a jeweler's saw (also called a "jigsaw" in hobbyist circles) with a thin blade. These slits are just wide enough to allow a single no. 26 wire to be

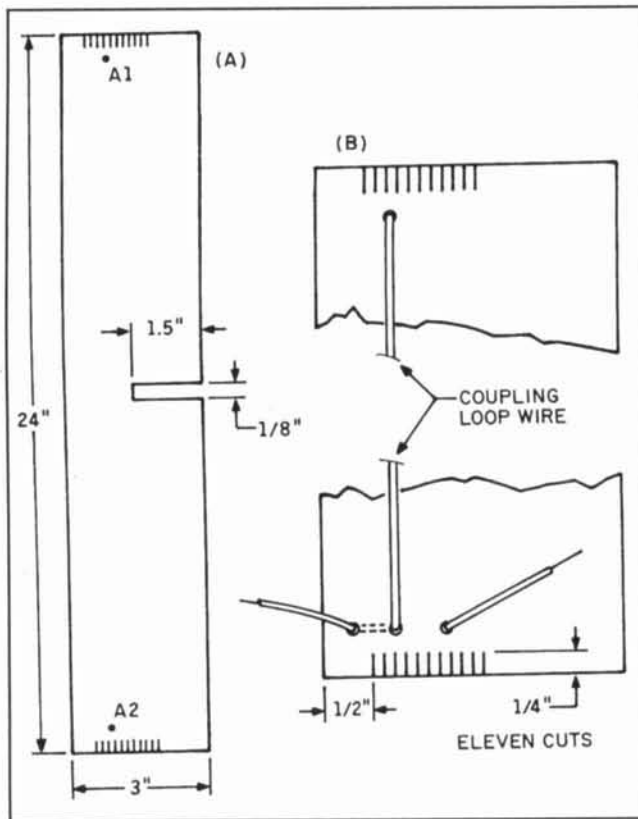


Figure 8. (C) Details for the boards.

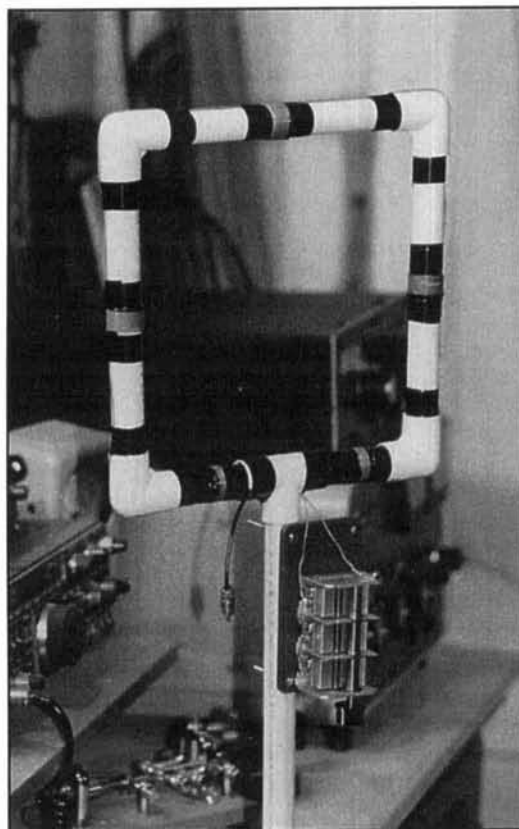


Photo E. The square loop loopstick antenna.



Photo F. Commercially available Palomar Engineers loop antenna. The Palomar LA-1 loop amplifier is used here and in Photo G.

inserted without slipping. The slits are 1/4-inch long, start 1/2 inch from one edge, and are 1/8-inch apart. There are ten slits on both ends of the horizontal piece, while the vertical piece has ten slits on the top end and eleven slits on the bottom end. The reason for offsetting the wire slits is to allow room on the other side of the 3-inch width of the vertical member for a mounting stick.

By the way, if you do a little mobile "hidden transmitter hunting," be careful when holding this antenna out the window of the car while searching for the transmitter. A friend of mine may have gotten a little too enthusiastic one Sunday morning as he homed in on his quarry. The police report said something about a nut speeding down the street waving a cross out the window of his car (*sigh*).

When assembling the antenna, use wood glue on the mating surfaces, square them to be at right angles to each other, and clamp the two pieces in a vice, or with C-clamps for thirty minutes (or longer if the glue manufacturer specifies). Next, glue the support blocks into place and clamp them for a similar period. The completed antenna is shown in **Photo C**.

Build the picture frame loop

When I went to the craft and hobbies shop to buy the embroidery hoops for the Embroider loop antenna, I ran across a display of raw, unfinished wooden picture frame lumber.

These sticks came in lengths up to 15 inches and were about 2-1/2 inches wide. Each piece of wood was cut at a 45-degree angle at each end, in a tongue-and-groove manner, so the pieces could be fitted together to make a frame. If four pieces are used, it's possible to make a 15-inch square loop. Larger loops can be made (in 15-inch increments) by connecting the pieces end-to-end, although the 15-inch variety seems sufficient. **Figure 9A** shows a square loop antenna made with picture framing stock for the loop, and 1x2 lumber for the vertical support.

The antenna is made using the same kind of 16-conductor computer ribbon cable used in the Embroider loop. Fifteen turns are used as the multi-turn loop, and one turn as a coupling loop to carry the signal to the receiver. The flat ribbon cable and the wood are smeared with wood contact cement, and then the cable is pressed onto the wood and held fast with clamps. The first loop of this type that I built used ordinary wood glue, and didn't hold the wire well. As a substitute, I held the cable to the wood with ordinary thumb tacks (one every three inches or so). The tacks I chose had plastic bodies, which reduced coupling between loops through the tack head.

Figure 9B shows the wiring scheme to make the parallel ribbon cable conductors into a single, multi-turn loop. The color scheme is standard, but make sure that your cable is of the same standard before relying on the diagram. The idea is to connect each conductor to its adjacent conductor at the feedpoint. The connections are made on a piece of integrated circuit (IC) printed wiring prototyping board designed for DIP ICs. Radio Shack carries several varieties, most of which are suitable. Vector and other suppliers to the hobby market also make such stock. The finished antenna is shown in **Photo D**.

The square loop loopstick

Loopstick antennas work differently from the small loop antennas that are the subject of this article. For one thing, the null is off the ends of the antenna, and the peak reception is orthogonal to the antenna. But if you arrange four loopsticks in a square (**Figure 10A**), the antenna works like a conventional loop antenna. Marris described this type of antenna previously, although the design was slightly different.²

Figure 10B shows the mechanical form of this antenna. It consists of four elements made from 0.75 inch PVC plumbing pipe. Each element contains an Amidon Associates, Inc. (2216 East Gladwick Street, Dominguez Hills, California 90220) type R61-050-750 ferrite

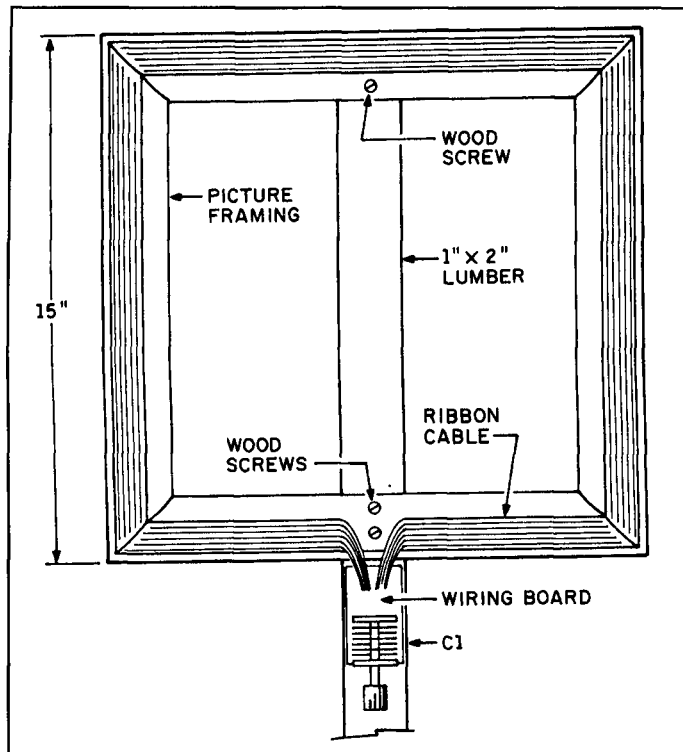


Figure 9. (A) square loop antenna made using picture framing stock for the loop and 1x2 lumber for the vertical support.

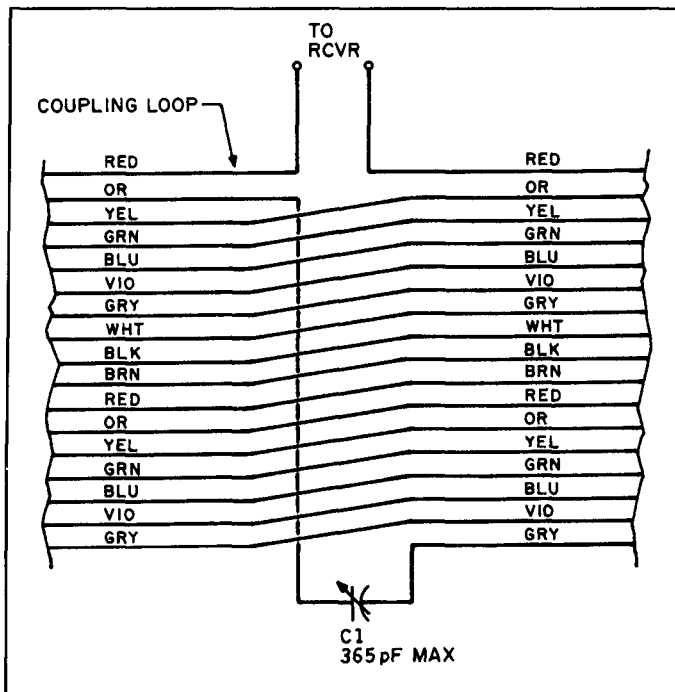


Figure 9. (B) The wiring scheme to make the parallel ribbon cable conductors into a single, multi-turn loop.

rod. These rods are 0.50 inch diameter, and 7.5 inches long. The type R61 nickel-zinc material is useful from 200 kHz to 15 MHz, so is suitable for all of the amateur bands where loops are reasonably useful. The ends of each rod are

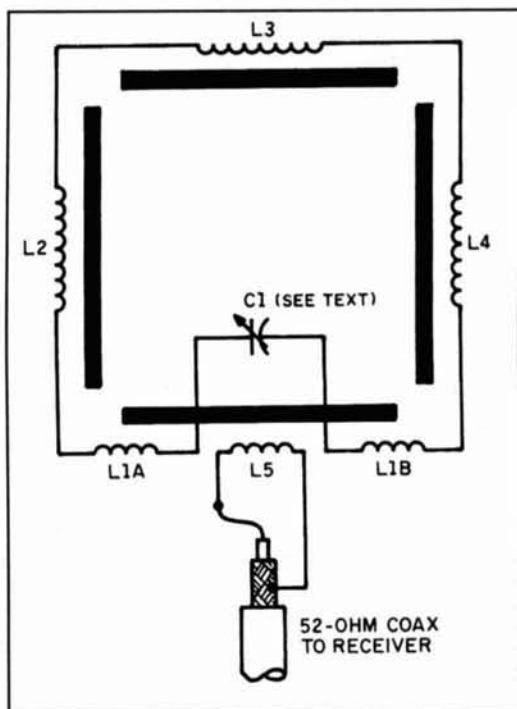


Figure 10. (A) The circuit diagram of the loopstick antenna.

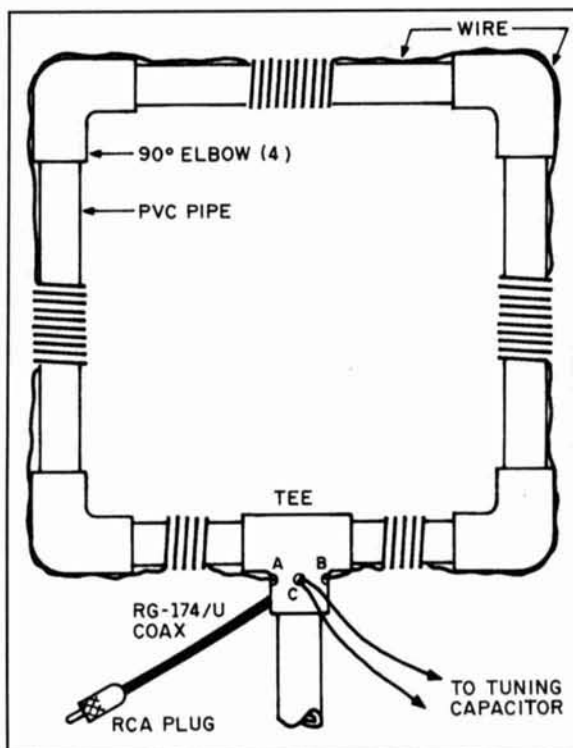


Figure 10. (B) The mechanical form of the loopstick antenna.

wrapped with sufficient black electrical tape to make them a snug fit inside the PVC pipe. I found that 3M tape required 13 or 14 wraps, depending on how snugly it was wrapped.

The inductors can be wound on the ferrite

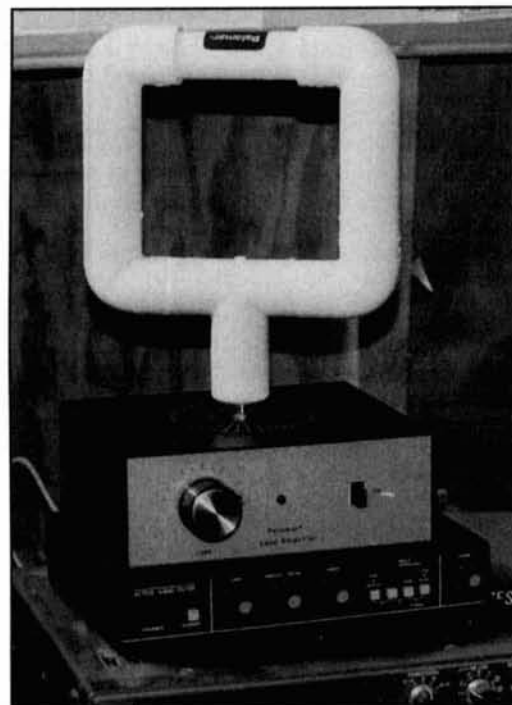


Photo G. Another Palomar Engineer's loop antenna in use at K4IPV.

rods, but it's easier, if you are experimenting, to wind them on the outside of the PVC pipe instead. Marris used four turns for L2, L3, and L4 for 80 meters, and two turns each for L1A and L2B. In the antenna I build for the AM broadcast band, ten turns for L2 through L4, and five turns for L1A and L1B, provided resonance in the 160-meter band and the upper edge of the AM broadcast band. Resonance in the low end of the broadcast band required a three section 3 x 365-pF variable capacitor.

A difference between my antenna and Marris' design is that I used a coupling loop (L5 in Figure 10A) wound on the ferrite rod used for L1A/L1B. The coupling loop is wound in the center of the rod, and connected to thin RG-174/U coaxial cable. The coax is routed through a small hole in the PVC "tee" piece that also serves for mounting. The completed antenna is shown in Photo E.

If you don't want to bother building a loop antenna, but still want to experiment with one, Palomar Engineers (P.O. Box 462222, Escondido, California 92046; 619-747-3343) makes several suitable models. Photos F and G show two Palomar products, along with their LA-1 loop amplifier, in use at my ham station in Virginia. ■

REFERENCES

1. Ken Cornell, W2IMB, "Loop Antenna Receiving Aid," *Ham Radio*, May 1975, pages 66-70.
2. Richard Q. Marris, G2BZQ, "Experimental Quadraform Ferrite Transmit/Receive Antenna," *Elektor Electronics USA*, November 1991, pages 57-59.

TECH NOTES

It's antenna time!

Once again, the warm spring weather brings us out of our houses and onto our roofs to check on the condition of our winter-ravaged antennas. This edition of *Tech Notes* presents two techniques you should find useful as you tend to your antenna farm.

Peter Bertini, K1ZJH, Senior Technical Editor

This first piece gives hints on how to estimate electrically small antenna resistance. It's reprinted with permission from the September 1992 issue of "Break-In".

Practical Estimation of Electrically Small Antenna Resistance

Bob Vernall, ZL2CA

Amateur antenna systems for the lower frequency bands generally have practical limits for size and height; consequently, the impedances of verticals can be highly capacitive, loops can be highly inductive and, in either case, radiation resistance can be very small. System losses can swamp the radiation resis-

tance, inevitably resulting in low transmission efficiency. These are characteristics of typical electrically small antennas.

I'll discuss an expedient way of estimating the total system resistance in situ, without needing to use an RF bridge or other type of measuring set. The method was developed primarily for checking LF antennas (in the 165 to 190-kHz band), but is also applicable to electrically small antennas in other bands.

It is assumed that the amateur station already has a tuning system for the antenna and a meter to monitor antenna current.

Relevant theory

At resonance, the antenna system can be considered to be a series resistive circuit, with the system resistance comprising the sum of radiation resistance, conductor resistance, earth resistance, and equivalent series resistance for tuning components and insulator leakage (see **Figure 1**). The value of each resistive component depends on a specific antenna and environment. The tuner takes care of canceling reactance and transforming system resistance to a suitable value for loading the transmitter.

For an antenna system with the tuner adjust-

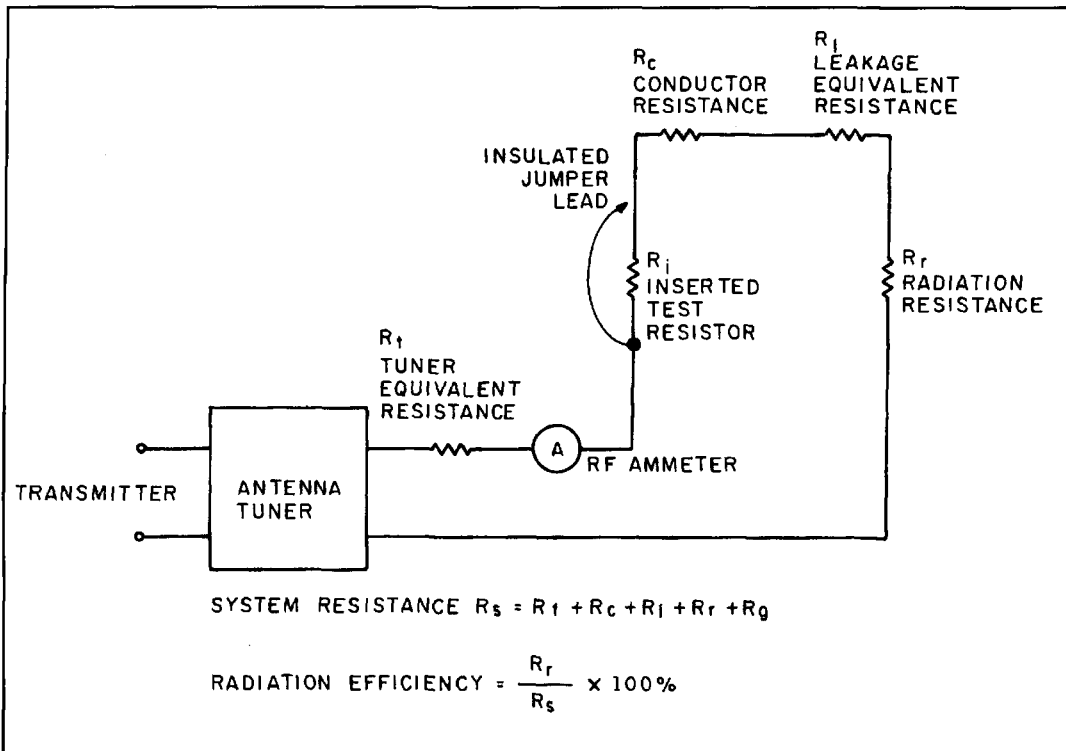


Figure 1. Test setup for measuring the velocity factor for coaxial cable or transmission line.

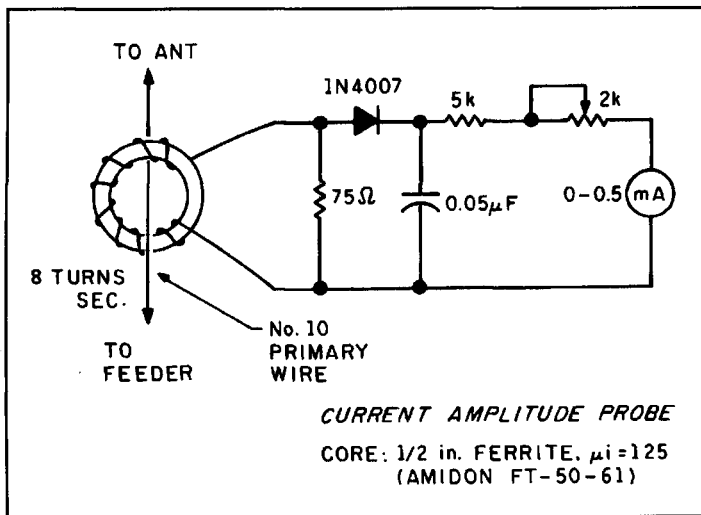


Figure 2. Measurements of RG58/U cable using methods outlined in the text.

ed for resonance, a certain antenna current will flow for a certain transmitter power. Adding a known series resistor will reduce the antenna current, for the same transmitter drive. The system resistance can be calculated from the following formula — an ohms law derivation:

$$R_s = d \cdot R_i / (1 - d) \quad (1)$$

where:

R_s is the (unknown) system series resistance.

R_i is the value of inserted series resistance.

d is the resulting current ratio, which is always less than 1, by dividing the resulting by initial current.

The inserted resistor should be placed at a point where the total antenna current flows (at or near a current node), which is generally the feed point for electrically small antennas. (And where does the current go? You may well ask, but that is beyond the scope of this article.)

Practical considerations

Use of the above formula can introduce errors if the current ratio is either too high or too low. If the current ratio is too high (d around a value of 1) then the numerator in the formula tends to zero, which leads to absurd numerical results. So the value of the inserted resistance should not be so low as to have difficulty observing a decrease in current. On the other hand, the value of the inserted resistance should not be so low as to have difficulty observing a decrease in the current. On the other hand, the value of the inserted resistor should not be so high as to cause significant change to the antenna impedance, otherwise any rematching of the tuner to restore conditions would introduce uncertainty. The best

conditions are suggested as when the observed current drops by around 10 percent (d around 0.9), as this gives sufficient change to observe a useful ratio, but not enough for mismatching to be significant. A further advantage of seeking only a modest current drop is to keep away from any nonlinear indication that may exist over lower parts of the meter scale.

So for vertical antennas, try starting with a known value of resistance of around 10 ohms. For loops try starting with a known value of resistance of around 1 ohm.

The current ratio can be observed directly if either the transmitter power or RF current meter sensitivity can be adjusted to obtain a convenient indication—such as 100 percent, or 1.0, on the scale. Then, after inserting the resistor, the current ratio can be observed directly. Using a jumper lead to temporarily short out the test resistor is a convenient way to find the current ratio, but the jumper lead must have excellent insulation. (Electrically short antennas can have several thousand volts present; consequently, the transmitter should be off while inserting or removing the resistor. The objective is to test the antenna system resistance, not human resistance!)

Carbon resistors are recommended for testing, as they are confirmed to not change the before and after tuning. Brief checks with wire-wound resistors show reactance which required retuning for resonance—even at LF. As well, they could have RF resistance that is different to the DC value, hence wire-wound resistors are not recommended for this test. Carbon resistors need to have adequate power rating so a series/parallel arrangement of lower-rating resistors can be used, so long as the net resistance is known (checking at DC is satisfactory).

Earth resistance decreases as moisture content increases, so checks done in different seasons will likely show significant variations in system resistance. Ground resistance in typical amateur situations can range from tens of ohms when many radial wires are used in damp soil to hundreds of ohms for a single earth rod in dry soil, and somewhere in between for any intermediate combination.

RF current monitoring

Antenna current can be monitored either with an in-line RF ammeter or by using a nearby inductively coupled meter. A broadband RF ammeter can be either a hot-wire ammeter (which is a rather rare item these days) or a home-built current transformer on a high permeability toroidal ferrite core. (See Figure 2. A wider display range could be obtained with a larger value variable resistor. Suitable insulation is needed for passing the antenna wire

through hole.) An inductively coupled meter can also be home-built¹ using a tuned winding on a ferrite rod and an enclosure with partial metal screening to reduce response to electrical pickup. With a band switch and suitable windings, it is capable of covering the entire LF/MF/HF range (a ferrite rod with multiple windings is the key component, the rest is like building a crystal set, but with a meter output instead of headphones).

Results

ZL2CA has used the described method to check antennas on three bands. For an LF vertical with flat top loading, a 15-ohm resistor reduced the current to 70 percent ($d = 0.70$), which indicates a system resistance of 35 ohms. A later check with a 9.5-ohm resistor gave $d = 0.78$ and a system resistance of 33 ohms, so the results are consistent. System resistance would be mainly ground loss followed by tuner loss (in which a big inductor with $+j2000$ ohms range is needed). The system resistance was lower than expected, but was checked during the winter, with wet soil conditions. The ground system uses three separate earth rods and fourteen buried radials of mixed length, going to all accessible parts of a suburban-sized property. In hindsight, testing this antenna with a 3.3-ohm resistor would be an optimum choice, but one was not on hand at the time.

A similar vertical to the LF arrangement was checked on 1.83 MHz and gave a system resistance of 28 ohms, which would be mainly earth loss. This same vertical checked on 3.6 MHz gave a system resistance of 135 ohms, which would be mainly radiation resistance as the electrical length is well over a quarter wavelength on this band.

REFERENCE

1. *ARRL Antenna Book*, 15th edition, American Radio Relay League, Newington, Connecticut, page 27-29.

George Wilson, WIOLP, passed along this piece which appeared in the Boston, Massachusetts Minute Man Repeater Association (MMRA) newsletter, fittingly called The Minuteman. It's excerpted from some earlier work done by Chet Smith, K1CCL, and published in QST in June 1979 on pages 27 and 28. In The Minuteman article, K1CCL's method is described by George Downs, W1CT; however, WIOLP has added some data of his own.

Measurement of velocity factor on coaxial cables and other lines

A simple direct method devised by Chet Smith, K1CCL, and described by George Downs, W1CT.

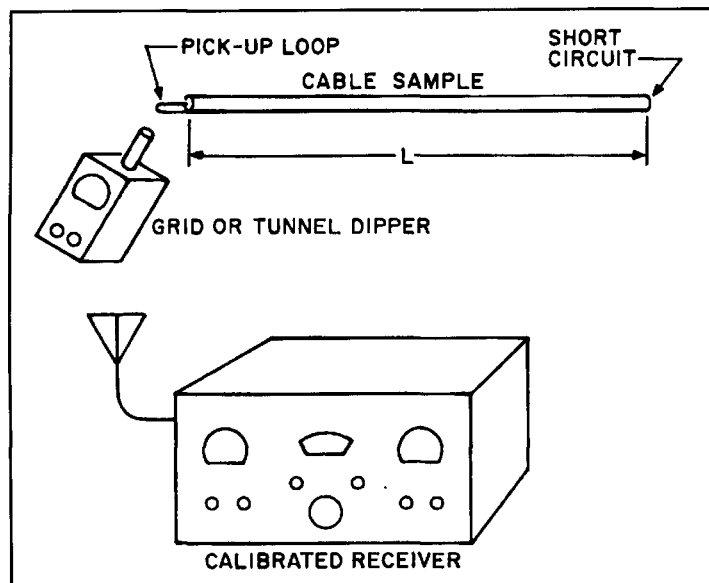


Figure 1. Diagram showing test arrangement and components of system resistance.

When cutting a piece of coax or twinlead to a specific fraction of a wavelength for impedance matching or phase shifting, the physical length of the cable is shorter than the free-space dimension calculated from the formula. This is because the wave travels slower in the cable than it does in free space. The ratio of the speed of the wave in the cable to the speed in free space is called the *velocity factor* (V_f). For example, the V_f for RG58/U is given in *The ARRL Handbook* as 66 percent. This means that instead of being 40.174 inches long, a half wave of RG58/U at 147 MHz would be $40.174 \times 0.66 = 26.515$ inches.

The problem is that, particularly with regard to the foam dielectric types, the rating in the tables is only an approximation. The actual cables vary as much as 10 percent from batch to batch, and sometimes (but much less so) from one part of a reel to another. In doing precise work, it becomes desirable to measure the V_f of the actual cable being used.

The K1CCL method is simple and straightforward:

- Take a piece of the cable to be used, preferably a piece more than 5 feet long.
- Short one end of the cable.
- Peel back just enough of the other end to solder to.
- Prepare three coils by winding no. 20 bare wire around the shank of a $1/4$ -inch drill. Make the coils 1 turn, 2 turns, and 3 turns. Space the turns $1/16$ -inch apart. Leave $1/2$ -inch pigtailed.
- Solder each of the three coils in turn between the shield and center conductor of the unshorted end of the coax, and determine the

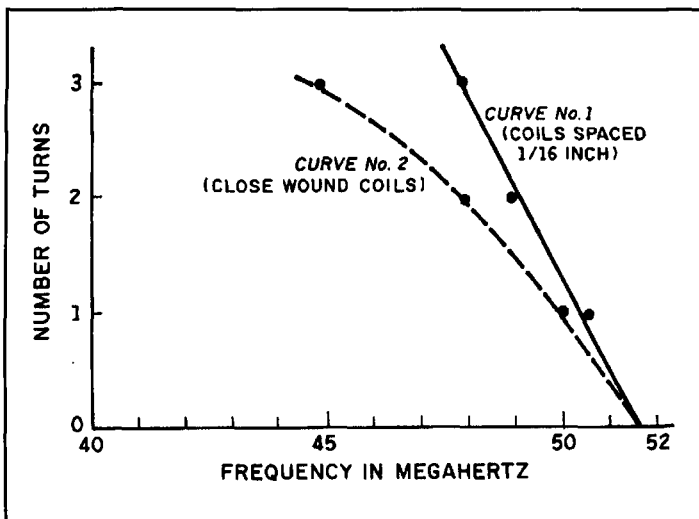


Figure 2. RF current transformer. (Source *Low Band DXING*, 1st edition, American Radio Relay League, Newington, Connecticut, 1987, pages 11–99.)

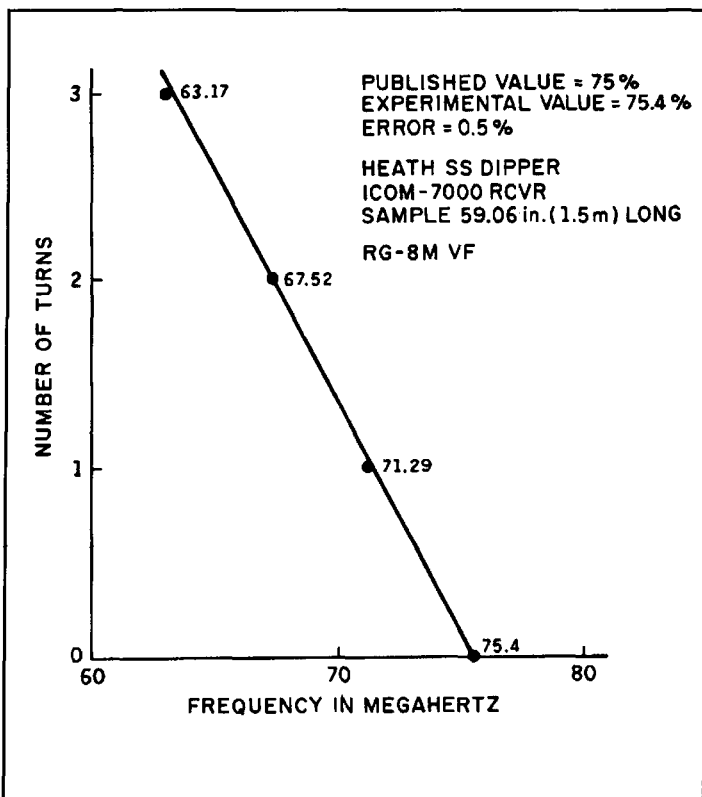


Figure 3. Measurements obtained by George Wilson, WIOLP, using the K1CCL method.

resonant frequency with a grid (or tunnel) dipper (see **Figure 1**). Use the minimum coupling that will give a positive reading. The calibrated receiver (or a wavemeter) is an option to give greater accuracy. Grid dippers are not usually that well calibrated.

- Plot your results (turns versus frequency) for 3 turns, 2 turns, and one turn on a piece of graph paper (see **Figure 2**). By eye and using a straightedge, French curve, or whatever tool you like, extrapolate to find the resonant frequency at zero turns. Call this frequency “F” (in MHz).

- Measure the overall length of the cable you have been using. Convert this length to meters (one meter = 39.37 inches). Call this length “L” (in meters).

- Substitute these values in the formula:

$$V_f = 2FL/3 \text{ percent}$$

where:

V_f is the velocity factor expressed in percent.

F is the resonant frequency at zero turns in MHz.

L is the length of the cable in meters.

For those who want an elegant solution, make the cable length (L) 1.5 meters (59.06 inches). The $2/3$ then cancels and the formula becomes $V_f = F$.

Here’s an actual example:

Cable : RG58/U

Length: $76 - 1/4$ inches = 1.937 meters

Data : $F_{3t} = 48$ MHz, $F_{2t} = 49$ MHz, $F_{1t} = 50.5$ MHz

F (extrapolated) = 51.7 MHz (See **Figure 2**.)

$$V_f \text{ (measured)} = 2FL/3 = \frac{2 \times 51.7 \times 1.937}{3} = 66.76 \text{ percent}$$

which is pretty close to the value that’s given in the handbook.

Those without a calibrated receiver in the 100-MHz range could use a cable sample of 4 meters or more to get the readings down into the HF range.

It’s important that the spacing between turns be the same for each of the three coils. The $1/16$ -inch spacing seems to give fairly linear results as shown on Curve No. 1 of **Figure 2**. Curve No. 2 was made with another set of coils. These were wound on a $1/4$ -inch drill shank of no. 22 enameled wire and were close spaced. The curve wasn’t linear, but the answer came out the same.

Figure 3 shows WIOLP’s attempt at measuring the velocity factor based on the published technique. Ed. ■

BUILD A HIGH-PERFORMANCE LOW-PROFILE 20-METER BEAM

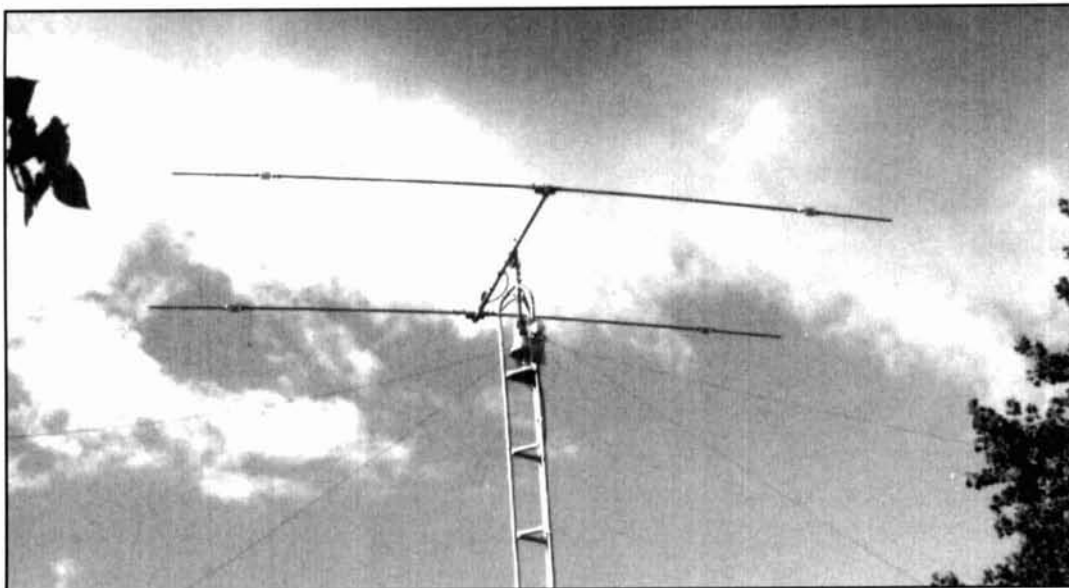
This neighbor-friendly antenna performs like a big beam

With the fast declining sunspot cycle, 20 meters remains one of the best choices for amateur radio communications. However, due to current marginal conditions, once successful wire and vertical antennas have become increasingly difficult to use on this band. In addition, local QRN can make DX almost impossible to work with these antennas. I soon discovered for myself the old Ham saying, "if you can't hear them, you can't work them."

For the past several years, I've tried some

compromise antennas that promised the world but delivered little. I needed good performance in a simple, but efficient antenna that would work well under local QRN. A full-size 20-meter beam was desirable but out of the question because of limited space, and a tower height of just thirty feet. While miniature antennas were commercially available for 20 through 10 meters, I decided against them because of their many compromises.

In the end I concluded that, for my application, it would be better to design and develop



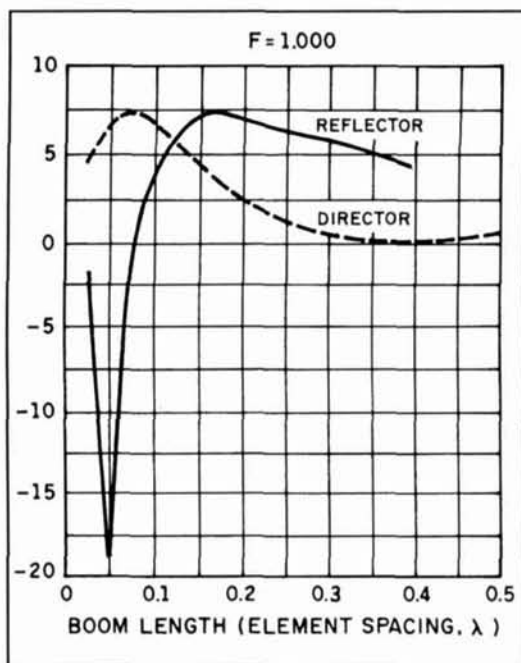


Figure 1. Isotropic gain for a two-element beam as a function of boom length (element spacing). Contrary to popular belief, shorter booms provide better gains. Theoretically, booms of 0.15 wavelength are desirable. However, booms of 0.1 wavelength can provide about 7.1 dBi.

my own antenna. The result is an easy-to-construct, two-element beam optimized for 20 meters. It's small and easy to install and rotate.

If you live in an apartment and have had a hard time convincing your landlord or neighbors that beams are beautiful, if you can only afford a tripod on your roof instead of a 100-foot tower in your back yard, or if you simply want to use your chimney as a support, this lightweight beam will help you obtain remark-

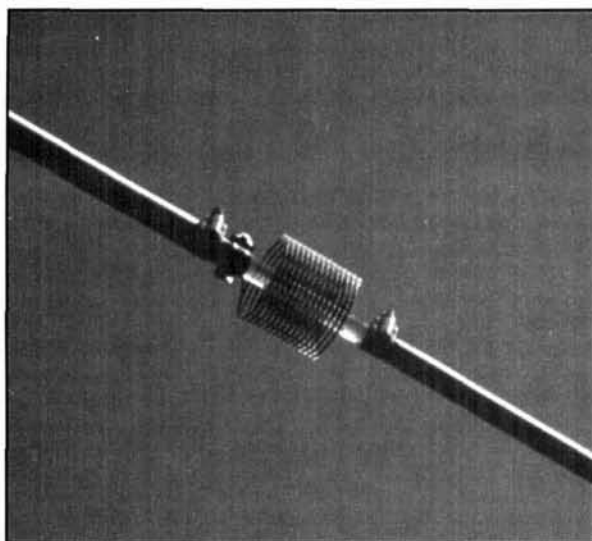


Photo A. Detail for the coil assemblies with their electrical and mechanical supports.

able results. The difference between this beam and the dipole, long wire, or inverted Vee you have been using will be substantial.

Design criteria

I started work on the project with a literature and product search which revealed that two-element short beams perform well when compared with full-size beams having the same number of elements. I then used a computer program to model the design. The reference literature stated clearly that in two-element Yagi designs, long booms are undesirable because gain falls off with boom length. This is primarily due to reduced excitation of the parasitic element as the spacing increases. Invariably, the largest gains are obtained at small spacing. The paper design showed a theoretical isotropic gain of 7.1 dBi for the two-element beam.

(Note: Antenna gain is referred to as the gain over a fictional isotropic antenna which has the same radiation intensity in any direction.)

According to the literature, if one wishes to obtain good gain with a fair front-to-back ratio over a bandwidth of say ± 10 percent, a boom length of about 0.15 wavelength will provide an optimum isotropic gain of about 7.5 dBi.^{1,2} However, experience proves that the boom's length can be reduced further. A more practical 0.1 wavelength will provide almost similar performance if care is exercised when tuning the elements (see **Figure 1**).

The literature search showed that efficient short beam antennas have been described by Walsh,³ by the ARRL's 1970 edition of *The Radio Amateur's Handbook*,⁴ and by several others. A short two-element beam known as the Trim-Tenna (TT-20) was manufactured briefly by Dentron Radio in the 70s.⁵

With the exception of the TT-20, these older designs used capacitive hats and gamma matches to achieve efficiency. I decided to take such a design, streamline its physical characteristics, and optimize it electrically for even better performance. To achieve this goal, the new beam uses very high-Q inductors for increased gain. The enhanced performance is also due to the proper placement of these inductors and a simple method of matching the driven element to the transmission line.

The final beam uses short, electrically lengthened elements of about 16 feet for the driven element and 14 feet for the reflector. The boom 0.1 wavelength on 20 meters (7 feet). I selected this length after considering the previously stated criterion. I also felt it was the best compromise between efficiency and bandwidth on one hand, and gain versus front-to-back ratio on the other.⁶

In the actual implementation of the beam, I

have optimized the peaks on the gain curves from **Figure 1** for the 0.1-wavelength spacing. I did this by adjusting the lengths of the two elements so that their coupled mutual reactance would be optimum at resonance. Consequently, the forward gain performance of the two-element beam approached that of a theoretical three-element beam. The only difference I noticed was in the front-to-back ratio, which is better for the three-element design. I rationalized that the limited front-to-back ratio (6 dB) was a good tradeoff for the reduced size of this antenna.

The key to the beam's success lies in the location of its coils. I minimized losses by placing the loading coils closer to the ends of the elements, where loss-resistance is at a minimum. This resulted in lower current distribution patterns that increase current flow on the elements rather than the coils, thus increasing gain. (Note: In some antennas loading coils are installed closer to the center of the elements and, therefore, can be lossier because of the higher current-distribution patterns that exist there.) I replaced the capacitance hats from the previous designs with simple aluminum tubing extensions. These were placed on the outer sides of the coils and mechanically supported with solid acrylic rods, also intended to provide DC isolation between the elements and their extensions (see **Photo A**).

The reflector is tuned lower in frequency (13.880 MHz) when compared with the driven element (14.200 MHz), providing improved front-to-side ratio. The element sizes were optimized through extensive experimentation. The front-to-side ratio for this beam tested at better

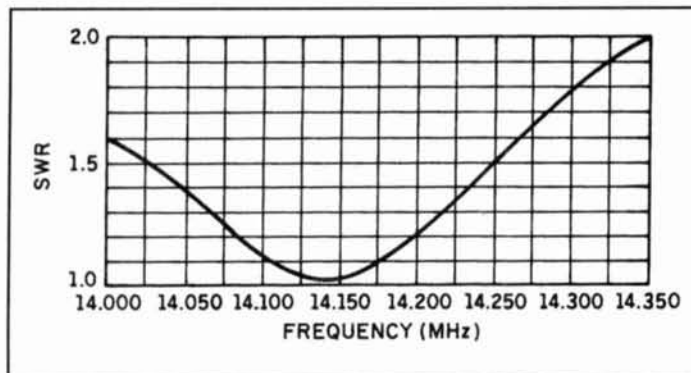


Figure 2. SWR curve for the two-element short beam as installed at WB3JZO.

than 18 dB. The beam has a minimum SWR of 1:1 and a maximum of 2:1 across the entire 20-meter band (see **Figure 2**).

Actual gain measurements were made between my QTH and a friend's house located 5 miles away. We placed the beam at the same height as a 20-meter reference dipole and took alternating measurements using an HP-8551 B spectrum analyzer as a field strength meter. We consistently measured a forward gain of 8 dB, although theory shows that the gain of a small beam over a perfect dipole is only 3 to 5 dB. Tests over the long path with DL3KCT and others produced consistently superior results when compared with wire or vertical antennas.

The difference between the calculated and measured gain values can be attributed to having to make the measurements in an environment other than free space. After all, it isn't possible to construct an absolutely perfect dipole; those exist only in theory. Ground con-

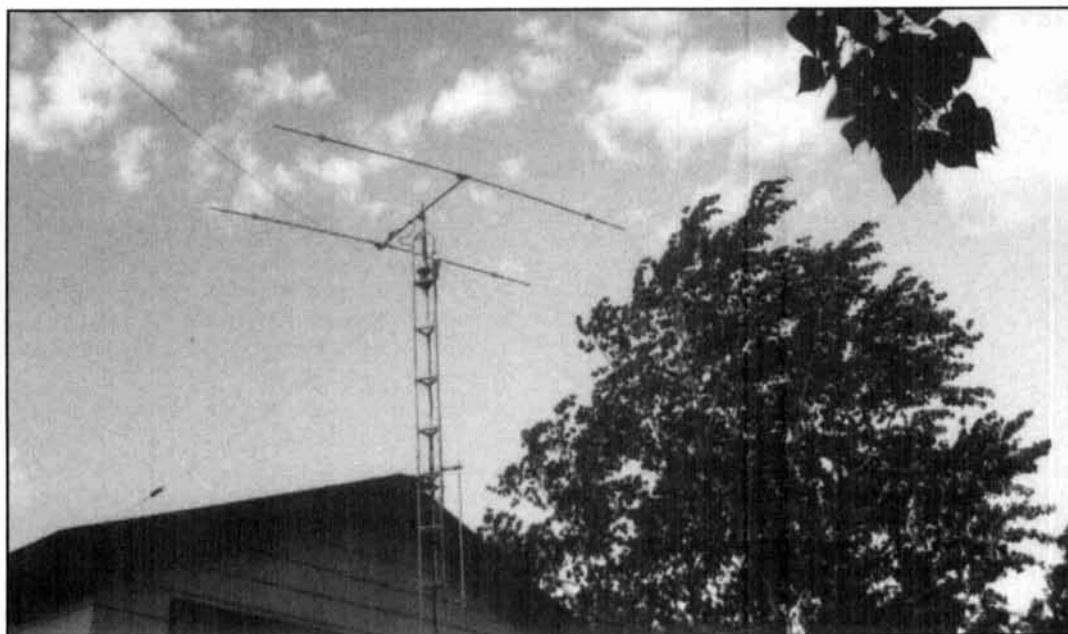


Photo B. Prototype installed on a 30-foot tower.

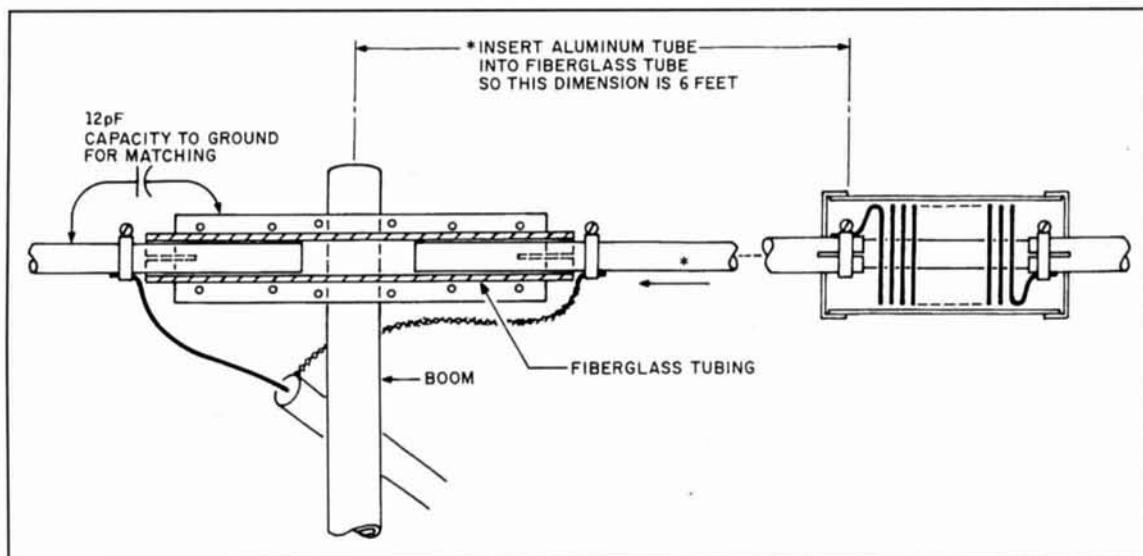


Figure 3. The beam is matched to 50 ohms through a simple technique that uses the natural capacitance of the driven elements to ground through the dielectric of the fiber glass support tube. This capacitance was optimized at 12 pF.

ductivity and proximity, the direction in which the antenna is aimed, interaction between the beam and dipole, and inability to construct a perfect dipole are all contributing factors to the difference in values.

Other design considerations

Mechanically, the beam uses seamless heavy-gauge hardened aluminum and has been tested over extended periods of time in the extreme climate of Minnesota. It has resisted winds of more than eighty miles per hour and temperatures ranging from -40 degrees Fahrenheit (also Centigrade) to +100 degrees Fahrenheit (or +38 degrees Centigrade) with no mechanical or electrical impacts. You could use softer aluminum tubing, like the kind found in hardware stores, at the cost of reduced mechan-



Photo C. A low-profile tripod installation..

ical strength (see **Photo B**).

I fed the system with RG/8 coaxial cable, matching it to 50 ohms through the natural capacitance of the driven elements to ground through the dielectric of the supporting fiber glass tubing. **Figure 3** shows the details.

Construction

Figure 4 gives physical specifications for the beam. The construction is straightforward. I used $1/16$ inch thick tempered aluminum tubing sold under the name ALCAD 6061-T6. This is the strongest aluminum alloy tubing commercially available and is especially intended for antenna work. In addition to aluminum and other elements, this alloy contains Titanium, Zinc, Chromium, Magnesium, and Copper.

I used several diameters of tubing in the beam. The boom was made of $1\ 1/2$ inch diameter tubing. The elements were constructed of a combination of several diameters, from 1 inch through $7/8$ inch, and $3/4$ inch. You may choose to simplify this design at the cost of mechanical strength. For the inserts supporting and isolating the coils, I used $5/8$ inch acrylic stock obtained from a local plastics house. The fiber glass insulating tube support for the driven element was $8\ 1/2$ inches long and $1/8$ inch thick. I inserted the two tubes for the driven element into the fiber glass tubing as shown in **Figure 3** to form two 12-pF capacitors to ground. You can substitute a fiber glass vaulting pole from a sporting-goods store for the fiber glass.

I made the coils from B&W no. 3029-3905-1 stock. They are constructed of no. 12 silver-plated wire. The front coils have sixteen turns ($11\ \mu\text{H}$), the back seventeen ($12\ \mu\text{H}$). The prototype beam didn't have coil covers. I observed

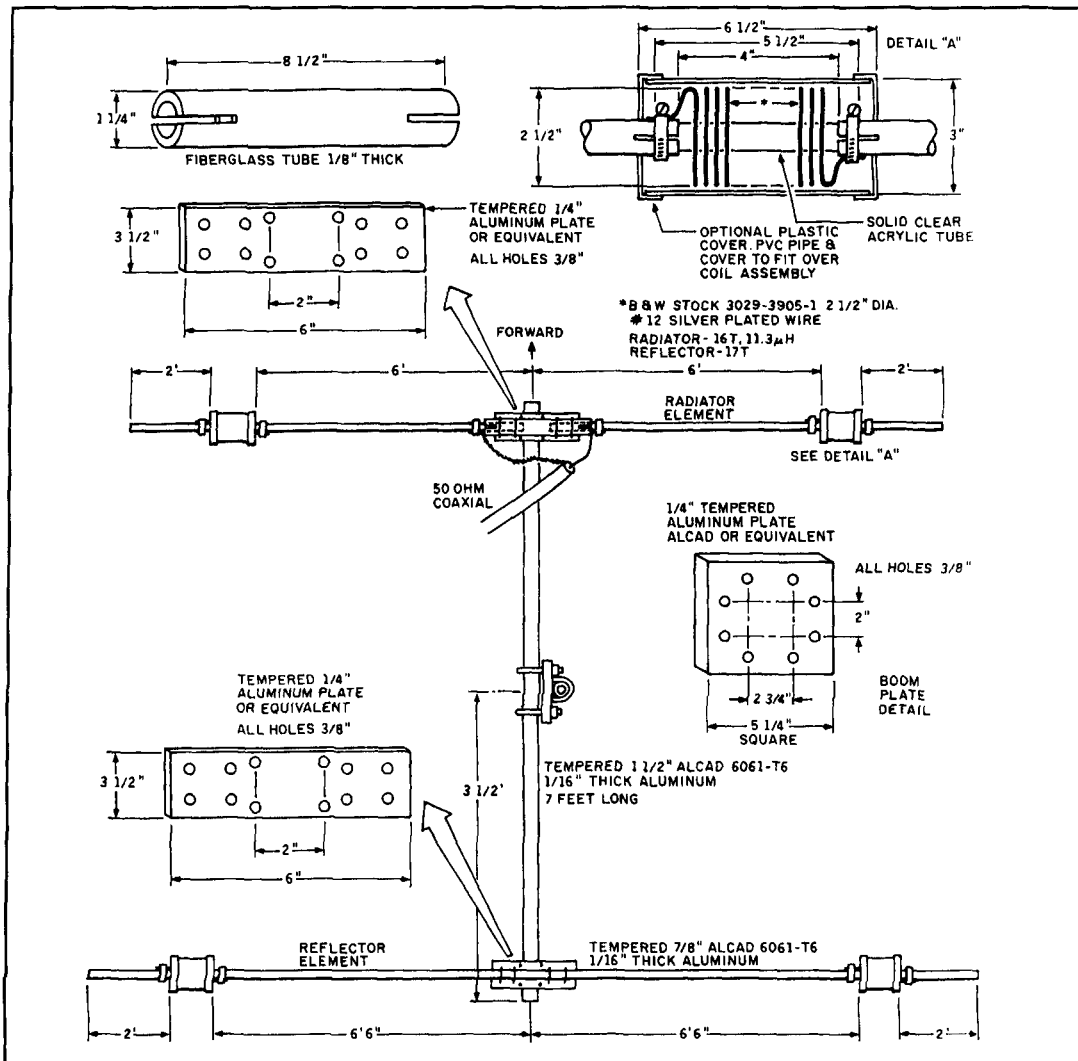


Figure 4. Physical specifications for the beam.

no signs of reduced performance on rainy days. In the final model, I fitted plastic covers made of PVC over the coils. Ample amounts of stainless U-bolts, saddles, and All-Stainless brand hose clamps were used throughout the design. Tubing was cut to size and slots were sawed at the ends for fitting.

Assembly and testing

For all parts and installation refer to **Figure 4**. The boom is assembled by installing the 5 1/4 inch mast-mounting plate in the center of the boom (3 1/2 feet away from either end). Tighten the 1 1/2 inch U-Bolts. Mount the element brackets at each end of the boom as shown in the **Figure 4**. Make sure you align these brackets at a 90-degree angle to the mast bracket plane and tighten using 2-inch U-bolts.

To assemble the reflector, insert the two 6' 6" x 7/8" tubes into one of the mounting brackets until they meet at the center of the 1" aluminum tube. Tighten the clamps and U-bolts.

To assemble the radiator, insert the other two

6' 6" x 7/8" elements into the fiber glass insulator as shown. Make sure that only 6 inches of each element's end is inserted into the assembly. Make sure the measurement for each side of the element is exactly 6 feet from the side of the boom to the tip of the pipe. Tighten all hardware except for the feed point wire clamps, which will be tightened later. With an ohmmeter, check that the two pipes of the radiator aren't DC shorted together or to the boom.

Insert the radiator and reflector loading coils into the elements as shown using the acrylic plastic inserts and the 2' x 3/4" extenders and clamps. Make sure that the smaller coils (16 T) are installed on the radiator and the larger (17 T) on the reflector. Tighten all clamps. This completes the general assembly.

Final tweaking

With the dimensions indicated, the resonant frequency of the beam should be around 14.150

Frequency (MHz) (SWR 1.01:1 minimum, 2:1 maximum)	14
Other frequencies (MHz) with tuner	18, 21, 24, 28
Number of elements	2
Theoretical isotropic gain (dBi)	7.1
Measured front-to-side ratio (dB) *	≥18
Measured front-to-back ratio (dB) **	≥ 6
Maximum continuous (or PEP) power (watts)	1500
Boom length (wavelength)	0.1
(feet)	7
Radiating element length (feet)	16
Director element length (Feet and Inches)	13.6
Turning radius (feet)	21
Calculated wind load (FT ²)	2.9
(Note: Wind load is specified for materials and construction recommended in this article. Other constructions may vary.)	
Weight (lbs.)	15

* Using HP 8551B spectrum analyzer located 5 miles from source.
** Using FT-ONE T/RX S-meter.

Table 1. Beam specifications.

MHz. However, the actual resonant frequency of the antenna may vary from one location to another because of height and local environmental conditions.

Tune the beam at least 10 feet above ground and preferably at the operational height. If you make your final adjustments close to the ground, you'll find the resonant points will change when the beam is erected at its final height. You can use a grid-dip meter loosely coupled to one of the coils to verify resonance. A VSWR meter will also work well at the transceiver point. You can optimize the antenna for any portion of the 20-meter band. To increase the frequency of the beam, remove half a turn from the radiator element coils to gain about 150 kHz. With variations in the electrical ground, I advise you to use a tuner at the transceiver point to allow for any impedance changes. The tuner will also help to operate the beam as a dipole on the higher bands.

Performance

The antenna operates well without a balun. You could use a 1:1 balun at the feed point; this will improve the radiated pattern somewhat. However, use only heavy-duty, toroid-type baluns for good reliability. Avoid ferrite bifilar-type baluns. Similarly, you could use 300-ohm balanced transmission line to feed this antenna, if you also use a tuner equipped with a balun.

This beam exhibits a good SWR ratio on 20 meters. However, if it isn't erected over a fixed groundplane, the variable water content in the soil can change the antenna's resonant frequen-

cy. (This problem occurs with most antennas.) Minimize this effect by using a tuner at the transmitting point.

Although I don't dispute the theoretical optimum height over a perfect ground, using this beam at heights of only thirty feet showed good low angle performance. This antenna far outperformed my dipoles and the like.

Specifications for the beam are given in **Table 1**.

Conclusion

This beam has provided consistently good performance on 20 meters when compared with my dipoles, inverted Vees, long wires, and verticals. The array does a better job, not only from a signal report point of view, but also by greatly attenuating the signals off the sides due to its outstanding front-to-side ratio of more than 18 dB. From an aesthetic point of view, the antenna keeps a very low profile. Although beauty is in the eyes of the beholder, this antenna has attracted positive remarks from neighbors and visitors alike.

REFERENCES

1. John D. Kraus, *Antennas*, McGraw-Hill, New York, 1950.
2. P. Viezbicke, "Yagi Antenna Design," NBS Technical Note 688. U.S. Department of Commerce, Washington, D.C., 1976.
3. David L. Walsh, "The W8 High Frequency Short Beam," *QST*, February 1969.
4. "A 20-meter Short Beam," *The Radio Amateur's Handbook*, American Radio Relay League, 1970, Pages 384-386.
5. Dentron Radio, TT-20 Instruction Manual.
6. James L. Lawson, *Yagi Antenna Design*, American Radio Relay League, 1991.

ONETWOES

A C-64 program for the analysis of single and double-resonant dipoles

Antenna analysis computer programs are popular these days. But those of us with old Commodore C-64 computers, may have trouble finding a program we can use. If you have a C-64 or a C-128, **Listing 1** provides a program with which you can calculate nearly all the properties of a center-fed loaded dipole with arbitrary dimensions and either one or two resonant frequencies of your choice. You can also modify the program to make it more convenient for special purposes.*

Once you've input and saved the program, type "RUN." The computer will ask for the length of the dipole (LW), and the separation of the loading reactors (LL) in wavelengths. Next, you'll be asked to specify the lower and upper resonant frequencies, FL and FU in MHz. For a single frequency, type in identical numbers for FL and FU. The program responds accordingly.

What the program calculates

After you answer a few obvious questions, the computer displays, either on screen or paper, data for plotting the pattern. It then displays the following quantities, in relation to your answers to the questions:

- XL(N), the reactance you need to insert on each side at the locations you've chosen. The subscript N = 1 for FL or FL = FU, and N = 2 for FU.
- R(N), the radiation resistance.
- RT(N), the sum of the radiation resistance and the loss resistances of the loading reactors. (The program doesn't include any other losses.)
- D(N), the directivity in dB relative to a half-wave dipole.

- G(N), the gain relative to a half-wave dipole in dB. (The efficiency in dB is G(N)-D(N).)
- The values of whatever inductances and/or capacitances are required.
- F0, the resonant frequency of the L-C circuits (with a double resonance).
- If you have specified a transmitter output power it displays IL, the rms current through the inductors, and VC, the rms voltage across the capacitors.
- LE, the length of the antenna in feet and LT, the distance between the resonators in feet.

The program provides all the information you might want—except for the frequency response. You determine this experimentally. The program is based on the simplest possible assumptions that can give these results. The section Physical Concepts discusses them in detail. Because of the program's simplicity, the absolute accuracy is poor. The output numbers don't have an absolute accuracy greater than 20 percent; however, this is enough to make antenna trimming easy. Usually, the adjustments needed are much smaller than 20 percent. Most of the differences between computed and experimental numbers are also much smaller.

Some definitions

With a single-frequency antenna, you can always find a reactance—inductive or capacitive—that can produce resonance. With an antenna for two frequencies, there are four situations you may encounter:

Type 1. When XL(1) is positive (inductive) and XL(2) is negative (capacitive), double resonance can be achieved by suitable parallel L-C circuits. The so-called "trap" antennas are among this type.

Type 2. When XL(1) is negative and XL(2)

*The author has included information and theory pertaining to such antennas in a soon to be published piece: "Double-Resonant Antennas Containing Loading Reactors (Traps)." Ed.

```

I REM "ONETWORES", Y. BEERS, W0JF,
BOULDER, COLO, SEPT 28, 1992.
90 POKE 53280,0: POKE 53281,0:REM DIMS
SCREEN AND BORDER.
100 PRINT "DO YOU WISH A PRINTOUT
(Y/N)?"
110 INPUT A$
125 PRINT " "
130 PRINT "DO YOU WISH PATTERN
DETAILS (Y/N)?"
135 INPUT B$
140 PRINT " "
145 PRINT "IF YOU WANT CURRENTS AND
VOLTAGES ENTER POWER: OTHERWISE 0"
150 INPUT P
155 PRINT " "
160 DIM J (90)
175 Q = 200: REM Q OF RESONATOR
INDUCTOR
180 ZA = 500: REM CHARACTERISTIC
IMPEDANCE OF SECTION BETWEEN FEED
AND REACTORS
185 ZB = 500: REM CHARACTERISTIC
IMPEDANCE OF SECTIONS BETWEEN
REACTORS AND ENDS
190 KZ = 20: REM NR OF INCREMENTS IN
LENGTH
195 DX = 10: REM INCREMENT IN ANGLE
(DEG)
196 REM DX MUST BE 90 DIVIDED BY AN
INTEGER
200 PRINT "ENTER RES FREQS, FL AND FU,
(MHZ) (FU>=FL)
202 INPUT FL,FU
203 F1=2*PI*FL*1E+6: F2=2*PI*FU*1E+6: REM
CONVERTS TO RAD/SEC
204 PRINT " "
206 FR = F1/F2
208 PRINT "ENTER DIPOLE LENGTH IN FU
WAVELENGTHS"
210 INPUT LW
214 LE = LW*936/FU: REM LENGTH OF
DIPOLE IN FEET WITH FU END CORRECTION
216 LW = 2*PI*LW: PRINT " ":REM ANGULAR
LENGTH
218 PRINT "ENTER SEPARATION OF
LOADING REACTORS IN FU WAVELENGTHS
220 INPUT LL
222 LT = LL*984/FU: REM SEPARATION OF
LOADING REACTORS IN FEET
224 LL=2*LL: REM ANGULAR SEPARATION
225 PRINT " "
226 KX = 90/DX
228 DX = PI*DX/180
230 IF A$ = "N" THEN 236
232 OPEN 15,4
234 CMD 15
236 IF FL = FU THEN 243
238 N2 = 2
240 GO TO 247
243 N2 = 1
247 FOR N = 1 TO N2
250 IF N = 2 THEN 270
253 A=FR*LL/2: B = FR* (LW-LL) / 2
257 F = FL: AL = A: FF = F1
260 REM A IS ANGULAR DISTANCE FROM
CENTER FEED POINT TO LOADING REAC-
TOR
263 REM B IS ANGULAR DISTANCE FROM
LOADING REACTOR TO END
267 GO TO 277
270 A = LL/2: B = (LW-LL) / 2
273 F = FU: AU = A: FF = F2
277 X2 = - ZB/ (TAN(B))
280 XL (N) = - X2 - ZA*TAN (A)
287 DZ = (A+B) / KZ: REM INCREMENT IN
LENGTH
290 KA = A/DZ: KB = B/DZ: K = KA + KB
292 IF A = 0 THEN 298
294 B2 = -ATN (ZB/ (XL (N) +X2)): REM
EQUIV ANG LENGTH OF LENGTH B WITH XL
ON IT
296 GO TO 380
298 B2 = B
380 IF B$ = "N" THEN 400
385 PRINT "PATTERN FOR FREQ (MHZ) = " ; F
386 PRINT " "
390 PRINT "ANGLE", "AMPLITUDE"
395 PRINT " "
397 JM = 0
400 FOR HX = 1 TO KX
410 ET = 0
420 X = (HX - 0.5) *DX: REM ANGLE OF
FIELD POINT
470 FOR HZ = 1 TO K
480 Z = (HZ - 0.5) *DZ: REM POSITION OF
CURRENT ELEMENT
500 IF HZ>KA THEN 540
505 I1= SIN (A + B2 - Z) / (SIN ( A + B2))
510 E = 2*I1*COS (Z * COS(X)) *SIN (X) *DZ
520 GO TO 600
540 I2 = (SIN (B2)) * (SIN (A+B-Z)) / ((SIN
(A+B2)) * (SIN (B)))
550 E = 2*I2*COS (Z*COS (X)) *SIN (X) *DZ
600 ET = ET + E
605 J (HX) = ET*ET
607 IF JM>J (HX) THEN 640
608 JM = J (HX)
640 NEXT HZ
645 XD = 180*X / PI
670 IF B$ = "N" THEN 800
679 ET = ET*1E3:ET=INT (ET) :ET=ET*1E-3
680 PRINT XD,ET
800 NEXT HX
900 R (N) = 0
910 FOR HX = 1 TO KX
915 X = (HX - 0.5) *DX
920 R7 = 4*PI* (SIN(X)) *J (HX) *DX: REM
CONTAINS FACTOR OF 2 FOR INT. TO 180
DEG.
930 R (N) = R (N) + R7
940 NEXT HX
950 D (N) = 4*PI*JM / R (N)
954 PRINT " "
955 D (N) = 73*D (N) / 120: REM CONVERTS
FROM ISOTROPIC TO HALF WAVE REFER-
ENCE
965 R (N) = R (N) *2.3961858: REM NORMAL-
IZES R TO GIVE 73 OHMS FOR HW DIPOLE
990 NEXT N
995 IF FL = FU THEN 1700
1000 PRINT " "
1010 X5 = XL (1) *F1-XL (2) *F2:X6 = XL (1)
*F2-XL (2) *F1
1015 IF XL (2) *XL(1)) >0 AND XL (2) <XL (1)
THEN 1698: REM COND. FOR NO SOLUTION.
1020 IF XL (2) >0 AND XL (1) <0 THEN 1050:
REM COND. FOR SERIES L-C

```

Listing 1. ONETWORES program listing.

```

1022 IF XL (1) *XL (2) >0 AND XL (2) >XL (1)
THEN 1024: REM COND. FOR 2 SOLUTIONS.
1023 GO TO 1030
1024 PRINT "DO YOU PREFER PARALLEL
L-C (P) OR SERIES L-C (S)?"
1025 INPUT D$
1026 ""
1027 IF D$ = "S" THEN 1050
1030 F0 = SQR (ABS(F1*F2*X5 / X6))
1032 X0 = XL (1) * (F1*F1-F0*F0) / (F1*F0)
1035 REM F0 = RES FREQ OF RESONATOR IN
RADIANS/SEC
1040 GO TO 1070
1050 F0 = SQR (ABS(F1*F2*X6/X5))
1060 X0 = XL (1) *F0*F1/ (F1*F1-F0*F0)
1070 X0 = ABS (X0)
1080 L = X0/F0: C = 1/ (X0*F0)
1100 FOR N = 1 TO N2
1110 IF N = 2 THEN 1140
1120 F = FL: FF = F1: A = AL
1130 GO TO 1161
1140 F = FU: FF = F2: A = AU
1161 PRINT ""
1190 IF XL (2) >0 AND XL (1) <0 THEN 1250
1192 IF D$ = "S" THEN 1250
1200 Y = L*C*FF*FF -1
1205 RP = Q*X0
1210 Y2 = L*L*FF*FF+RP*RP*Y*Y
1215 RS = RP*L*L*FF*FF/Y2: REM SERIES
RESISTANCE OF RESONATOR (PAR L-C)
1220 GO TO 1290
1250 RS = X0/Q: REM RES OF RESONATOR
SERIES L-C)
1290 RT = R(N) + 2*RS* (COS (A))* (COS(A)):
REM TOTAL EQ RESISTANCE AT FEED
1295 EF = R(N) / RT: REM EFFICIENCY
1300 X3 = 2*XL(N)*(COS(A))*(COS(A))
1310 X3 = ABS (X3)
1351 PRINT ""
1359 XL (N) = INT (XL(N))
1360 PRINT "F(MHZ)=";F; "REACTANCE=";
XL (N)
1364 G = D(N) *EF
1366 G = 4.342*LOG(G) : D(N) = 4.342*LOG
(D(N))
1367 D(N) = D(N) *1E+2:D(N) = INT (D (N)) : D
(N) = D (N) *1E-2
1368 G=G*1E+2: G=INT (G) : G=G*1E-2
1369 R(N) = R(N) *1E+2:R(N) = INT (R(N)) :
R(N) = R (N) *1E-2
1370 RT=RT*1E+2:RT=INT (RT) :RT=RT*1E-2
1373 PRINT "RAD RES="; R(N); "TOTAL
RES="; RT
1380 PRINT "D (DB) ="; D(N); "G(DB)="; G
1390 IF P=0 THEN 1495
1395 IF XL (2) *XL (1) >0 AND XL (2) <XL (1)
THEN 1450
1396 IF D$ = "S" THEN 1450
1400 VC = COS (A) *COS (A) *ABS (XL(N))
*SQR (P/RT)
1410 IL = VC/ (L*FF)
1430 GO TO 1465
1450 VC = COS (A) *COS (A) * (1/ (C*FF))
*SQR (P/RT)
1460 IL = COS (A) *COS (A) *SQR (P/RT)
1465 VC = INT (VC)
1470 IL = IL*100:IL=INT (IL) : IL=IL/100
1480 PRINT "CAP VOLTS="; VC; "IND CUR="
; IL
1495 NEXT N
1500 PRINT ""
1600 F0 = F0* (1/ (2*pi)) *1E-6: REM IN MHZ
1601 F0 = F0*1E+2:F0=INT(F0) :F0=F0*1E-2
1610 C=C*1E+14:C=INT (C) : C=C*1E-2:REM C
IN PF
1620 L = L*1E6: REM IN MUHY
1621 L=L*1E+2:L=INT (L) :L=L*1E-2
1630 PRINT "RESONATOR PARAMETERS"
1631 PRINT ""
1633 IF XL (2) >0 AND XL (1) <0 THEN 1638
1634 IF D$ = "S" THEN 1638
1635 PRINT "L AND C IN PARALLEL"
1636 GO TO 1640
1638 PRINT "L AND C IN SERIES"
1640 PRINT "RES FREQ (MHZ) =" ;F0;"Q=";Q
1650 PRINT "C(PF) =" ;C;"L(MUHY)=" ;L
1658 LE=LE*1E+2:LE=INT (LE) :LE=LE*1E-2
1659 LT=LT*1E+2:LT=INT (LT) :LT=LT*1E-2
1690 PRINT "LENGTH (FT) =" ; LE; "REACTOR
SEP (FT) =" ; LT
1692 LW=LW/ (2*pi) :LL=LL/ (2*pi)
1693 PRINT "LENGTH (WLU) =" ; LW; "SEP
(WLU) =" ; LL
1695 GO TO 1900
1698 PRINT "NO SIMPLE L-C CIRCUIT CAN
PRODUCE AN IMPEDANCE MATCH"
1699 GO TO 1900
1700 IF XL (1) <0 THEN 1760
1705 L = XL (1)/F1
1706 L = L*1E+6: REM CONVERTS TO MUHY
1710 RT = R (1) +2*COS (A) *COS (A) *XL (1)/Q
1715 G = R (1) *D (1) / RT
1720 RT= 10*RT:RT=INT(RT) : RT=RT / 10
1735 IF P = 0 THEN 1748
1740 IL = COS (A) *COS (A) *SQR (P / RT)
1741 IL = IL*100:IL=INT (IL) : IL=IL / 100
1745 PRINT "IND CUR="; IL
1748 L = L*1E+3:L=INT (L) : L=L*1E-3
1749 XL (1) =INT (XL(1))
1750 PRINT "REACTANCE="; XL (1); "L
(MUHY) =" ; L; "Q="; Q
1758 GO TO 1774
1760 C = -1/(F1*XL (1))
1761 IF P = 0 THEN 1768
1762 VC = (1/(C*F1)) *SQR (P/R (1))
1763 VC =INT (VC)
1765 PRINT "CAP VOLTS="; VC
1768 C=C*1E+14:C=INT (C) :C=C*1E-2:REM C
IN PF
1770 XL (1) =INT (XL(1))
1771 PRINT "REACTANCE="; XL (1); "CAP
(PF)="; C
1772 G = D(1): RT=10*R (1): RT=INT (RT) :
RT=RT / 10
1774 D (1) = 434.2*LOG (D(1)): D(1) = INT
(D(1)) : D(1) = D(1) *1E-2
1776 G = 434.2*LOG (G) :G=INT (G) :
G=G*1E-2
1777 R (1) = 10*R (1) : R (1) = INT (R(1)): R (1) =
R (1) / 10
1778 PRINT "RAD RES="; R (1); "TOTAL R=";
RT
1779 PRINT "D (DB) =" ; D (1); "G (DB) =" ; G
1780 LW=LW/ (2*pi) : LL=LL/ (2*pi)
1782 LE=LE*1E+1:LE=INT (LE) : LE=LE*1E-1
1784 LT=LT*1E+1:LT=INT (LT) : LT=LT*1E-1
1786 PRINT ""
1788 PRINT "LENGTH (FT) =" ; LE; "REACTOR
SEP (FT) =" ; LT
1790 PRINT "LENGTH (WLU) =" ; LW; "SEP
(WLU) =" ; LL
1900 IF A$ = "N" THEN 2000
1990 CLOSE 15
2000 END

```


is positive, double resonance can be achieved with series L-C circuits.

Type 3. When $XL(1)$ and $XL(2)$ are of the same sign and when $XL(2)$ is larger or less negative than $XL(1)$, double resonance can be achieved by either parallel or series L-C circuits. In this case, ONETWORES gives you a choice.

Type 4. When $XL(1)$ and $XL(2)$ are of the same sign, and when $XL(2)$ is smaller or more negative than $XL(1)$, it's not possible to achieve double resonance.

A future article will give some examples of type 2 and 3 antennas, which exist but aren't well known. It will also discuss some common type 4 situations.

Typing and saving the program

ONETWORES is the merger of several shorter programs I've developed over a five-year period. Using ONETWORES, I can compare the properties of a double-resonant antenna with what they would be if tuned up for single-frequency operation, using the same program. However, in my files, I have a separate predecessor program that works with only one resonant frequency. It's faster than ONETWORES when used with a single frequency. Therefore, I suggest you type in the program in stages, some of which you may wish to save as separate programs.

The paragraphs that follow specify the stages I suggest you use. Numbers in parentheses denote statement numbers. A hyphen between two numbers denotes a consecutive set of statements. Commas separate individual statement numbers. If you wish, you can omit all REM (reminder) statements.

Stage 1. Input information (1–225). These statements include all of the information that the program needs to run. You must enter some of the information with every run; however, the rest of the data in declarative statements (175–200) that remain unchanged until you alter them deliberately.

First, it asks three questions: whether you want screen or printer output, and whether you require a display of the pattern. If you want current through inductances and voltage across capacitors, you must give the transmitter output power.

Calculating the data for the pattern takes a long time. These data are needed for later calculations. If you opt for screen display, you'll find the screen blanks during these calculations. Therefore, I suggest that you display the pattern, even though you may not be interested in it, just to monitor computer action.

Five declarative statements follow (175–196). (175) the Q of inductors, is self-explanatory. (I assume that the losses in capacitors are negligible.)

My model makes the basic simplifying assumption that load reactances required for resonance can be calculated by assuming that each half of the antenna acts as a transmission line of characteristic impedance Z_0 . Until I started to write this article, I had a single statement $Z_0 = 500$, instead of (180, 185). However, in special applications like vertical and "L" antennas, there's some logic for using different values for the sections inside and outside the loading reactors. I've modified the program to provide this option. In the section **Physical Concepts** I pay a lot of attention to Z_0 . Mathematically, if $Z_A = Z_B = Z_0$, it acts as a scale factor for computed reactances.

Later the program, when determining the pattern, assumes an angle, X , between the axis of the antenna and a line to the observer—assumed to be the "far" field. The program then calculates the field at the observer by assuming a current distribution in the antenna whose half angular length is Z . It breaks this up into KZ sections, computes the field for each section, and adds them up. Angle X then increases in steps of DX .

There's a tradeoff between numerical accuracy and running time. The values for KZ and DX here are a reasonable compromise. However, to debug the program or obtain rough answers, reduce the running time using $KZ = 5$ and $DX = 18$ or 30 .

(200–225) contain the input of the frequency and dimensional information. These data are processed because the forms convenient to the user must be converted into forms acceptable to the computer. Frequencies in MHz are converted to radian/sec (203). Later, in (253 and 270), the dimensional information is converted into the angular distance in radians from the center feed point to one of the loading reactors, **A**, and the angular distance from a reactor to the end of the antenna, **B**.

This is the portion of the program you are most likely to modify. The user chooses the information that's entered every run via "INPUT" statement. Other needed data are contained in declarative statements like (175-196), which are a semi-permanent part of the program. You're likely to make a series of runs in which everything but one parameter is held constant. In these cases, you can save time by replacing some of the "INPUT" statements with declarative ones. The easiest way to accomplish this is to place the declarative statement and a "GO TO" statement before the existing "INPUT," thus bypassing it. To restore the use of the original "INPUT," delete the

statements you have inserted.

Stage 2. (226–800, 990, 1900–2000). After typing in these statements, you should have a working program for calculating the numbers needed for plotting the pattern. This stage contains three nested FOR-NEXT loops (called “DO” loops in Fortran). The outside one (247, 990), with the index N, controls the frequency used. The middle one, with the index HX (400, 800), selects the angle, and the inner one, with the index HZ (470, 640), selects the section of the antenna whose field, E, and intensity, E*E, are calculated. I1 (505) gives the current distribution in the inner portions of the antenna; I2 (540) gives the current distribution in the outer portions. The KZ loop sums the contributions from the separate sections. Although you skip (950–965), you need (990) to close the N loop. (1900–2000) are needed to end the program.

Stage 3. The subroutine in statements (900–965) is a small one, and you may want to combine it with Stage 2. It uses another KX loop imbedded in the same N loop to sum E and E*E over angle. (Actually it is integrating these expressions over solid angle.) The results are then used to calculate the directivity (as a ratio) and the radiation resistance. The fact that E is in arbitrary units has no effect on the directivity, but a constant of proportionality must be supplied on the E*E sum to make the radiation resistance equal the textbook value of 73 ohms for the half-wave dipole.

In the complete program, these quantities are displayed by later “PRINT” statements. You may want to insert the following temporary statement to display them and make this stage a working program:

```
980 Print “Rad Res=”;R(N);“D=”;D(N).
```

which lets you display these quantities.

Stages 4 and 5. The remainder of the program is concerned mainly with finding the actual inductors and capacitors you need to supply the required reactances. This portion displays the final results of the entire program. With (995), the program divides into two parts that are independent except for the closing statements (1900–2000).

The statements in (1000–1695) pertain to double resonant antennas, while those in (1700–1790) pertain to single resonant ones.

I suggest that you enter (1700–1790) as **Stage 4**, skipping those for double resonant antennas. This will give you a complete program for a single-frequency antenna. You can make the program slightly shorter and easier to run by deleting or modifying a few of the previous statements; I leave the details to you.

Statement (1720) is one of several similar statements in the program. Its purpose is to

“format” the output; that is, round off undesirable digits in the output. Without it, the output number would contain about eight digits—most of which would be of no significance.

The statement consists of three commands. The first multiplies the number to be displayed by an appropriate power of 10 (in this case, the first power), so the digits to be retained are on the left side of the decimal point. Next, the INT (integer) command deletes all the digits to the right of the decimal point. Finally the third command multiplies the number by the inverse power of 10, restoring the decimal point to its original position.

To complete ONETWOES, I suggest that you enter (1000–1699) as **Stage 5**.

When you run the program with a Type 3 antenna, the computer asks if you want a parallel (P) or series (S) L-C circuit. If you want data on both, you must run the program twice. I turn next to a few “bugs” in this program.

Program “bugs”

In the midst of a run, the program may stop and give you an error message indicating you have divided by zero. This could happen when the length LW is exactly 0.5 or 1.5 wavelengths (with LL=0). The physical situation is such that the results must be finite; consequently the fault lies in the mathematics.

The standard way to deal with a mathematical infinity is to approach it as a limit. With LW = 0.499 and LW = 0.501, the program runs and provides finite answers. In principle, one should take the average of the two. However, the differences are trivial compared to other errors. Therefore, in practice, I have considered a half-wave dipole to be 0.499 wavelengths long, and have used this value in determining the constant in 965.

You could encounter a similar problem when LL is exactly 0.5. When I wish to specify that the loading reactors are a half wavelength apart, I use LL = 0.499.

In a computer program, it’s often possible to circumvent an infinity by using “IF” and “GO TO” statements, as, for example, in Lines 292–298. Because, I haven’t found a practical way to do so with the present cases, I advise you how to deal with them if they occur.

Physical assumptions

This program is based upon four assumptions: (a) that the antenna is “thin” and that the currents within the separate sections vary with distance as portions of sine curves; (b) that the radiation resistances are those of antennas in empty space; (c) for calculating loading reactances that the antenna can be represented by a

transmission line of characteristic impedance that's independent of length and frequency; and (d) it's necessary to make a small "end" correction because the physical and electrical lengths aren't equal. These are all approximations, and, strictly speaking, (b) and (c) are inconsistent.

Current distribution. A sinusoidal variation of current with distance is valid for short antennas constructed from thin wire.¹ This approximation is often used and many published patterns are based on it. The approximation neglects the fact that the current is attenuated as it travels out from the feed point as the result of radiation and conductor losses. The approximation is valid with relatively short antennas.

Radiation resistance. For this calculation, I assume that the antenna is in free space and that the wire is thin. In (965), I have adjusted a constant of proportionality to make the radiation resistance of the half-wave dipole the textbook value of 73 ohms.² As a check on the correctness of the program, I have reproduced the values in Moxon's graph as accurately as I can read the numbers.³

While writing this manuscript, I received the Summer 1992 *Communications Quarterly* containing an article by Charles J. Michaels, W7XC.⁴ Mr. Michaels has calculated many of the same things I have. In comparing a few sample values, I note that my radiation resistances are consistently higher. (Michaels doesn't display them directly, but it's easy to determine them by multiplying the total resistances by the efficiencies. In the few cases I checked, I, correctly, obtained the same value from Q = 100 data as from Q = 300 data.)

I think the discrepancies can be traced to his assumptions that the half-wave dipole radiation resistance is about 65 ohms, which is more correct, than the 73 ohms I have used, because it compensates for end effects. Michaels also considers wire diameter effects. If you prefer 65 ohms, reduce the constant in (965) in the ratio 65/73.

My calculations do not give consideration to the effect of the ground. ONETWORES gives free-space results.

Reactance calculation. For calculating reactances, I assume that one half of the antenna can be represented as a wire parallel to a perfectly conducting earth. The outer portion appears at the position of the reactor, as a section of open circuited line. By standard theory it has a reactance:

$$Z2 = -Z0/(\tan(B)) \quad (1)$$

The view from the same point of the inner section is that of a transmission line terminated by half the output impedance of the cable to the transmitter, which, as an approximation, can be

neglected. In other words, this line is short-circuited. Its reactance is:

$$Z1 = Z0*\tan(A) \quad (2)$$

Then, for resonance, the reactor must have the reactance:

$$ZL = Z0((1/(\tan(B)) - \tan(A))) \quad (3)$$

The reactance calculations in the program are based on **Equation 3**. There are theoretical estimates of Z0, but my value of 500 ohms is determined by experimentation.

By standard theory, if the transmission line has no ohmic losses:

$$Z0 = \text{SQR}(L/C) \quad (4)$$

where L and C are the inductance and the capacitance per unit length.

One theoretical estimate comes from the formula for a wire of diameter D parallel at a distance H from a plane conducting earth, which has been derived from **Equation 4**:

$$Z0 = 138 \log(4H/D) \quad (5)$$

For no. 14 wire 20 feet high, typical of my experiments, **Equation 5** gives 576 ohms. For Michael's antenna of no. 10 wire at 40 feet, it gives 590 ohms.

Of course, the use of this formula is contradictory to the assumption that the antenna is in the free-space situation used in the radiation resistance measurements. The theory of the reactance of a dipole in free space is very complicated, mathematically using what I call "mode theory."^{5,6}

Because the capacitance and inductance per unit length vary, Z0 depends on the length. References 3 and 6 summarize the results from this theory in the form of straight-line graphs. The values from Reference 6 are a few percent higher. In these graphs, Z0 pertains to the whole dipole, while I consider only one side here. Therefore, I divide by two. Including this factor, a straight line representing their average can be represented by the following:

$$Z0 = 118 \text{ LOG } (L/R) - 63 \quad (6)$$

For a no. 14 wire 30 feet long, which is typical of my work, **Equation 6** gives 474 ohms. For Michael's half-wave dipole made of no. 10 wire, it gives 497 ohms. This is in the same "ball park" as the previous value, but about 15 percent lower.

Equation 6, which shows a dependence on length, indicates that my hypothesis isn't correct. But the dependence is logarithmic, which

is very slow. When a quantity varies very slowly, one can obtain a good approximation in most calculations by considering the quantity fixed and giving it an average value. Therefore, I feel that ONETWORES gives valid results.

Because the calculated reactances are proportional to Z0, it is possible to determine Z0 experimentally by adjusting Z0 in such a way that the experimental values agree with the calculated ones, on the average. The comparison of calculated and experimental values is easiest to do with a center-loaded dipole an odd number of quarter wavelengths long. In this case A = 0 and Tan (B) is ±1 with the result that the reactance on each side of the feed is Z0 for antennas 1,5,9....quarter wavelengths long and - Z0 for antennas 3,7,11.....quarter wavelengths long.

For an earlier article,⁷ I built 5/4 wavelength dipoles for 21, 28, and 50 MHz and showed how to match their impedance to 50 ohms coax. (These antennas, when fed with open wire line, are popularly called extended double Zepps.) In the course of that work, I measured the reactances necessary, and, after correcting for distributed capacitance, concluded that the total reactance needed is 990 ±170 ohms, where the error is one standard deviation. This is the total reactance, or Z0 = 495 ±85 ohms, which is in good agreement with the theoretical values.

It's possible to extract a value of Z0 from every number in the table contained in Michael's **Figure 2**. I selected the two values easiest to calculate as samples. In accordance with the previous discussion, a quarter-wave center-loaded dipole (his 1/2, 0 case) needs a reactance of Z0 on each side for resonance. From the inductance of 11.5 µH, I calculated Z0 = 517 ohms. Furthermore, if the loading reactors are moved out to the middle position (his 1/2, 0.5 case) the reactance must be 2*Z0. From the inductance of 21 µH, I calculated Z0 = 472 ohms.

I've concluded that 500 ohms is a convenient round number which represents these values, and that 20 percent is a valid estimate for the absolute errors in the calculations. However, the precision in comparing different calculated values is much greater.

End corrections. I mentioned that a dipole physically one-half wave long has a reactance. To obtain resonance, it's necessary to reduce the length by about 5 percent for a thin wire antenna (more for a thick one). A physical half wave length:

$$HW = 492/F \quad (7)$$

while an electrical half wave length:

$$HW = 468/F \quad (8)$$

where HW is in feet and F is in MHz.

I estimate such a correction for antennas of other lengths. I believe the difference is due to an end effect. If electrostatic charges are placed on a insulated wire, they are distributed uniformly everywhere but near the ends, where they are more dense. This gives rise to more lines of force per unit length. Therefore, it's logical to attribute the difference between **Equations 8 and 9** to a correction:

$$DL = 12/F \quad (9)$$

to be subtracted from each end of the electrical length to determine the physical length, or vice versa. One approach is to use this as a correction for other lengths. Another approach is to assume the contraction is uniformly distributed.

I've used an intermediate correction. My conversion of the total length in wavelengths, LW, to physical length in feet, LE (214), is based on **Equation 8**; the conversion for the resonator separation, LL (222) to LT is based on **Equation 7**. As a result, it appears that between the resonators, the electrical and physical lengths are the same, while the outer portions are uniformly contracted to account for the total.

I don't plead a strong defense for making the correction is this way. Indeed, because the size of the correction is smaller than some of the errors, I don't plead a strong defense for making any correction at all. At any rate, the correction doesn't affect any of the calculations, except for explicit conversions between physical and electrical lengths.

Vertical antennas

For the ideal situation of a vertical antenna mounted over perfectly conducting ground, the pattern is the same as one half of the free-space pattern of a free-space dipole twice the length, and radiation resistance is one half as large. On such premises, one can apply ONETWORES to vertical ground-mounted dipoles. However, you must expect poorer agreement between any experiment and the results calculated by the program. While there is some basis to support the assumption that Z0 is independent of length in the case of horizontal antennas, as in practice their situations are intermediate between being close to the ground and being in free space, there is little justification for assuming a constant Z0 for vertical antennas.

The characteristic impedance for a ground-mounted vertical antenna is given by:⁸

$$Z0 = 60(LN(2*H/R) - 1) \quad (10)$$

where H is now the length of the antenna, R is its radius in the same units, and LN is the natural logarithm.

For representative values of $H = 35$ feet and $R = 0.5$ feet, Z_0 is about 390 ohms. For a vertical antenna without a loading reactor at the base, it's logical to use the value from this formula for ZA and a slightly larger value for ZB, but I have no reliable way of estimating ZB.

I haven't made a thorough comparison between the values from the program and the experimental data on vertical antennas. I do have two verticals. Both are base loaded. Qualitatively, they conform with the conclusions of my theoretical calculations; but in the few cases in which I have been able to make comparisons, the calculated reactances don't agree very well with the experimental ones.

Concluding remarks

I've described a program capable of providing useful information about loaded dipoles. Because it's based upon very simplifying assumptions, the program has limited accuracy. The inclusion of two adjustable constants—(1) the normalization factor that makes the radiation resistance of the half-wave dipole conform with a selected value (965) and (2) the characteristic impedance, Z_0 , (180-185)—compensates for the program's lack of accuracy.

ONETWORES has never been a static program. I continually modify and adapt it for special problems. **Listing 1**, as it stands is likely to be most useful to a user who's not interested in making program changes.

I have made adaptations for off-center fed dipoles, continuously loaded antennas, and antennas with capacitive hats.

With one known exception, radiation resistance are not additive. For dipoles short compared to the wavelength, they are proportional to the square of the length. In a private communication, L.A. Moxon, G6XN, called my atten-

tion to the one known exception. It occurs, when two portions of an antenna are at right angles, as happens with an "L" antenna. Those for the horizontal and vertical parts can be calculated independently and added together—assuming that they are referred to the same feed location. On this basis, ONETWORES can yield the reactance and radiation resistance of an "L" antenna.

I have little information as to how calculations pertaining to these adaptations compare with experiments or with the results of other computer programs. For this reason, I leave the details for you to work out, or for subject of correspondence.

It's interesting to note how the current sometimes varies with the distance, Z. I have modified a version of ONETWORES, creating a short program for calculating the current as a function of Z. You can easily modify a listing of **Stage 2** into such a program. First insert these statements:

```
460 PRINT "FREQUENCY" (MHZ)=";F
620 PRINT "Z =" ;Z;"I1=" ;I1;"I2=" ;I2
```

Delete all statements related to the angle X (385-420, 800). To avoid the need to answer irrelevant questions, delete (130-155). You may, if you wish, delete all statements involving E, ET, J, and JT—although their presence doesn't prevent operation of the program. ■

REFERENCES

1. John D. Kraus, *Antennas*, McGraw-Hill Book Company, New York, 2nd Edition, 1988, page 218.
2. Kraus, page 227.
3. L.A. Moxon, G6XN, *HF Antennas for All Locations*, Radio Society of Great Britain, London, 1st Edition, 1982, Figure 4.40, page 61.
4. Charles J. Michaels, W7XC, "How Short Can You Make a Loaded Antenna?," *Communications Quarterly*, Summer 1992, page 73.
5. Kraus, Chapters 8 and 9.
6. S.A. Schelkunoff and H.T. Friis, *Antennas Theory and Practice*, John Wiley and Sons, New York, 1952, Chapter 13.
7. Yardley Beers, W0JF, "The 5/4-Wavelength Dipole: A Revival," *Communications Quarterly*, Fall 1990, page 40.
8. *The ARRL Antenna Book*, Equation 2, page 2-38.

PRODUCT INFORMATION

270Vdc, 10 Amp Power Switching Relay

Kilovac Corporation introduces a new 270Vdc, 10A multi-purpose power switching relay for the Military, Space and Commercial Aerospace industries. Kilovac's new AP10A relay is rated from -55°C to 85°C, and can be used for 270Vdc applications as well as 155Vdc for submarines and ships, 120Vdc battery operated systems like the space station, and a variety of other applications.

The AP10A is 1.7 cubic inches in size and can provide over 2000 volts of isolation across

open contacts for system safety. It can switch at up to 1000Vdc at lower currents without damage, and dissipates less than 3 watts at 10A, eliminating the need for heat sinks and fans or de-rating for temperature.

The AP10A also offers mounting options including through the chassis, PC board, and panel mounting.

For more information contact Kilovac Corporation, P.O. Box 4422, Santa Barbara, CA 93140 or call 805-684-4560.

GILBERT CELL ACTIVE MIXERS

A quick look at an interesting device

The main disadvantages of diode mixers have always been the amount of dynamic range available and intermodulation problems caused by strong signals. Various active mixer designs have been used for many years in an attempt to decrease these problems by using active devices in place of diodes. Most designs incorporating discrete devices relied on FETS because of their predictable bias and temperature characteristics. When integrated circuit technology became available, numerous bipolar transistors could be laid out on a single substrate to increase stable operation with external bias components. The Motorola MC14/1596 series of balanced modulators/demodulators is an example of this technique. A close look at the schematic of this device shows that it has the characteristics of a Gilbert cell mixer.

Reviewing the fundamentals

In order to understand the operation of Gilbert cell active mixers, it's necessary to review the fundamentals of the double-balanced mixer configuration consisting of a quad layout of Schottky barrier diodes and two balun transformers. Schottky diodes ensure quick turn-on when the threshold voltage level is reached and when placed in a matched configuration, as shown in **Figure 1**. The high level of the LO injection signal attempts to maintain a square-wave effect in the diodes, providing a nonlinear mixing (multiplication as opposed to addition) process with the low level RF signal.

The waveform shown at the LO input is actually the signal that results after passage through the diode quad; the positive part of the waveform produces a +1 value and the negative portion a -1 value. This action is the outcome of the square wave alternately causing opposite sides of the quad to conduct. The product (or IF output), then, is multiplication of the RF signal

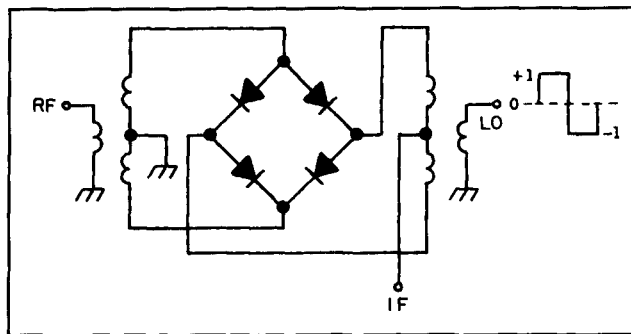


Figure 1. Double-balanced diode mixer configuration.

by ± 1 at the LO frequency. This can be shown algebraically as:

When the LO is >0

$$\begin{aligned} V_{out} &= +V_{lo} - V_{rf} \\ V_{out} &= -V_{lo} - V_{rf} \\ V_{out} &= -V_{rf} \end{aligned}$$

When the LO is <0

$$\begin{aligned} V_{out} &= -V_{lo} + V_{rf} \\ V_{out} &= +V_{lo} + V_{rf} \\ V_{out} &= V_{rf} \end{aligned}$$

The resulting multiplication in a real double-balanced diode mixer produces a 6 to 8 dB conversion loss and 20 to 40 dB suppression of the RF and LO signals. Higher harmonics of the LO also create high order frequency terms at the IF output at a level consistent with a Sin x/x distribution. Fast switching action of the diode quad is probably the most important parameter of the diode mixer.

In the Gilbert cell active mixer of **Figure 2**, the configuration consists of an emitter-coupled transistor pair Q1, Q2, and four collector-cross-coupled transistors Q3-Q6. The RF input is coupled to Q1 while Q2 is AC grounded. As

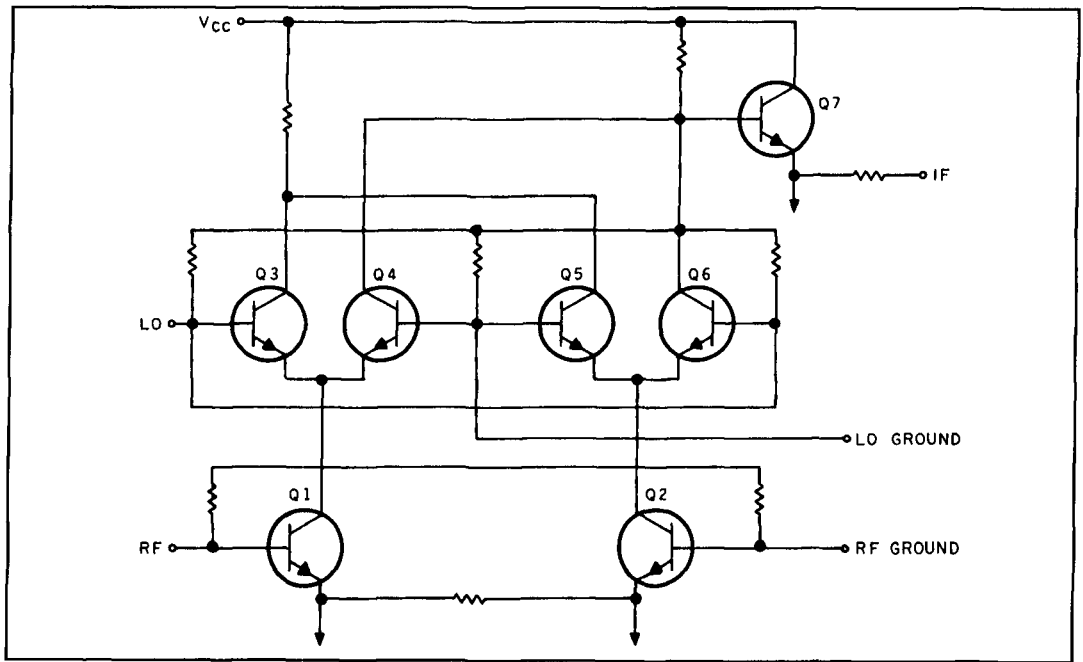


Figure 2. Gilbert cell active mixer configuration.

the RF signal proceeds from Q1 through Q2 via the emitters and from base to base, a differential mode due to the 180-degree phase shift changes the current between the two branches, creating a wideband input at low VSWR.

However, this configuration also loads the output of Q2. By using Q7 in an emitter follower scheme, the output port loading is reduced and the output impedance is set by a resistor.

To multiply the RF input by ± 1 at the LO frequency and ensure the required switching action, four cross-coupled transistors Q3-Q6 are used. The LO is injected at Q3 to Q6 with the opposite side Q4 and Q5 AC grounded. Positive values of the LO square wave turn on Q3 and Q6 and multiply the RF by ± 1 at the LO frequency while negative values activate Q4 and Q5, multiplying the RF signal by -1 at the LO frequency. Not shown on the diagram are the various voltage and current sources that ensure wideband matching and load insensitive operation—important parameters that are difficult to satisfy with other mixer designs.

The emitter-coupled amplifier configuration of the Gilbert cell mixer is responsible for the available conversion gain. By using incorporated AC grounding capacitors, the low end frequency response is set to 50 MHz. Pin connections provide for the addition of external capacitors to lower this frequency. Although diode mixers may appear to provide better RF/IF isolation, the higher required LO injection level and conversion loss tend to outweigh any possible benefit. Additionally, the high power levels at opposing ports in the diode mixer make it prone to producing undesired intermodulation products.

One disadvantage

Although the Gilbert cell active mixer has many advantages over a diode design, such as gain, low oscillator injection level, and load insensitivity, it also has one main disadvantage—high noise figure. It is not a low-noise device, due mainly to the shot noise of the four collector-cross-coupled transistors and the emitter-coupled pair, as well as thermal noise introduced by the feedback resistor.

Additionally, attempts to operate this device at low LO frequencies will increase the noise figure due to the mixer quad being unable to conduct during portions of each cycle. This anomaly can be lessened somewhat by using a square wave to drive the LO port.

Common uses for Gilbert cell mixers

The most common use for the Gilbert cell active mixer IC is in UHF and low microwave up and downconverters. Although the noise figure precludes using this device at the antenna input, a low-noise amplifier with 20dB gain preceding it will provide comparable performance to a passive mixer/IF amplifier combination. The RF and LO ports will accept signals up to 5 GHz with IF gain to 2 GHz.

As with passive mixers, the Gilbert cell configuration can also be used to design harmonic mixers, frequency dividers, multipliers, and phase detectors.

BIBLIOGRAPHY

1. Avantek Application Notes AN-S013, AN-S010, data sheets.

A Ready-Made MMIC Active Mixer/Amplifier Combination

The AvanteK IAM-81008 (see **Figure 3**) is a silicon bipolar monolithic microwave integrated circuit (MMIC) active double-balanced mixer with integral IF amplifier. Supplied in an SO-8 surface mount package, this device makes a small 0.2 x 0.24 square inch footprint. Using the Gilbert cell active mixer, conversion gain is available at frequencies from 50 to 5000 MHz at the RF input and to 1 GHz at the IF port.

According to the specifications, the gain varies from about 8 dB at an IF of 40 MHz to 9 dB at 100 to 200 MHz and down to 0 dB at about 1.8 GHz. If the LO frequency is lower than the RF, there is 2 dB less gain from about 400 MHz to the high end of 1 GHz at the IF. An additional advantage is that the maximum conversion gain is obtained with -5 dBm LO level. At -15 dBm, the gain drops to 5 dB.

With a noise figure of 17 dB, this is not a front-end mixer, nor is the third order intercept point of 3 dB especially attractive for HF work. It is probably best suited to UHF and low microwave downconverters. When I tested the device on double-sided G10 circuit board with microstrip techniques and surface mount components, an RF input at 900 MHz and -84 dBm, with an LO input of 871 MHz at -5 dBm produced a 29 MHz output at -73 dBm, providing a conversion gain of 11 dB. To determine the low level sensitivity of the mixer, I adjusted the RF input level until the IF output was 3 dB above the noise level with a spectrum analyzer bandwidth of 10 kHz. In this test, I used a 50 ohm splitter to couple the input RF; the other splitter input was terminated in 50 ohms. After I measured the IF output, I placed the 50 ohm termination on the IF port of the mixer and connected the other splitter input port to the spectrum analyzer to measure the LO leakage to the mixer RF port. **Figure 4** shows the various signal levels at the 50 ohm RF and IF ports. For this test, the RF frequency was 900 MHz with an LO of 871 MHz at -5 dBm and the amplifier bias was set at 18 mA with a 100 ohm resistor. Capacitors Cd were 0.001 μ F and capacitors Cb were 0.01 μ F.

MMIC amplifiers

The MSA-0611 and MSA-2111 silicon bipolar MMIC amplifiers are similar cascaded 50 ohm gain blocks with response curves differentiated only by the available circuit current. While the MSA-0611 provides 18 dB gain at 500 MHz, the MSA-2111 is optimized for 16.5 dB gain at 900 MHz. When using the MSA-2111 amplifier ahead of the mixer, the minimum input signal to produce an IF output 3 dB above the noise level was reduced to -100

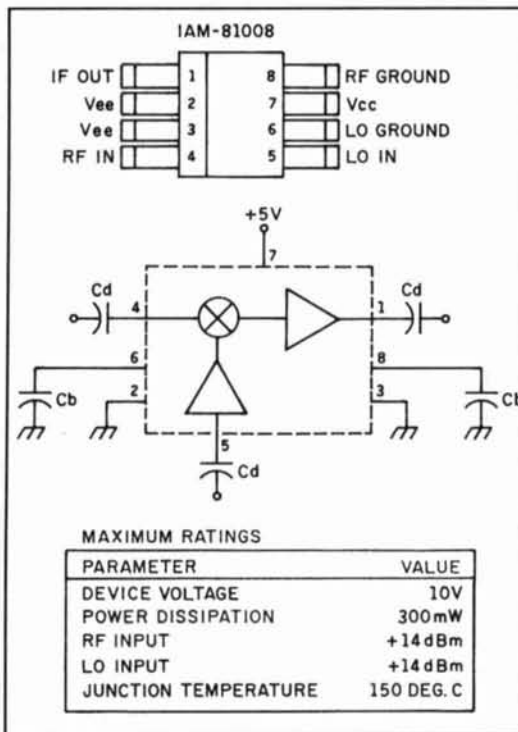
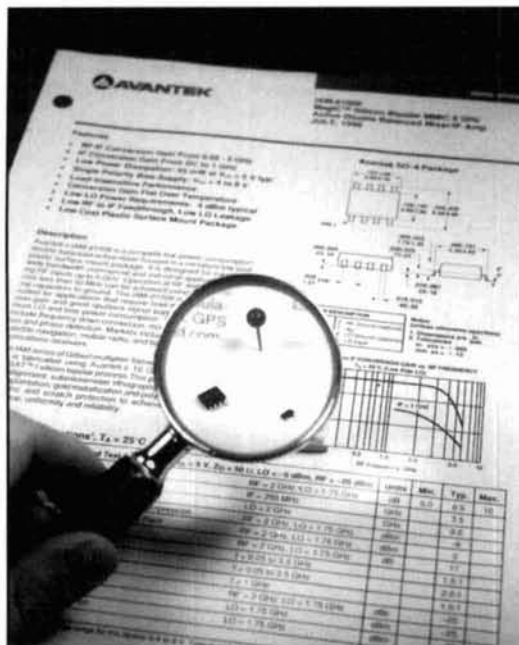


Figure 3. AvanteK Gilbert cell active mixer data.

dBm. For an RF input of -73 dBm, the IF output was -49 dBm. Since the specified noise figure of the MSA-2111 is 3.3 dB at 900 MHz, the cascaded noise figure for this configuration should be about 4.3 dB. By limiting the current to 16 mA, a 3 dB bandwidth of about 700 MHz is possible.

As shown in **Figure 5**, these MMIC ampli-

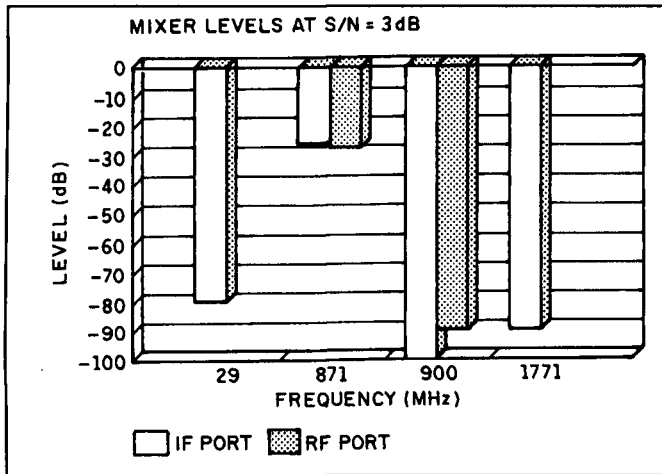


Figure 4. Active mixer test signal levels.

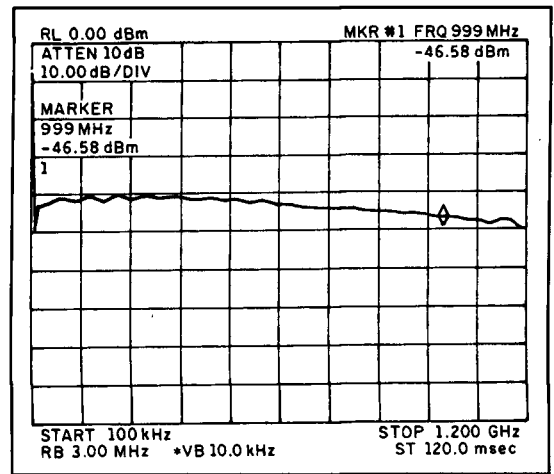


Figure 6. MMIC amplifier swept frequency response.

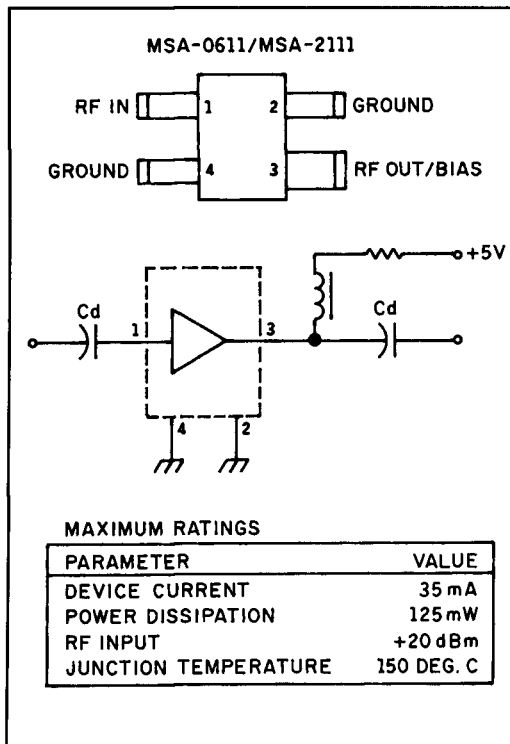


Figure 5. Avantek MMIC amplifier data.

fiers are supplied in SOT-143 surface-mount packages with a common pin for RF output and bias. Since the input/output impedance is nominally 50 ohms, the amplifier can be cascaded. The swept frequency response of the MSA-2111 from DC to 1 GHz, at 5 volts and 18 mA and without input/output matching networks, is shown in **Figure 6**. For an RF input of 900 MHz at -60 dBm, the output was -45 dBm, showing a gain of 15 dBm. In many instances, a wideband "matched" MMIC amplifier will exhibit effects outside the frequency range of interest. As an example, if we wanted to ampli-

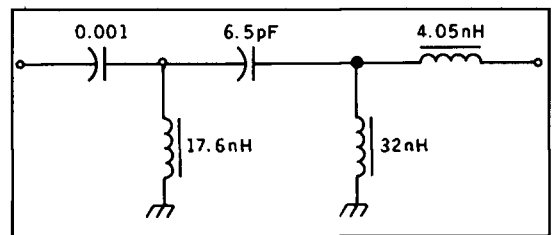


Figure 7. Matching network for cascading MMICs.

fy a narrowband FM signal at 440 MHz, any other signals—either from outside or created by the amplifier—would affect the level and possibly the response of the desired signal. This would be true even if the undesired effects appeared at 10 MHz or some other frequency far outside the range of interest. To check for such an effect with the MSA-2111, I injected an FM signal at 440 MHz and a level of -80 dBm into the same configuration used for the swept frequency test. Monitoring the amplifier output on a spectrum analyzer in 100 MHz steps (10 MHz/division) from DC to 1 GHz showed no extraneous products in this frequency range.

Wideband impedance matching

Because it's very difficult to make a wideband amplifier maintain a purely resistive impedance across a frequency range of many decades, it is important to verify the actual impedances in the frequency range of interest. Avantek supplies S-parameter information on their data sheets, so this is relatively easy to accomplish. Between 100 MHz and 1 GHz, at the specified bias level of 3.3 volts at pin 3 and a current of 16 mA, S11, the input reflection coefficient varies from 0.04 to 0.13 in magnitude and from a phase angle of -176 to -139 degrees. The output reflection coefficient, S22,

varies from 0.04 to 0.18 in magnitude and from -57 to -160 degrees in phase angle.

Transforming these values to input and output impedance provides the data for **Table 1**.

Above 1 GHz, the phase angle quickly changes to a positive value, indicating that inductive effects have overcome the capacitive influence at the lower frequencies. At a bias current of 29 mA, the phase angle of both S11 and S22 are inductive to 1.5 GHz. To cascade two devices over a frequency range of 500 MHz to 1 GHz, the computer-simulated matching network of **Figure 7** would be required for optimum response in the lower current configu-

MSA-0611/MSA-2111 Series Z		
Input	Output	Frequency
46.1-0.25j	52.1-3.5j	100 MHz
46.9-3.75j	42.6-9.54j	500 MHz
40.5-7.0j	35.3-4.5j	1 GHz

Table 1. Amplifier impedance at 5 volts, 16 mA.

ration. Data sheets for these devices can be obtained from AvanteK by calling 1-800-AvanteK. Additional information and applications can be found in Application Notes AN-S013 and ANS010. ■

PRODUCT INFORMATION

255 Spectrum Probe.®

Smith Design introduces their new 255 Spectrum Probe. The Probe is used primarily for EMI investigation, and complements the 107 Spectrum Probe. It also measures RF voltage and RF current entering or exiting electronic equipment. The Probe reads voltage or current on the power line, estimates the field strength on emissions so they can be reduced, sees RF currents and voltages on their PC board, searches for EMI sources that may be interfering with board RF "signatures" equipment and checks the HF portion of infrared, ultrasonic, and line conducted remote controls.



The frequency range for the Probe is less than 30 kHz up to 2.5 MHz, display dynamic range is 60 dB min.; tangential sensitivity, 40 μ V \pm 3 dB at 1 MHz flatness \pm 2 dB .05-2.5 MHz; intermodulation, 3rd. order: -50 dB at 40 mV; vertical output, 5 mV/dB \pm 10%; vertical logarithmic linearity \pm 3 dB; spurious responses, about -50 dB; IF bandwidth, 10 kHz at -6 dB; damage input 1 KVDC, linear reduction to 100v at 500 kHz; sweep rate, 20

ms/2.5 MHz \pm 10%; horizontal linearity, \pm 2%; tip conducted radiation, nil; power 117v \pm 10%, 60Hz, 4W max UL 1319 listed. The Probe weighs 2 oz (18 oz. when shipped) and is 7.5" long, 0.9" diam.; the wall transformer and cable weigh 8 oz. and the transformer is 1.8" wide, 1.8" long, and 1.6" project.

To receive more information on Smith Design's Spectrum Probe, write to: Smith Design, 207 E. Prospect, N. Wales, PA 19454 or call (215) 661-9107.

New! WWV-Format Universal Time Piece

The new Spectra-Com Universal Time Piece (copyright 1993 by JZO Research) combines WWV time and frequency services information from the National Institute of Science and Technology with a real time analog clock.



It shows WWV events and local or universal time with precision in a twelve and twenty-four hour format.

The information is displayed with quartz accuracy (\pm 1 second). It is designed in red, white, blue, yellow and black. The unit uses one AA battery (not included). Use it to pin down such information as storm alerts, GPS and Omega navigation reports, solar activity and more.

For more information contact JZO Research, 7140 Colorado Avenue North, Minneapolis, MN 55429.

Anne Prather, KA9EHV
2776 S. Monroe Street
Denver, Colorado

THE FINAL TRANSMISSION

A new place for an old art

Morse code is a dying art—or so some people would have us believe. We passed a codeless license because we felt that code was a barrier to the younger people trying to get into the hobby. The FCC now allows certain disabled people to waive their code requirement on an individual case-by-case basis. Code is difficult, it's slow, it's boring, and there are much better communications methods out there today.

Statements like this may cause those of us who love the code to rally to its defense—declaring that we will never use sideband, that CW is the only way to enjoy ham radio. Under the onslaught, some of us may even be tempted to treat the new Communicators rudely because their codeless licenses fly in the face of our long-held preferences.

Such behavior is not helpful. The communications art is changing. As it changes, so will the use of the International Morse Code. Its usefulness in radio may well be fading, though you wouldn't know it listening to the CW bands. But there are other places in which Morse code may prove a useful and effective communications alternative. Those of us who want to preserve the code would do well to encourage its implementation in appropriate applications.

The ham radio subculture contains many tales of situations in which the code played a vital role. My personal favorite came from a young friend in college. When asked why she was late for our dinner appointment, she explained: "Well, the elevator got stuck, and the alarm sounds just like the telephone. I couldn't get anybody to come until I tapped out SOS on the alarm." Another friend tells of a blind-deaf amateur who used CW by sending with a key and putting his hand on the speaker cone to copy incoming signals.

Yes, there are better, more sophisticated communications methods out there. Packet,

spread-spectrum, digital signal processing—all of these contribute to the communications art we have pursued over the years. But it seems to me that if we allow ourselves to be deluded into letting the Morse code die, we are losing the ability hams have had since the inception of the hobby to provide ingenious and economical solutions to difficult problems. Further, we will lose a communications method that has three important qualities. First, it is simple. Second, because it is simple, its implementation is economical. And finally, it is universal.

SOS means "distress" whether you're in Hong Kong or Canada. An amateur from Uzbekistan knows that "QTH" means "location." Turks, Arabs, Americans, the blind, the blind-deaf, the quadriplegic all use the same International Morse Code. This is not as true of some of the other communication methods out there. Yes, digital standards are international, but the data communicated by these media are not, necessarily. Yes, the average DX station will know enough English to say "Your signal is S5"—but look where the signal-reading convention came from!

The universality of the International Morse Code may decline along with its use in commercial applications. But as long as there is need for a simple, bare-bones communications method that spans both cultural and physical parameters, it's well worth preserving the code.

I can send 15 words a minute with my lips, ten words a minute with my eyebrows, and at least that with most of my toes. Fortunately, except for the experimentation I did for this article, there's no real reason for me to improve these proficiencies. I have a friend with Lou Gehrig's disease, however, who needs to be able to send with something besides his hands. He is losing his voice, so the standard voice-recognition software typically used by high-level quadriplegics won't work. If he were to lose his eyesight as well, he would be unable to

use the point-and-click software he currently uses. How nice the Morse code would be in this situation. It's not fast—but neither are his current methods of computer interface.

Recently I found myself in need of something to help out with computer screens. Although I can see well enough to read most words, I have a problem if I have to look at individual letters. Further, I travel a lot, which makes portability a prime consideration in any adaptive method I use. I recently picked up a computer program that echoes ASCII characters in Morse code. It will go as high as 100 wpm and uses different pitches to indicate upper and lower-case characters. It took some getting used to, but the ability to distinguish case proved to be wonderful for checking capitalization and punctuation errors.

For those who need sophistication and convenience, IBM makes a whole series of voice-recognition and speech-synthesis products designed to work with PCs. They did not send me price information on any of these products, but I suspect they run into the thousands of dollars. They also require large amounts of RAM and disk space—limiting the amount of space available for other applications. They are faster and easier to use than Morse interfaces. But for those who know the Morse code, it may be just as effective in “spell-heavy” applications—like spell checkers, DOS hex editors, etc.—to use Morse code.

If I were completely blind, I would certainly investigate a speech synthesizer for use at home and for reading large volumes of material. Even at 80 wpm, a Morse code interface can't go as fast as a speech synthesizer. But it's nice to know that if I want to work on a writing project on a ski trip, I have an audible, self-contained method of reading the screen.

A blind friend recently spent \$900 for her speech synthesizer. A portable version of the one she bought is available for \$1295. The portable version is lightweight—13 oz.—and plugs into the serial port of a laptop or notebook. By contrast, I paid nothing for the Morse reader—at most I would be out a \$50 shareware fee for its use. Even though it's light, you still have to haul an extra piece of hardware around if you use a portable speech synthesizer. For my particular application and circumstance, the Morse reader turns out to be an economical and effective solution to an awkward problem.

When all is said and done, Morse code can be implemented with the simplest materials—a switch and an oscillator or beeper to turn on and off. Because it's so simple, it can be implemented inexpensively in many cases. And its use crosses both national and physical barriers. Blind-deaf hams can communicate with sighted hams by using the code. Quadriplegics can use the code to activate voice synthesizers if there is no other activation method that will work.

People severely handicapped enough to

require the use of Morse code are probably relatively few in number. But for those who really need it, the code can open up the world. And it is the world-opening ability of Morse code for those who need it that we will lose if we lose the code. When you operate CW, you can't tell if the person on the end is “normal” or severely handicapped with multiple sclerosis, for example. And it's a great relief to someone who is handicapped in any way to be able to interact in a societal circumstance in which the handicap is not central. As a matter of fact, in most CW QSOs, the handicap is irrelevant.

The code isn't ideal. It has many drawbacks—especially for those whose minds don't deal with languages easily or who, for whatever reason, find the code difficult. But those of us who love the code, if we want to preserve it, ought to be out there looking for other places besides radio in which it can be effectively implemented and used. If we want to keep the code a viable, dynamic communications method, we need to look honestly at its advantages and disadvantages, and encourage its use in places where its advantages are integral to the final application.

There is another aspect of the code that my discussion of its utilitarian uses has ignored. The code is a living language, complete with the idiom and nuance that make each language unique. I can infer a lot about a person by the way he or she sends code. Do they use IMI as a question mark or an indication of repetition? If weighting is selectable on their keyer or keyboard, what have they chosen? Do they send “fer” to mean “for?” How do they laugh? Each person seems to develop a set of code-sending habits that come through whether they use a keyboard or a bug.

Viewing the code as a language takes it beyond the dimension of something boring and utilitarian and brings it into the realm of codes and intrigue. It seems to me that CW fans could do a lot to further the art of Morse code by emphasizing the intrigue. If you love the code and want your kids to love it, help them build a wire telegraph between their house and a neighbor's. Play games with the code. Tell stories about it. See if there is someone in your life for whom the code can solve a perplexing communication problem.

It is perfectly possible to enjoy the code for what it is and what it can do. I enjoy the code. I also enjoy computers and SSB. I have used the code to deal with my difficulties using computers in a way suitable to me. I experience a deep sense of frustration at the attitude that CW is the be-all and end-all of ham radio. This flies in the face of reality as does the attitude that the code is obsolete and ought to be discontinued. Neither attitude is completely true, and the best thing those of us who love CW can do is admit that the code has its limitations and work to find it a place in modern communications art. ■

LETTERS (from page 6)

Ready to build

In the winter issue of *Communications Quarterly*, in an article about homebrewing, you mentioned radio kits, but I have never seen them advertised. Would you be so kind as to give me a list of companies dealing in radio kits?

The first rig I used was a Hammarlund 6L6 oscillator back in 1933, and enjoyed it very much!

Sidney Jump, KA1OJW
Lowell, Massachusetts

On the lookout

I have just received your winter 1993 issue. I enjoy your magazine very much.

I refer to the Editor's note on page 21, where you say "...I suggest using a slug-

tuned ceramic form..." I have been looking for a source of such forms and/or cores for quite some time, and have not found any, either manufacturer's or surplus. In fact, most salespersons these days haven't any idea what I am talking about.

In the course of my work, I am occasionally called on to design and build special-purpose filters, and when my junkbox runs out, I am going to be out of this business.

If you can suggest anyone who still makes or stocks coil forms and cores, I would be most grateful.

James Lomasney, WA1NIL
Menlo Park, California

*That's a toughie; any suggestions readers?
Ed.*

The Best Shortwave You Can Buy Comes From Drake



The Drake R8 Communications Receiver...simply the best shortwave clarity and fidelity you'll find, outperforming receivers costing much, much more.

Famous Drake technology gives you wide frequency coverage of all world bands and excellent dynamic range...in an uncluttered package with an ergonomic front panel, featuring keypad entry of functions.

For the best access to world events as they happen, buy yourself a Drake R8 shortwave receiver. Ask your dealer for more information, or contact a Drake sales office today at **1-800-9-DRAKE-0 (1-800-937-2530)**.

DRAKE

We Bring the World Closer

R. L. Drake Company, P.O. Box 3006, Miamisburg, OH 45343, USA Tel: 513-866-2421
Drake Canada, 655 The Queensway #16, Peterborough, Ontario K9J 7M1, Canada
Tel: 705-742-3122

1-800-9-DRAKE-0 (1-800-937-2530)

ASTRON POWER SUPPLIES

• HEAVY DUTY • HIGH QUALITY • RUGGED • RELIABLE •



MODEL VS-50M

SPECIAL FEATURES

- SOLID STATE ELECTRONICALLY REGULATED
- FOLD-BACK CURRENT LIMITING Protects Power Supply from excessive current & continuous shorted output
- CROWBAR OVER VOLTAGE PROTECTION on all Models except RS-3A, RS-4A, RS-5A, RS-4L, RS-5L
- MAINTAIN REGULATION & LOW RIPPLE at low line input Voltage
- HEAVY DUTY HEAT SINK • CHASSIS MOUNT FUSE
- THREE CONDUCTOR POWER CORD except for RS-3A
- ONE YEAR WARRANTY • MADE IN U.S.A.

PERFORMANCE SPECIFICATIONS

- INPUT VOLTAGE: 105-125 VAC
- OUTPUT VOLTAGE: 13.8 VDC \pm 0.05 volts (Internally Adjustable: 11-15 VDC)
- RIPPLE Less than 5mv peak to peak (full load & low line)
- All units available in 220 VAC input voltage (except for SL-11A)

SL SERIES



• LOW PROFILE POWER SUPPLY

MODEL	Colors		Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
	Gray	Black				
SL-11A	•	•	7	11	2 1/4 x 7 1/2 x 9 3/4	12
SL-11R	•	•	7	11	2 1/4 x 7 x 9 3/4	12
SL-11S	•	•	7	11	2 1/4 x 7 1/2 x 9 3/4	12
SL-11R-RA	•	•	7	11	4 1/4 x 7 x 9 3/4	13

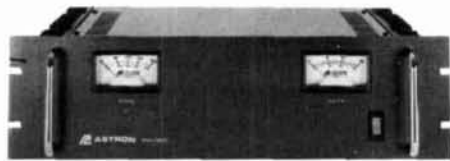
RS-L SERIES



• POWER SUPPLIES WITH BUILT IN CIGARETTE LIGHTER RECEPTACLE

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
RS-4L	3	4	3 1/2 x 6 1/2 x 7 1/4	6
RS-5L	4	5	3 1/2 x 6 1/2 x 7 1/4	7

RM SERIES



MODEL RM-35M

• 19" RACK MOUNT POWER SUPPLIES

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
RM-12A	9	12	5 1/4 x 19 x 8 1/4	16
RM-35A	25	35	5 1/4 x 19 x 12 1/2	38
RM-50A	37	50	5 1/4 x 19 x 12 1/2	50
RM-60A	50	55	7 x 19 x 12 1/2	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5 1/4 x 19 x 8 1/4	16
RM-35M	25	35	5 1/4 x 19 x 12 1/2	38
RM-50M	37	50	5 1/4 x 19 x 12 1/2	50
RM-60M	50	55	7 x 19 x 12 1/2	60

RS-A SERIES



MODEL RS-7A

MODEL	Colors		Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
	Gray	Black				
RS-3A	•	•	2.5	3	3 x 4 1/4 x 5 1/4	4
RS-4A	•	•	3	4	3 1/4 x 6 1/2 x 9	5
RS-5A	•	•	4	5	3 1/2 x 6 1/2 x 7 1/4	7
RS-7A	•	•	5	7	3 3/4 x 6 1/2 x 9	9
RS-7B	•	•	5	7	4 x 7 1/2 x 10 3/4	10
RS-10A	•	•	7.5	10	4 x 7 1/2 x 10 3/4	11
RS-12A	•	•	9	12	4 1/2 x 8 x 9	13
RS-12B	•	•	9	12	4 x 7 1/2 x 10 3/4	13
RS-20A	•	•	16	20	5 x 9 x 10 1/2	18
RS-35A	•	•	25	35	5 x 11 x 11	27
RS-50A	•	•	37	50	6 x 13 3/4 x 11	46
RS-70A	•	•	57	70	6 x 13 3/4 x 12 1/2	48

RS-M SERIES



MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4 1/2 x 8 x 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 x 9 x 10 1/2	18
RS-35M	25	35	5 x 11 x 11	27
RS-50M	37	50	6 x 13 3/4 x 11	46
RS-70M	57	70	6 x 13 3/4 x 12 1/2	48

VS-M AND VRM-M SERIES



MODEL VS-35M

• Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps)			ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
	@13.8VDC	@10VDC	@5VDC	@13.8V		
VS-12M	9	5	2	12	4 1/2 x 8 x 9	13
VS-20M	16	9	4	20	5 x 9 x 10 1/2	20
VS-35M	25	15	7	35	5 x 11 x 11	29
VS-50M	37	22	10	50	6 x 13 3/4 x 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5 1/4 x 19 x 12 1/2	38
VRM-50M	37	22	10	50	5 1/4 x 19 x 12 1/2	50

RS-S SERIES



MODEL RS-12S

• Built in speaker

MODEL	Colors		Continuous Duty (Amps)	ICS* Amps	Size (IN) H x W x D	Shipping Wt. (lbs.)
	Gray	Black				
RS-7S	•	•	5	7	4 x 7 1/2 x 10 3/4	10
RS-10S	•	•	7.5	10	4 x 7 1/2 x 10 3/4	12
RS-12S	•	•	9	12	4 1/2 x 8 x 9	13
RS-20S	•	•	16	20	5 x 9 x 10 1/2	18
SL-11S	•	•	7	11	2 1/4 x 7 1/2 x 9 3/4	12

W2FMI BALUNS AND UNUNS

How to put MORE power into your antenna and get better and higher signal strength ???

Use the all new MOST OUTSTANDING, HIGHEST EFFICIENCY (97% TO 99%), and the BEST BROADBAND UNUN (unbalanced to unbalanced) and BALUN (balanced to unbalanced) TRANSFORMERS exclusively designed and optimized by DR. JERRY SEVICK (W2FMI). Amidon Associates, Inc. now offers the widest selection of the W2FMI Baluns and Ununs to meet all your requirements for better and more consistent communication. Also available in do-it-yourself kits.

Use UNUN transformers for matching 50-ohm cable to:

- (1) Vertical antennas, inverted L's and ground fed slopers (all over good ground systems),
- (2) 75-OHM cable,
- (3) A junction of two 50-OHM cables,
- (4) Shunt-Fed towers performing as vertical antennas, and
- (5) The output of a transceiver or class B linear amplifier when an unfavorable VSWR condition exists.

Use BALUN transformers for matching:

- (1) Half-waves dipoles and inverted V's
- (2) Quad antennas
- (3) Folded dipoles
- (4) Rhombic and V antennas
- (5) Yagi beams



All designs, when used according to instructions, are GUARANTEED to give outstanding performances. Comparisons with other matching transformers are invited.

POWER RATING: at least 2KW continuous from 1MHz to 40 MHz.

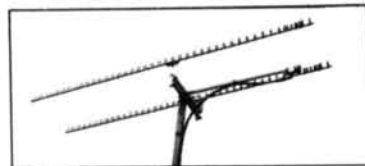
W2FMI HIGH POWER UNUNS	IMPEDANCE MATCH	COMPLETE UNIT PRICE
2.25:1HU50	50:22.22-OHMS	\$45.00
2:1-HU50	50:25-OHMS	\$45.00
2:1-HDU50	50:22.22-OHMS	\$50.00
	50:25-OHMS	
2.25:1-HU112.5	112.5:50-OHMS	\$45.00
2:1-HU100	100:50-OHMS	\$45.00
2:1-HDU100	112.5:50-OHMS	\$50.00
	100:50-OHMS	
1.78:1-HU50	50:28-OHMS	\$45.00
1.5:1-HU50	50:32-OHMS	\$45.00
1.5:1-HU75	75:50-OHMS	\$45.00
4:1-HRU50	50:12.5-OHMS	\$50.00
4:1-HCU50	50:12.5-OHMS	\$50.00
9:1-HRU50	50:5.56-OHMS	\$50.00
9:1-HU50	50:5.56-OHMS	\$45.00
1.78:1-HDU50	50:28-OHMS	\$50.00
	50:12.5-OHMS	
1.56:1-HDU50	50:32-OHMS	\$50.00
	50:18-OHMS	
HMMU50(5 diff. ratios)	MULTIMATCH UNUN	\$80.00
1.78:1-HMMU50	MULTIMATCH UNUN	\$90.00
(8 different ratios)		

W2FMI HIGH POWER BALUNS	IMPEDANCE MATCH	COMPLETE UNIT PRICE
1:1-HBL50	50:50-OHMS	\$55.00
4:1-HBM200	200:50-OHMS	\$55.00
6:1-HB300	300:50-OHMS	\$120.00
9:1-HB450	450:50-OHMS	\$150.00
2.25:1-HB112.5	112.5:50-OHMS	\$120.00
4:1-HB50	50:12.5-OHMS	\$55.00

NOTE: OTHER BALUN & UNUN TRANSFORMERS AVAILABLE. PLEASE CALL OR WRITE FOR ADDITIONAL DETAILS.
BOOKS: 1) "Transmission Line Transformers Design Handbook", by Jerry Sevick, W2FMI, Amidon Associates, Inc., 1991, \$3.00 ea.
2) "Transmission Line Transformers", by Jerry Sevick, W2FMI, APRIL, 1990, \$20.00 ea.

ALSO AVAILABLE:
Iron Powder and Ferrite Products
Millions of Parts in Stock for Immediate Delivery.
Low Cost Experimental Kits.
Call for our FREE tech. flyer.
Toroidal Cores • Shielding Beads • Shielded Coil Forms • Ferrite Rods • Pot Cores • Baluns • Etc.
AMIDON Associates, INC.
Fast reliable service since 1963
2216 East Gladwick Street, Dominguez Hills, CA 90220
TEL: (310) 763-5770 • FAX: (310) 763-2250

DOWN EAST MICROWAVE



MICROWAVE ANTENNAS AND EQUIPMENT

- Loop Yagis • Power Dividers • Dish Feeds
- Complete Antenna Arrays • Linear Amps
- Microwave Transverters & Kits • GaAs FET Preamps and Kits • Microwave Components
- Tropo • EME • Weak Signal • OSCAR • FM
- Packet • ATV
- 902 • 1269 • 1296 • 1691 • 2304 • 2400 • 3456 • 5760
- 10,386 MHz

ANTENNAS

2345 LYK	45 el	Loop Yagi Kit	1296 MHz	\$95
1345 LYK	45 el	Loop Yagi Kit	2304 MHz	\$79
3333 LYK	33 el	Loop Yagi Kit	902 MHz	\$95
1844 LY	44 el	Loop Yagi Assembled	1691 MHz	\$105
3B Fed Tri Band Dish Feed 2.3, 3.4, 5.7 GHz				\$15

Many others and assembled versions available. Shipping extra.

LINEAR AMPS AND PREAMPS

2303 PA	1.2 to 1.3 GHz	3w out	13.8 VDC	\$130
2318 PAM	1.24 to 1.3 GHz	20w out	13.8 VDC	\$205
2335 PA	1.24 to 1.3 GHz	35w out	13.8 VDC	\$325
2340 PA	1.24 to 1.3 GHz	higain 35w out	13.8 VDC	\$355
2370 PA	1.24 to 1.3 GHz	70w out	13.8 VDC	\$695
1302 PA	2.2 to 2.5 GHz	3w out	13.8 VDC	\$430
13 LNA	2.3 to 2.4 GHz	preamp	6 dB NF	\$140
23 LNA	1.2 to 1.3 GHz	preamp	6 dB NF	\$95
33 LNA	900 to 930 MHz	preamp	6 dB NF	\$95
1691 LNA WP	1691 MHz	mast mounted preamp	8 dB NF	\$140

Kits, Weatherproof Versions and other Frequencies Available

NO-TUNE TRANSVERTERS AND TRANSVERTER KITS

SHF 902K	902 MHz Transverter	40mW, 2m IF	Kit \$139
SHF 1296K	1296 MHz Transverter	10mW, 2m IF	Kit \$149
SHF 2304K	2304 MHz Transverter	10mW, 2m IF	Kit \$205
SHF 3456K	3456 MHz Transverter	10mW, 2m IF	Kit \$205

OSCAR and other frequencies available, also Amps and package versions wired and tested. Write for more information. Free catalog available.

DOWN EAST MICROWAVE
Bill Olson, W3HQT
RR1 Box 2310, Troy, ME 04987
(207) 948-3741 Fax: 207-948-5157

COLLINS

Manual Sale: Brand New, Fresh and Crisp! KWM-2 & 2A - \$29 • 51S-1-535 • 30S-1-529 • 32S-3-524 • 30L-1-524 • 75S-3/A - \$29
Plastic Trim Rings: Large fit all S-Line and KWM-2. Small fit 516F-2, B4, B5, etc. \$29.95
Bristol Wrench Set: Complete set of 9 Bristol Spline wrenches for Collins. \$29.95
TR Relay for KWM-2/A, 32S- & 30L-1: Authentic relay for Collins Equipment \$49.95
Panel Meters: New Meters for most all Collins equipment. \$85.95
Paint (12 oz. spray) Primer-Zn Chromate, Galv. Zinc Dry #12 Gray, Cabinet #180, Front Panel #29, 1/2. James Gray #210, RIB-380 Cabinet Lark Brown. (1/2 pt. cont.) RIB-380 Front Panel or Cabinet. \$12 ea. #1/50
Power Cards: 51S-1-535, 75S-515, 516F-2 power-5895, 516F-2 interconnect-5895
Crystals, operating frequencies: HC-6/U holders. \$ each
Crystals, BFO, Calibrator: 456.350, 455.650 & 455.8 KHz-515each. 100 KHz-515each
KWM-2A Plug-in Baluns: 970-2439-010 4POT (90%) or 970-2439-020 4POT (97%) \$3
KWM-2A Plug-in Relay Mod Kit: Converts open frame to plug-in relay. Incl. instructions, relays & Leds. \$124

KWM-380-HF-380 Conversion Kit includes A3 Filter, 060 ROM, and HF-380 Emblem. \$525
380 Reinforced Speaker Grill - \$45; Mic Gain Pot 376-0276-030 - \$52;
MM-280 Mobile Mike - \$45; AC-3802 Speech Processor - \$358; 140 Hz CW Filter - \$195; 360 Hz CW Filter or 1.7 KHz RTTY/SSB/Data Filter - \$109; 6 KHz AM Filter - \$29; Power Xformer - \$45; Front Panel Lense - \$45.

Collins equipment on hand: 51S-1, 32S-1, 32S-3, 75S-1, 75S-2, 75S-3, 75S-3A, 312B-4. Please Call

DOORKNOB CAPACITORS

HT59	500pf	15kv	23amps	\$99
HT57	25pf	15kv	13amps	\$33
HT57	40pf	15kv	13amps	\$29
HT57	50pf	15kv	14amps	\$33
HT57	100pf	15kv	14amps	\$33
HT57	200pf	15kv	15amps	\$30
HT57	500pf	15kv	7amps	\$30
HT57	1000pf	15kv	5.1amps	\$30
HT57	1500pf	15kv	5.3amps	\$33
HT58	10pf	5kv		\$12
HT58	25pf	7.5kv		\$12
HT58	30pf	5kv		\$12
HT58	50pf	5kv		\$12
HT58	66pf	5kv		\$13
HT58	80pf	5kv		\$13
HT58	100pf	5kv		\$12
HT58	120pf	5kv		\$13
HT58	140pf	5kv		\$13
HT58	200pf	5kv		\$12
HT58	300pf	5kv		\$13
HT58	400pf	5kv		\$13
HT58	500pf	5kv		\$12
HT58	1000pf	5kv		\$12

MINIATURE DOORKNOBS WITH WIRE LEADS

HT55	NPO	1pf	5kv	\$11
HT55	NPO	2pf	5kv	\$11
HT55	NPO	5pf	5kv	\$11
HT55	NPO	6pf	5kv	\$11
HT55	NPO	10pf	5kv	\$12
HT55	NPO	15pf	5kv	\$11

Times "IP-8" (RG-8 style COAX)
The best coax you'll ever need! Impervious cable by Times has self-healing Zelex® jacket, is moisture, sunlight, and corrosion resistant and is not affected by cold, heat, condensation or TIME. With IP-8, you have protection from all of nature's elements leading to eventual failure of other RG coax. Electrical specs and dimensions are comparable to other RG-8 cable, with less loss. It's flexible too!
35¢ per foot (plus shipping)

***** NEW ADDRESS *****
1502 Jones St. • Omaha, NE 68102
402-346-4750 • fax: 402-346-2939

Surplus Sales of Nebraska

"EIMAC" 8877 - 3CX1500A7

Last chance to stock up at up to 50% savings. New (original boxed, guaranteed) Gain and linearity curves included with each tube. Limited quantities available. First come first served. Shipping: Add \$7/tube in US. DX please call for proforma.
Individually: \$495 each (2-6) \$450 (7-24) \$425
Matched Pairs \$1099 per pair

SGC

SG-2000	HF Transceiver, 1.8-30 Mhz, 150 watts	\$1599
SG-2000	Extra Control Head (use up to 8 per installation)	\$ 695
SG-230	Smartuner Antenna Coupler	\$ 450
SG-303	Marine/Mobile Antenna (best one for SG-230)	\$ 339
QMS	Quick Mount System (SG-230, SG-303 & QMS)	\$1199

FREE! Just off the Press.... Our newest publication:
15th Anniversary Flyer 6A.

Amplifier parts, Ball Drives, Capacitors (Vacuum, Oil, Electrolytic), Relays (Vacuum, Power), SGC, NYE, DIAMOND, AEA, COLLINS PARTS, Fans, Blowers, Fingerstock, Hardline Connectors, Test Equipment, Meters, Microwave RF GOODIES, Non-Inductive Resistors, Power Supplies, Door Knob Caps!

Send a postcard w/ mailing info to new address below or \$1 for 1st class delivery.

MICROWAVE COMPONENTS
Flexible Waveguide 8.2-12.4GHz; 11-1/2" long; silver inside, resilient rubber outside, 1-5/8" square flange \$75 ea
Mini-Circuits MIXER-TFM-2-408-1-DC-1000Mhz \$8.95 3/\$24

INTERFERENCE REDUCTION FERRITE Split Beads
Beads are split in halves. Clamps around wire/ cable and chokes off unwanted interference from almost any radio transmitter.
Fits up to 1/4" cable, no cage \$3.45/set • 5 sets /\$16
Fits up to 3/8" cable, with cage \$4.95/set • 5 sets /\$23
Fits up to 1/2" cable, no cage \$4.35/set • 5 sets /\$20

HEAVY Tinned Copper Ground Braid
Double Silver Shield - 50Ω
RG-142 \$1.75/foot
RG-178 \$0.75/foot
Subminiature Coax - 50Ω
RG-174 \$0.19/foot
1" Wide \$3/ft. • 50+ feet/\$2.50
1/2" Wide \$1.20/ft. • 50+ feet/\$1

Anniversary Flyer 6A included with your order. Please add adequate shipping. In U.S.-48 figure 304 per lb. to 70 lbs. (1st - 3lbs add \$4 min). Others please call, fax or write for exact shipping total. We accept Visa, MC, Amexco, checks, Cash On Delivery (UPS only, add \$4) or wire transfer.

Surface Mount Chip Component Prototyping Kits—

INDIVIDUAL VALUES AVAILABLE

Only \$49.95



CC-1 Capacitor Kit contains 365 pieces, 5 ea. of every 10% value from 1 pf to .33μf. CR-1 Resistor Kit contains 1540 pieces; 10 ea. of every 5% value from 10Ω to 10 megΩ. Sizes are 0805 and 1206. Each kit is ONLY \$49.95 and available for Immediate One Day Delivery!

Order by toll-free phone, FAX, or mail. We accept VISA, MC, COD, or Pre-paid orders. Company P.O.'s accepted with approved credit. Call for free detailed brochure.

COMMUNICATIONS SPECIALISTS, INC.

426 West Taft Ave.
Orange, CA 92665-4296
Local (714) 998-3021
FAX (714) 974-3420

Entire USA 1-800-854-0547

ANTENNA OPTIMIZERS

AO 6.0 automatically optimizes antenna designs for best gain, pattern, impedance, SWR, and resonance. AO optimizes cubical quads, phased arrays, interlaced Yagis, or any other arrangement of wire or tubing. AO uses an enhanced, corrected MININEC algorithm for improved accuracy, assembly language for high speed, and protected mode for high capacity. AO features stunning 3-D radiation patterns, 3-D geometry and wire-current displays, 2-D polar and rectangular plots with overlays, automatic wire segmentation, automatic frequency sweep, symbolic dimensions, symbolic expressions, skin-effect modeling, current sources, polarization analysis, near-field analysis, up to 450 pulses, and pop-up menus. AO 6.0, \$100. AO-Pro 6.0 (5700 pulses), \$600. 386+387 or 486DX and VGA required. MN 4.5 (256 pulses; any PC; no optimizer, 3-D patterns, or assembly language), \$50. GUY 1.0 (guy-wire modeler), \$25.

YO 5.0 automatically optimizes monoband Yagi designs for maximum forward gain, best pattern, and minimum SWR. YO models stacked Yagis, dual driven elements, tapered elements, mounting brackets, matching networks, skin effect, ground effects, and construction tolerances. YO optimizes Yagis with up to 50 elements from HF to microwave. YO runs hundreds of times faster than MININEC. YO is calibrated to NEC for high accuracy and has been extensively validated against real antennas. YO is intuitive and highly graphical. YO 5.0, \$75. YOC 5.0 (assembly language, much faster), \$100. NEC/Yagis 1.0 (professional accuracy reference), \$50.

YOC and NEC/Yagis require a math coprocessor; MN and YO come with both coprocessor and noncoprocessor versions. All except AO run on any IBM PC compatible with graphics. All programs include extensive documentation. Add 7.25% CA, \$5 overseas. Visa, MasterCard, U.S. check, cash, or money order. 3.5" or 5.25" disk.

Brian Beezley, K6STI
507 1/2 Taylor, Vista, CA 92084
(619) 945-9824, 0700-1800 Pacific Time

CQ Video Library!

Only \$19.95 each

plus \$3.50 shipping and handling per tape.

Getting Started in Ham Radio.

The authoritative fast paced video introduction to the fascinating world of ham radio. Shows how to select equipment and antennas. How to use repeater stations for improved VHF coverage. The importance of grounding and the basics of soldering. How to get the most out of your station, whether it's home-based, mobile or hand held!

Mail Orders To: CQ Communications

76 North Broadway, Hicksville, New York 11801

For Faster Service Call 516-681-2922 or FAX 516-681-2926

ELENCO & HITACHI & B+K PRODUCTS AT DISCOUNT PRICES

TO ORDER CALL TOLL FREE 1-800-292-7711
1-800-445-3201 (Can.)

ELENCO OSCILLOSCOPES



S-1325 25MHz \$349

Dual Trace Oscilloscope

S-1340 40MHz \$495

Dual Trace Oscilloscope

S-1360 60MHz

Dual Trace, Delayed Sweep

\$775

- Automatic beam finder
- Built-in component tester
- 1mV sensitivity
- Dual time base

B-K OSCILLOSCOPES

- 2120 - 20MHz Dual Trace \$395
- 2125 - 20MHz Delayed Sweep \$539
- 1541B - 40MHz Dual Trace \$749
- 2160 - 60MHz Dual Trace, Delayed Sweep, Dual Time Base \$949
- 2190 - 100MHz Three Trace Dual Time Base, Delayed Sweep \$1,395
- 2522 - 20MHz / 10MS/s Storage \$869
- 1442 - 20MHz Portable \$1,229
- 1443 - 40MHz Battery / AC operated with Cursor & Readouts \$1,439

1.0GHz PORTABLE SPECTRUM ANALYZER

Model 2610 \$2,595.95

- AC/DC operation (battery included)
- 70dB dynamic range
- Resolution bandwidth of 10kHz
- 50Ω and 75Ω input impedance (switch selectable)
- Fixed bandwidth setting for viewing TV signals
- Field calibratable with internally generated 100MHz, 80dB signal

Hitachi Compact Series Scopes

- V-212 - 20MHz Dual Trace \$409
- V-525 50MHz, Cursors \$975
- V-523 50MHz, Delayed Sweep \$949
- V-522 50MHz, DC Offset \$849
- V-422 40MHz, DC Offset \$749
- V-222 - 20MHz, DC Offset \$625
- V-660 - 60MHz, Dual Trace \$1,095
- V-665A - 60MHz, DT, w/cursor \$1,325
- V-100A - 100MHz, Dual Trace \$1,375
- V-1065A - 100MHz, DT, w/cursor \$1,649
- V-1085 - 100MHz, DT, w/cursor \$1,995
- V-1100A - 100MHz, Quad Trace \$2,195
- V-1150 - 150MHz, Quad Trace \$2,695

Hitachi RSO Series

- RSO's feature: roll mode, averaging, save memory, smoothing, interpolation, pretriggering, cursor measurements
- VC-6023 - 20MHz, 20MS/s \$1,650
 - VC-6024 - 50MHz, 20MS/s \$1,950
 - VC-6025A - 50MHz, 20MS/s \$2,350
 - VC-6045A - 100MHz, 40MS/s Call
 - VC-6145 - 100MHz, 100MS/s Call

Digital Capacitance Meter

CM-1550B \$58.95
9 Ranges
.1pf-20,000pf
.5% basic acy.
Zero control w/ Case
Big 1" Display

Digital LCR Meter

LC-1801 \$125
Measures:
Coils 1uH-200H
Caps 1pf-200pf
Res. 0.1-20M

Multimeter with Capacitance & Transistor Tester

\$55 CM-1500B
Reads Volts, Ohms
Current, Capacitors,
1 Transistors and
Diodes / with case

FLUKE MULTIMETERS

Scopemeters (All Models Available Call)

Model 93	\$1,095.00	70 Series	
Model 95	\$1,395.00	Model 701	\$65.00
Model 97	\$1,695.00	Model 771	\$145.00
10 Series		Model 791	\$169.00
Model 10	\$62.95	80 Series	
Model 12	\$79.95	Model 87	\$289.00

CALL US FOR ALL YOUR COMPONENT NEEDS

Soldering Station Temperature Controlled SL-30 \$99
Digital Display
Temp Range: 300F-900F
Grounded Tip
Overheat Protect

Video Head Tester HT-200 \$44.95
Tells you if VHS head is defective or worn.

Digital Multimeter w/ Inductance & Capacitance \$75.00
LCM-1850
Ten Functions
by Elenco

Color Convergence Generator SG-250 \$89.95
Kit \$69.95
Finest in the industry 10 rock steady patterns RF & video output

Quad Power Supply XP-580 \$69.95
2-20V @ 2A
12V @ 1A
5V @ 3A
-5V @ .5A
Fully regulated and short circuit protected

Triple Power Supply XP-620 Assembled \$75
Kit \$50
2 to 15V @ 1A,
-2 to -15V @ 1A
(or 4 to 30V @ 1A)
and 5V @ 3A
All the desired features for doing experiments. Features short circuit protection, all supplies

AM/FM Transistor Radio Kit with Training Course Model AM/FM 108 \$27.95
14 Transistors • 5 Diodes
Makes a great school project

True RMS 4 1/2 Digit Multimeter M-700T \$135
55% DC Accuracy
with Freq. Counter
Data Hold

Sweep/Function Generator with Freq. Counter \$259
Elenco Model GF-9026
Sine, Square, Triangle, Pulse, Ramp
2 to 20MHz, Freq. Counter: 1-10MHz
Internal Linear & Logic Sweep

Function Generator Blox #9600 \$28.95
Kit \$26.95
Provides sine, triangle, square wave from 1Hz to 1MHz
AM or FM capability

XX-500 Digital / Analog Trainer
A complete mini-lab for building, testing, prototyping analog and digital circuits
Elenco's Digital/Analog Trainer is specially designed for school projects, with 5 built-in power supplies. Includes a function generator with continuously variable, sine, triangular, square wave forms. All power supplies are regulated and protected against shorts.

- Power Supplies**
- Variable Power Supply
 - +1.25 to 20VDC @ 5 Amp
 - +1 to 15VDC @ 1 Amp
 - +1.25 to 20VDC @ 5 Amp
 - +1.25 to -15VDC @ 1 Amp
 - 12VDC @ 1 Amp
 - 12VDC @ 1 Amp
 - 5VDC @ 1 Amp
 - 30VAC Center tap
 - 15VAC @ 1 Amp
- Analog - Section**
- Function Generator Sine, Triangular, Square wave forms
 - Frequency adjustable in five ranges from 1 to 1000kHz
 - Fine frequency adjust
 - Amplitude adjust
 - DC offset
 - Modulation FM-AM
- Digital - Section**
- Eight data switches
 - Two no bounce logic switches
 - 8 LED readouts TTL buffered
 - Clock frequency is 100kHz
 - Clock amplitude SVPP square wave
- Breadboards**
- 2 breadboards, each contains 80 tie points (total 1,600)



Learn to Build and Program Computers with this Kit Includes: All Parts, Assembly and Lesson Manual Model MM-8000 \$129.00

Elenco Wide Band Signal Generators

SG-9000 \$129
RF Freq 100K-450MHz AM Modulation at 10KHz Variable RF output
SG-9500 w/ Digital Display & 150 MHz built-in Counter \$249

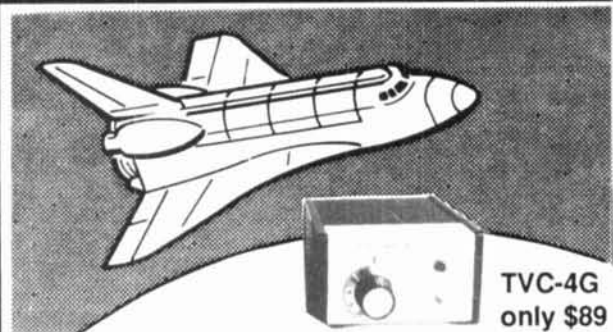
WE WILL NOT BE UNDERSOLD UPS SHIPPING - 48 STATES 97 IL RES 7.5% TAX (33 min \$10 min) PROBES INCL ALL SCOPES & METERS

C&S SALES INC. 1245 ROSEWOOD, DEERFIELD, IL 60015 FAX: 708-520-0085 • (708) 541-0710



15 DAY MONEY BACK GUARANTEE FULL FACTORY WARRANTY WRITE FOR FREE CATALOG

AMATEUR TELEVISION



**TVC-4G
only \$89**

SEE THE SPACE SHUTTLE VIDEO

Many ATV repeaters and individuals are retransmitting Space Shuttle Video & Audio from their TVRO's tuned to Satcom F2-R transponder 13. Others may be retransmitting weather radar during significant storms. If it is being done in your area on 70 CM - check page 413 in the 91-92 ARRL Repeater Directory or call us, ATV repeaters are springing up all over - all you need is one of the TVC-4G ATV 420-450 MHz downconverters, add any TV set to ch 2, 3 or 4 and a 70 CM antenna. We also have downconverters and antennas for the 900 and 1200 MHz amateur bands. In fact we are your one stop for all your ATV needs and info. Hams, call for our complete ATV catalog - antennas, transceivers, amplifiers. We ship most items within 24 hours after you call.

(818) 447-4565 m-f 8am-5:30pm pst.

Visa, MC, COD

P.C. ELECTRONICS

Tom (W6ORG)

2522-CQ Paxson Ln Arcadia CA 91007

Maryann (WB6YSS)

Low Cost RF Software

The RF Design Software Service distributes programs published in RF Design magazine. These aren't commercial software packages, just accurate, working programs by some of the best engineers in the world! Nearly every monthly issue of RF Design since February 1989 has had at least one RF computer program published, for a total of 80 programs — and the collection keeps growing.

The most recent programs include:

RFD-0193 (January 1993)

"The Design of Constant Phase Difference Networks" by Robert Dehoney. Computes poles, zeros, and component values for active and L-C phase shift networks.

RFD-0293 (February 1993)

"A Program for the Design of Lowpass Coaxial Filters" by Eric Stasik. Designs Butterworth and Chebyshev lowpass filters using coaxial line sections as the reactances.

RFD-0393 (March 1993)

"MNAP9 Nodal Network Analysis Program" by John Eisenberg. Full nodal analysis of circuits using resistors, capacitors, inductors, transmission lines, or s-parameter data.

RFD-0393-MAC (Two Macintosh Programs from March 1993)

"Series Feedback Oscillator Design" by Ted Grosch. Maps circles of constant reflection coefficient circles for oscillator design.

"Log Mac: Frequency and Time Domain Analysis" by Jeff Crawford. Circuit transfer function is analyzed by frequency response or step function time domain response.

RFD-0493 (April 1993)

"A Resistive Attenuator Program" by Jouni Verronen. A utility for calculation of resistor values for T, Pi and minimum-loss pads.

Call or write for a list of all available programs

Each month's program disk is \$15.00 postpaid to U.S. and Canada. Other foreign orders add \$8.00 per order S&H. Specify 5-1/4" or 3-1/2" disks. Master Card, VISA and American Express charges accepted by telephone, or send check or money order to:

RF Design Software Service
P.O. Box 3702
Littleton, CO 80161-3702
(303) 770-4709

Measure Up With Coaxial Dynamics Model 83000A RF Peak Reading Wattmeter

Take a PEAK with Coaxial Dynamics "NEW" Model 83000A, designed to measure both FWD/RFL power in CW and FM systems simply and quickly. Then with a "FLIP" of a switch, measure "PEAK POWER" in most AM, SSB or pulse systems. Our Model 83000A features a complete selection of plug-in-elements plus a 2 year warranty. This makes the Model 83000A an investment worth looking at. So go ahead, take a "PEAK", you'll like "WATT" you see!

Contact us for your nearest authorized Coaxial Dynamics representative or distributor in our world-wide sales network.

Available With Elements Up To 1800 MHz!
Freq. Range - .45-2300 MHz



**COAXIAL
DYNAMICS,
INC.**

15210 Industrial Parkway
Cleveland, Ohio 44135
216-267-2233
1-800-COAXIAL
FAX 216-267-3142

Service and Dependability... a Part of Every Product



ELNEC Advanced Antenna Analysis Program

Fast to learn and easy to use, ELNEC lets you analyze nearly any type of antenna in its actual operating environment. Describe your antenna with ELNEC's unique menu structure and spreadsheet-like entry system and watch it make azimuth and elevation plots, report beamwidth, f/b ratio, takeoff angle, gain, and more. Save and recall antenna files. Print plots on your dot-matrix or laser printer.

ELNEC uses the full power of MININEC code but makes description, analysis, and changes worlds easier and faster. With ELNEC there's no messing with "pulses" - just tell it where on a wire you want a source or load to go, and ELNEC puts it there. And keeps it there, even if you change the antenna. ELNEC has a host of other features to make analysis fast and easy. The MaxP option extends ELNEC's capability to very complex antennas (up to 260 "pulses").

ELNEC runs on any PC-compatible computer with at least 512k RAM, CGA/EGA/VGA/Hercules, and 8/9 or 24 pin Epson-compatible or HP LaserJet/DeskJet printer. (The MaxP option requires a hard disk, coprocessor, and 640k RAM.)

There's no copy-protection hassle with ELNEC - it's not copy protected. And of course there's extensive documentation.

ELNEC is only \$49.00 postpaid. MaxP is \$25.00. (Please add \$3.00 per order for airmail outside N. America.) VISA and MasterCard orders are accepted -- please include card number and expiration date. Specify coprocessor or noncoprocessor type. Order or write for more information from:

Roy Lewallen, W7EL
P.O. Box 6658
Beaverton, OR 97007

1691 MHZ WEATHER SATELLITE SYSTEM

- 1691 MHz HEMT Pre-amp.
model TS-1691-P Amp\$450
- 1691 MHz Receiver
model TS-1691-Recvr\$450
- Decoder Board & Software
Model TS-VGA-SAT4\$399
- Low Loss (microwave) Coaxial Cable
(65ft) with connectors
model 1691-coax ass'y\$70
- Track II IBM Satellite Orbital
Program Tracks All Satellites,
World Map, Print Out\$99
- 1691 MHz Loop-Yagi Antenna
model 1691-LY(N)\$99
- 1691 MHz Loop-Yagi Extension
Model 1691-LY-XTN\$85
- Demonstration Disc (IBM-PC VGA
compatible) of signals recorded from
WX-SAT system\$3

J-BEAM'S AND LOOP YAGIS

ANTENNAS

137 MHz	2XY-137-C	\$95.00
137 MHz	5XY-137-C	135.00
1691MHz	1691LY:N:	99.00
1691 MHz	1691 LY EXTN.	85.00
70/MBM48		115.00
70/MBM88		160.00
1268 MHz	1268LY:N:	99.00
1268 MHz	1268 LY EXTN.	85.00
1296 MHz	1296 LY:N:	99.00
1296 MHz	1296 LY EXTN.	85.00

HARNESSE KITS

PMH 2-70.....	29.00
PMH 4-70.....	55.00

LOOP YAGI ACCESSORIES AVAILABLE

Prices Subject to Change Without Prior Notice.
All Antennas include 50 ohms Balun.
Write for detailed specs.



si

Spectrum International, Inc.
P.O. Box 1084 Dept. Q
Concord, MA 01742 U.S.A.
Phone: 508-263-2145
FAX: 508-263-7008

DIGITAL METER & LCR METER



- DMM 2360**
\$129.95
DMM + LCR Meter
Very Versatile DMM
- Inductance: 1μH - 40H
 - Capacitance: 1pF - 40 μF
 - Temperature: -40 - 302°F using diode-type temp. probe TP-DO5 \$7.00
 - Frequency: 1Hz - 4MHz
 - Logic test: 20MHz
 - Diode test
 - Continuity beeper
 - Volt, current, ohm
 - 3999 count display
 - Peak hold
 - Auto power off
 - 10 MΩ input impedance
 - Complete with test leads, soft case, 9 V battery, spare fuse and manual



- LCR 814**
\$199.95
Best Handheld LCR Meter
- Inductance: 0.1 μH-200H
 - Capacitance: 0.1 pF-20,000 μF
 - Impedance: 1 mΩ-20MΩ at 1kHz
 - Dissipation Factor / Q Factor measurements
 - Zero Adjustment to compensate for parasitics
 - Test frequency 1kHz
 - Best for small component for high frequency application
 - Tweezer type test fixture for surface mount component \$25.00
 - Padded deluxe case \$5.00

ALFA ELECTRONICS

P.O. BOX 8089, PRINCETON, NJ 08543
(800) 526-ALFA / (609) 275-0220
FAX: (609) 275-9536

QUALITY DMM, GOOD RELIABILITY
15 DAY MONEY BACK GUARANTEE
1 YEAR REPLACEMENT WARRANTY
WRITE OR CALL FOR FREE CATALOG
Visa, MasterCard, American Express C.O.D. Purchase Order Welcome

ATV CONVERTERS • HF LINEAR AMPLIFIERS

DISCOVER THE WORLD OF FAST SCAN TELEVISION



AMATEUR TELEVISION CONVERTERS
ATV3 420-450 (GaAs-FET) . \$ 49.95 Kit
ATV4 902-928 (GaAs-FET) . \$ 59.95 Kit

2 METER VHF AMPLIFIERS
35 Watt Model 335A \$ 79.95 Kit
75 Watt Model 875A \$119.95 Kit
Available in kit or wired/tested

For detailed information and prices, call or write for our free catalog.

HF AMPLIFIERS per MOTOROLA BULLETINS
Complete Parts List for HF Amplifiers Described in the MOTOROLA Bulletins.

AN758 300W	\$154.15	AR313 300W	\$366.00
AN762 140W	\$ 95.15	AR305 300W	\$346.82
AN779L 20W	\$ 83.79	EB63 140W	\$ 89.85
AN779H 20W	\$ 93.20	EB27A 300W	\$136.80
		EB104 600W	\$371.85

New!! 1K WATT 2-50 MHz Amplifier
POWER SPLITTERS and COMBINERS
2-30MHz

600 Watt PEP 2-Port \$ 69.95
1000 Watt PEP 2-Port \$ 79.95
1200 Watt PEP 4-Port \$ 89.95

100 WATT 420-450 MHz PUSH-PULL LINEAR AMPLIFIER - SSB-FM-ATV
KEB67-PK (Kit) \$159.95
KEB67-PCB (PC Board) \$ 18.00
KEB67-I (Manual) \$ 5.00

UNIVERSAL DIGITAL FREQUENCY READOUT
TK-1 (Wired/tested) \$149.95

HEAT SINK MATERIAL
Model 99 Heat Sink (6.5 x 12 x 1.6) \$ 24.00
CHS-8 Copper Spreader (8 x 6 x 1/8) \$ 22.00

We also stock Hard-to-Find parts

CHIP CAPS - Kemet/ATC
METALCLAD MICA CAPS - Unelco/Semco
RF POWER TRANSISTORS
Mini-Circuit Mixers

SBL-1 (1-500Mz) \$ 6.50
SBL-1X (10-1000Mz) \$ 7.95

ARCO TRIMMER CAPACITORS
VK200-20/4B RF Choke \$ 1.20
56-590-65-3B Ferrite Bead \$.20
Broadband HF Transformers
Low Pass Filters
for Harmonics (Up to 300W)
10m, 15m, 20m, 40m, 80m & 160m

Add \$4.00 for shipping and handling

CCI Communication Concepts Inc.

508 Millstone Drive • Beavercreek, Ohio 45434-5840
(513) 426-8600 • FAX (513) 429-3811



VARI-NOTCH® DUPLEXERS

FOR 2 METERS

The TX RX Systems Inc. patented Vari-Notch filter circuit, a pseudo-bandpass design, provides low loss, high TX to RX, and between-channel isolation, excellent for amateur band applications. TX RX Systems Inc. has been manufacturing multicoupling systems since 1976. Other models available for 220 and 440 MHz, UHF ATV and 1.2 GHz.

MODEL 28-37-02A

144-174 MHz
92 dB ISOLATION AT 0.6 MHz SEPARATION
400 WATT POWER RATING

TX RX SYSTEMS INC.

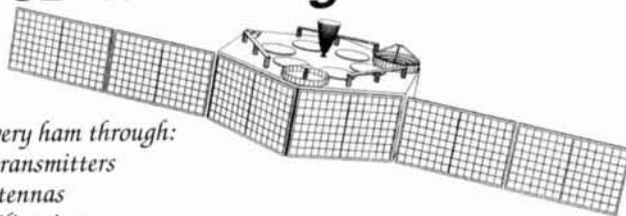
8625 INDUSTRIAL PARKWAY, ANGOLA, NY 14006
TELEPHONE 716-549-4700 FAX 716-549-4772 (24 HRS.)



19" RACK MOUNT



Phase 3D Will Bring:



Space to virtually every ham through:

- Higher power transmitters
- Higher gain antennas
- Three axis stabilization

More bands:

- Downlinks on 10 Meters, 2 Meters; 70, 13, 5, and 3 Cm.
- Uplinks on 70, 23 and 5 Cm.

Higher Orbit:

- 48,000 km. (29,000 mile), 16 hour orbit for increased coverage, and visibility for many hours each day.

Phase 3D Needs Your Help!

A contribution of just a few dollars from *EVERY* active amateur will insure that this new marvel gets into orbit.

Make Your Contribution Today!

You are also invited to join AMSAT and receive the AMSAT Journal so that you too, can follow the progress of Phase 3D and other exciting amateur satellite activities. Dues are just \$30 annually in the U.S. and \$36 in Canada and Mexico - \$45 elsewhere



AMSAT

P.O. Box 27
Washington, D.C. 20044
(301) 589-6062, FAX: (301) 608-3410

TOROID CORES



- Iron Powder
- Ferrite
- Shielding Beads
- Ferrite Rods
- Split Beads

Small orders welcome. All items in stock for immediate delivery. Low cost experimenter's kits: **Iron Powder, Ferrite.** The dependable source for toroidal cores for 25 years.

Call or write for free catalog and tech data sheet.

PALOMAR ENGINEERS

Box 462222, Escondido, CA 92046
Phone: (619) 747-3343
Fax: (619) 747-3346

ORGANIZE AND PROTECT YOUR COPIES OF Communications Quarterly

Now there's an easy way to organize and keep copies of your favorite magazine readily available for future reference.

Designed exclusively for **Communications Quarterly** by Jesse Jones Industries, these custom-made titled cases and binders provide the luxury look that makes them attractive additions to your bookshelf, desk or any location in your home or office.

Each binder/case is covered with durable leather like material in burgundy, title is hot-stamped in gold.

Whether you choose cases or binders, you'll have a storage system that's durable and well organized to help protect your valuable copies from damage.

Communications Quarterly

Jesse Jones Industries, Dept. Quarterly
499 East Erie Avenue,
Philadelphia, PA 19134

Please send _____ cases; _____ binders

Enclosed is \$ _____

Charge my: (Minimum \$15)

American Express

Visa

Mastercard

Diners Club

Card # _____

Exp. Date _____

Signature _____

Print Name _____

Address _____

No P.O. Box Numbers Please

City/State/

Zip _____

PA Residents add 7% sales tax, Call TOLL FREE 7 days, 24 hours 1-800-825-6690

Quantity	Cases	Binders
One	\$ 7.95	\$ 9.96
Three	\$21.95	\$27.95
Six	\$39.95	\$52.95

Add \$1 per case/binder postage and handling. Outside USA \$2.50 per case/binder. (U.S. funds only)

ADVERTISER'S INDEX

AEA/Adv. Elec. Applications.....	5
AMSAT.....	112
Alfa Electronics.....	111
Alinco Electronics.....	COV. II, 1
Amidon Associates.....	108
Astron Corporation.....	107
Beezley, Brian, K6STI.....	109
C & S Sales.....	109
Coaxial Dynamics.....	110
Communications Concepts, Inc.....	111
Communications Specialists.....	109
CQ Video Productions.....	11, 109
Down East Microwave.....	108
Drake, R.L.....	106
HAL Communications.....	8
Japan Radio, JRC.....	7
Jesse Jones Industries.....	112
Kenwood USA.....	COV.IV, 2
L.L. Grace Comm. Products.....	10
Lewallen, Roy, W7EL.....	110
Motorola.....	12
Nemal Electronics.....	108
OPTOelectronics.....	9
PC Electronics.....	110
Palomar Engineers.....	112
R.F. Design Software Service.....	110
Spectrum International, Inc.....	111
Surplus Sales of Nebraska.....	108
TX RX Systems Inc.....	111
Yaesu.....	COV.III

Only the best companies advertise in CQ. Learn how easy it is to enjoy the benefits the great CQ audience has to offer. Call Arnie Sposato, N2IQO, at (516) 681-2922 or FAX (516) 681-2926.

"This FT-890AT is great in the field!"

"Yaesu did it again!"

FT-890AT Compact HF Transceiver

- Automatic Antenna Tuner Hybrid High Speed Design Covers 160-10 meters
- I.P.O. Intercept Point Optimization
- DDS-Direct Digital Synthesis
- F.S.P. Frequency Shift Speech Processor
- General Coverage Receiver 100 KHz to 30 MHz
- Pass Band Shift and 30db Notch Filter
- Noise Blanker with Adjustable Pulse Width
- Built-In Iambic Keyer
- 32 Memories plus 20 VFOs
- FM Repeater Operation Automatic 10 Meter Repeater Offset w/Selectable CTCSS Encode
- All Mode Squelch
- DFCS-Duct Flow Cooling System
- **Accessories:**
Contact your Dealer for full details.



"It's the world's smallest HF rig with a built-in antenna tuner."

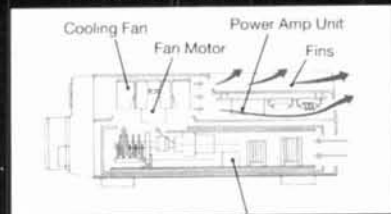
Field days and contesting are challenging. We built the FT-890AT for times when you need the high performance of a base station – like the FT-1000 – but the practicality of a compact, rugged mobile. In fact, the FT-890AT is the mobile version of the FT-1000. Designed to be the world's smallest HF with a built-in antenna tuner, its superior receiving performance is a direct result of FT-1000 technology.

Since field work is demanding, the front panel has been simplified. Seldom used VOX controls are on the back. For faster TX/RX switching, the FT-890AT has two direct digital synthesizers (DDSs). With its unique duct flow cooling system, die-cast aluminum upper case and heat sink, the FT-890AT can't be beat for superior field work and DX-peditions.

Not just a "field" radio, with the optional FP-800 AC Power Supply, MD-1C8 Desk Mic and YH-77ST Headphones you've got a performance-plus base station loaded with features and affordably priced.

To see what that means for you, contact your Yaesu dealer today.

Duct Flow Cooling System (DFCS)
Rugged aluminum top panel heat sink and internal thermally switched fan draws air through the heart of the transceiver.



Built-in Antenna Tuner
Tunes most antennas 160-10 meters.

YAESU
Performance without compromise.SM

Usually if a rig is this loaded, the price is too.



DVS-2 (Optional)
Digital Voice Recorder
with 16 seconds of memory
each on RX and TX.

© 1992 Yaesu USA, 17210 Edwards Road, Cerritos, CA 90701 (310) 404-2700

Specifications subject to change without notice. Specifications guaranteed only within amateur bands. Some accessories and/or options are standard in certain areas. Check with your local Yaesu dealer for specific details.

KENWOOD

...pacesetter in Amateur Radio

Compact Milestone

Kenwood unveils the world's smallest HF transceiver

HF is going places—thanks to the smallest transceiver of its kind in the world. Providing high-performance communications with go-anywhere convenience, the freedom. And whether used

TS-50S
HF TRANSCEIVER

DX-peditions, or in a fixed installation, this rig packs a powerful punch. Maximum output is 100W, and there's a full range of advanced features—including 100 memory channels, DDS with innovative "fuzzy" control, and AIP for superior dynamic range IF shift and CW reverse mode help reduce interference, while a noise blanker improves clarity. For user-friendly operation on the move, there's a multi-function microphone and powerful menu system. And the TS-50S is fully equipped for split-frequency operations. Test drive one today.

to Kenwood's new TS-50S, its kind in the world. Providing communications with go-anywhere convenience, the TS-50S is your passport to freedom for mobile operations and



NEW

KENWOOD COMMUNICATIONS CORPORATION
AMATEUR RADIO PRODUCTS GROUP
P.O. BOX 22745, 2201 E. Dominguez Street
Long Beach, CA 90801-5745

KENWOOD ELECTRONICS CANADA INC.
6070 Kestrel Road, Mississauga,
Ontario, Canada L5T 1S8

93-ARD-0599