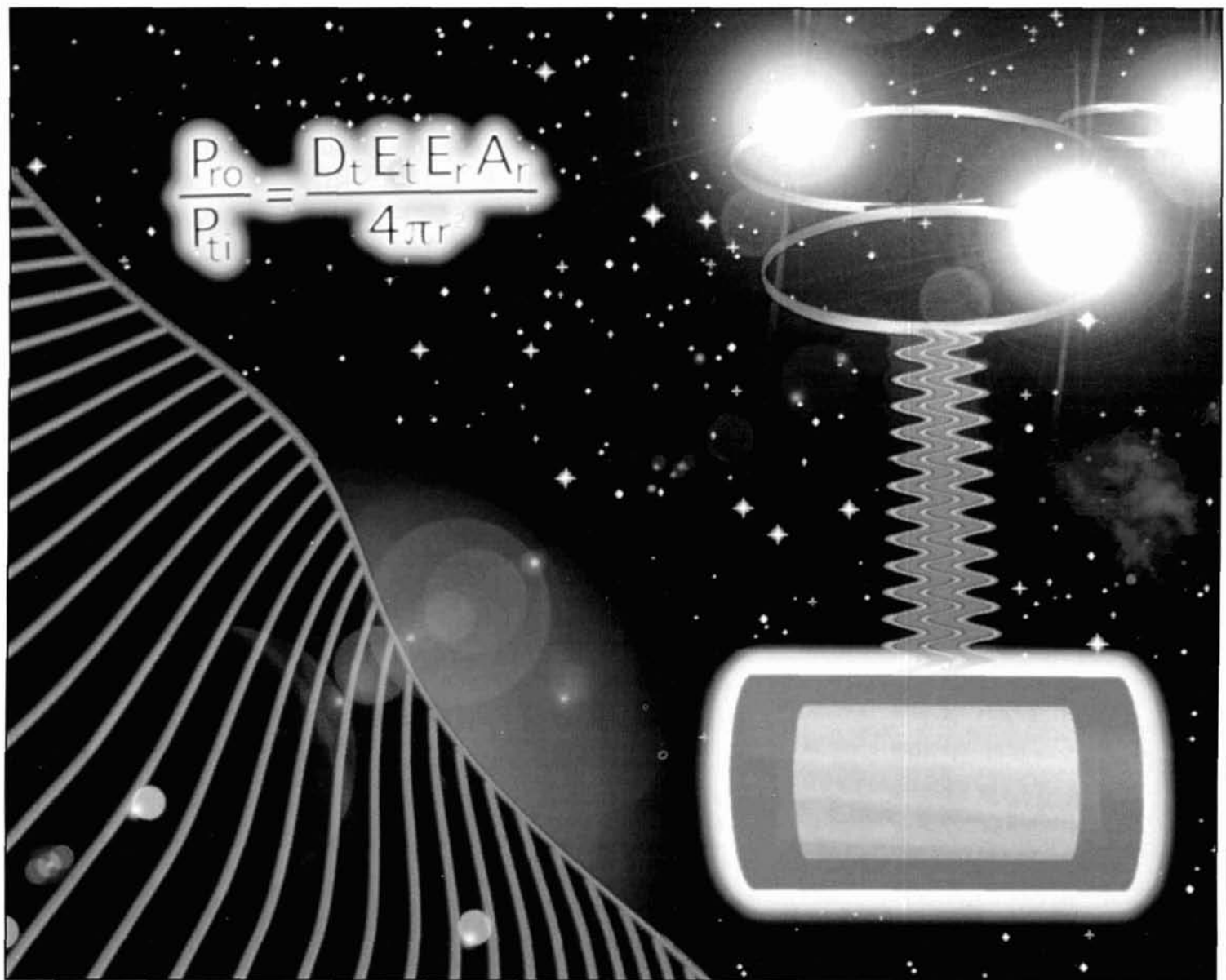


# COMMUNICATIONS QUARTERLY

THE JOURNAL OF  
COMMUNICATIONS  
TECHNOLOGY

Fall 1993

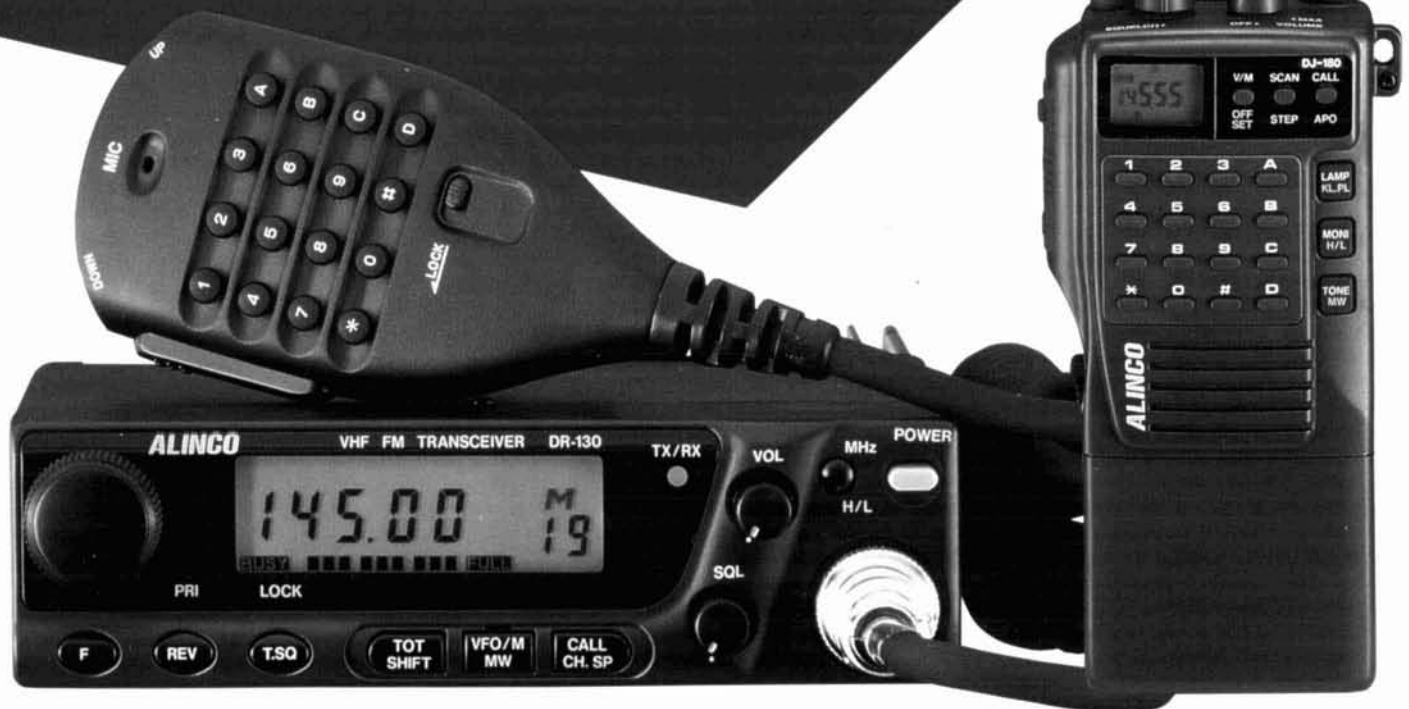
\$9.95



- Gravity Wave Communications
- Connecting Computers To Radios
- Spread Spectrum Communications
- High-Frequency Bypass Capacitors

- Transmitting Short Loop Antennas  
For The HF Bands: Part 2
- The Case Of The Invisible Meteor Storm
- The Uni-Directional Long Wire Antenna

# DYNAMIC DUO



## DR-130T

### Works Hard / Runs Cool

ALINCO's newest 2-Meter mobile, the DR-130T, packs a big punch. This compact radio delivers 50 Watts of cool running power, and offers the durability and reliability that Hams have come to expect from ALINCO.

Standard features include 50 CTCSS Tones, Programmed Memory scan, Programmable "Time-Out" Timer, CTCSS Encode, and others. With the optional EJ-19U plug-in module, 100 memory channels are available. All memory channels can store "odd-split" frequencies, and also store CTCSS Encode/Decode status.

CHECK OUT THE AFFORDABLE  
TECHNOLOGY OF THE 90'S.  
CHECK OUT ALINCO.

 **ALINCO**  
ELECTRONICS INC.

**NEW**

## DJ-180T

### Clearly Superior

Ergonomic, rugged design, combined with excellent sensitivity and great sound make this the radio of choice for demanding operators.

This rig comes standard with CTCSS encode and decode. Ten memory channels come standard, and the unit can be upgraded to 50, or even 200 Memory Channels with optional plug-in chips.

Odd Splits! This radio can store repeater offsets from 0 to 15.995 MHz. A different offset can be stored in each memory channel, and most other functions can also be stored independently in each memory channel.

For brochures and operation manuals,  
try our new 24 hour Hot Line.

Telephone: **(310) 618-9017**

438 Amapola Ave., #130 Torrance, CA 90501  
Phone:(310) 618-8616 / Fax:(310) 618-8758

# LIGHTS OUT!!

## OPTOELECTRONICS turns out the lights on LED Counters...

...with the new 10 digit LCD MiniCounter.  
More PERFORMANCE  
In a SMALLER package

- 1 MHz - 2.8GHz Frequency Range
- Direct Count Range From 1MHz to 250MHz with 1Hz/Sec high resolution display.
- Prescaled Range from 10MHz to 2.8GHz for virtually all two way communications
- Select up to 6 Gate/Measurement Periods
- 10MHz Industry Standard Time Base
- Ultra Compact - true pocket size
- Maximized sensitivity for picking up radio transmissions from the greatest distance
- Display Hold Switch Locks current Measurement

Introductory Price  
**\$129.00**  
MADE IN THE USA

The New Model 3300 MiniCounter breaks the LED counter tradition. From our first Model FC50 (1976) to our latest Model 2300, we have always had quality LED counters. But LEDs are power hungry, failure prone and unreadable in sun light. So the LED counter has now gone the way of the LED wristwatch—replaced by LCD. The LCD advantages include lower power, better reliability, smaller package, more information displayed and better viewing characteristics.

**ACTUAL SIZE!**  
(Actual Size Hand!)

**Model M1 & 3000A Hottest on the Market with:**  
**Digital Filtering** for the fastest method to reduce false counts—no loss of sensitivity & **Digital Auto Capture** that auto holds and stores - working even near strong RF fields!

- **3000A-Multifunction HandiCounter®**
  - 15 gate times selectable • 7 hrs. Batt. life
- **M1-Full Range HandiCounter®**
  - 10 Gate times • 6 hrs. Battery life

**Both offer:**

- OE10 Ultra Fast & Reliable Counter IC
- Standard Backlit 10 digit LCD Display
- 16 Segment Signal Strength Bargraph
- 3 Data Storage Registers
- 1.3% of a second Measurement Rate
- 1Hz Resolution in 1 Sec. up to 250MHz
- 2 Wire Serial Output for Data Logging

**3000A** .....\$329.  
**M1** .....\$229.



# OPTOELECTRONICS

**TOLL FREE ORDER LINE**  
**1-800-327-5912**

305-771-2050 • FAX 305-771-2052  
5821 NE 14th Ave, Ft. Laud., FL 33334  
5% Ship/Handling (Min \$5 & Max \$10) U.S. & Canada.  
15% outside continental U.S.  
Visa, Master Card, C.O.D., Cash or Money Order only.

# Kenwood's TS-50S - Awesome Versatility



## TS-50S HF TRANSCEIVER

### Features

- 500kHz-30MHz general coverage receiver
- DDS (Direct Digital Synthesizer) with fuzzy logic control
- Large LCD panel with digital bar meter
- Auto-mode capability
- Menu system
- AIP (Advanced Intercept Point)
- Switchable AGC circuit (SLOW/FAST)
- All-mode squelch
- CW reverse mode
- Full break-in and semi break-in
- 20dB attenuator
- Multi-function microphone supplied
- RF output power control (100W, 50W, 10W)
- Optional 500Hz CW filter (YK-107C)
- Optional external antenna tuner (AT-50)

For the Amateur Radio enthusiast, going "beyond bounds" is what it's all about. That's why Kenwood created the TS-50S, the world's smallest and smartest HF transceiver. The choice is yours: you can mount it in a vehicle, take it on a DX-pedition, or even install it permanently as a base station transceiver. Yet despite its size, the TS-50S provides a maximum output of 100W and the sort of sophisticated features normally found only inside a shack. Take for example the 100 memory channels for independent storage of transmit/receive parameters, the microprocessor-controlled DDS with innovative "fuzzy" control, and Kenwood's own AIP for superior dynamic range. There's also a powerful menu system, IF shift and CW reverse mode for interference reduction, TF-SET, and a noise blanker—plus everything you need for split-frequency operations. Heavyweight performance no longer means heavyweight dimensions. So if you feel the call of the wild, check out the TS-50S.

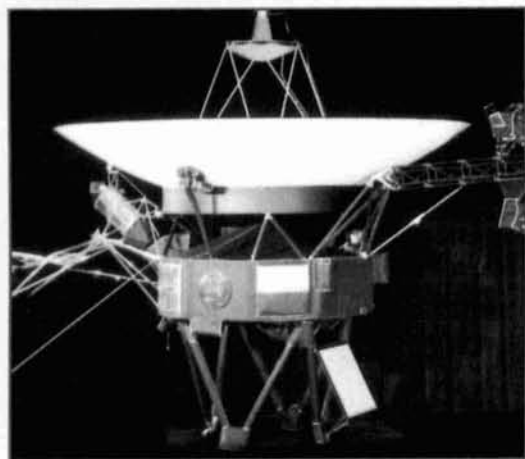
# COMMUNICATIONS QUARTERLY

THE JOURNAL OF  
COMMUNICATIONS  
TECHNOLOGY

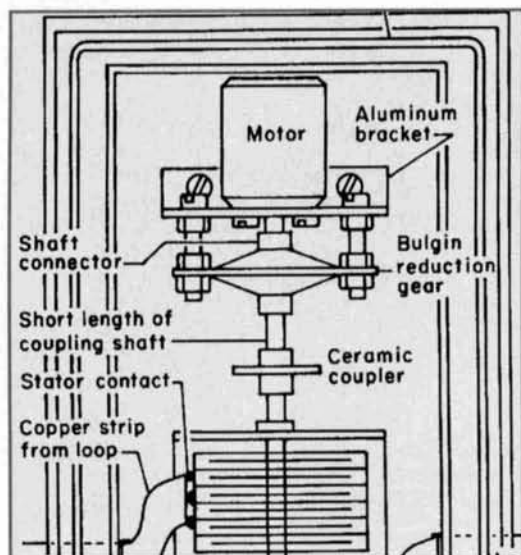
## CONTENTS

Volume 3, Number 4

Fall 1993



Taylor, page 84



Craighero, page 95

- 6 Technical Conversations**  
*Jerry Sevick, W2FMI; and Mason A. Logan, K4MT*
- 13 Connecting Computers To Radios**  
*Howie Cahn, WB2CPU*
- 30 Gravity Wave Communications**  
*Jim Peterson, AA6OZ*
- 35 The Uni-Directional Long Wire Antenna**  
*R.P. Haviland, W4MB*
- 42 The Case Of The Invisible Meteor Storm**  
*Joseph L. Lynch, N6CL*
- 45 High-Frequency Bypass Capacitors**  
*M.E. Gruchalla, P.E.*
- 67 Quarterly Devices**  
*Rick Littlefield, K1BQT*
- 71 Spread Spectrum Communications**  
*Bryan Bergeron, NU1N*
- 84 The Solar Spectrum**  
*Peter O. Taylor*
- 87 Tech Notes**  
*Peter Bertini, K1ZJH; Robin Ramsey, ZL3TCM; and Pat Hawker*
- 95 Transmitting Short Loop Antennas For The HF Bands:  
Part 2**  
*Roberto Craighero, I1ARZ*

Cover photo: Could gravity wave communications be the next amateur radio frontier? Our cover photo, by Bryan Bergeron, NU1N depicts this intriguing new mode. To find out more, see this issue's feature article "Gravity Wave Communications—The Ultimate Frontier" by Jim Peterson, AA6OZ, on page 30.

# EDITORIAL

## Seamless Communications

The FCC has shaken up the commercial Land Mobile Radio industry with its proposed rewrite of Part 90 rules governing the LMR spectrum. Part 88 would replace the existing Part 90 rules in two stages; by the year 1996, LMR users in the 150 to 174-MHz band will need to reduce deviation from 5-kHz to 3 kHz, and to 2 kHz by the year 2006. The 421 to 512-MHz users (excluding amateurs, who are governed under Part 97 of the rules) will be limited to a maximum 2.5-kHz deviation in the last stage. The Part 88 rules, if enacted, will lead to a \$40 billion economic stimulus package. By the year 1996, amateurs could be reaping a windfall of prime used commercial mobile and repeater equipment that no longer meets FCC type acceptance—and at firesale prices. Will this potential windfall benefit amateur radio, or will it delay the inevitable by further locking us into outmoded technology?

Combining the best of technologies now in use for cellular and 800-MHz trunking systems with the new digital communications modes that must be developed to meet the requirements of the new Part 88 rules in its last stages of implementation, could lead amateurs into a new era. "Seamless Transportation" is the latest buzzword in the transportation industry. In effect, it means a shipper is assured of timely and efficient delivery of his product between points A and B by dealing with a single vendor—despite the number of shippers handling the product en route. Should amateur VHF operators be working toward developing a "Seamless Communications" system, where we can communicate with friends over large distances through a complex, digital mode trunking system, without concerning ourselves with what "channel" or band we are on, or what clubs or satellites are offering the trunking and linking services providing the communications? The day when we can commute to work and talk with a VHF mobile station in England heading home from a hard day at the office isn't that far away. The lines dividing the worlds between the analog RF and digital fields are fading rapidly.

Unfortunately, amateur claims on our limited VHF/UHF spectrum are presently territorial at best. Repeater frequencies are coordinated, and then jealously guarded against intruders by their "owners." In most metropolitan areas, repeater pairs are assigned in crowded 15-kHz channel spacings—a marginally workable compromise for our present 5 kHz "narrowband" standard. Frequencies once assigned for simplex FM voice operations are quickly being overrun by the increasing demands of the packet community—

using a protocol that is now outdated, excruciatingly slow, and overloaded by modern digital standards. The demands for more VHF spectrum will not abate. The recent influx of new amateurs attests to that. Another commercial claim on our VHF/UHF spectrum, mirroring the loss of the 220 to 222-MHz amateur spectrum, could bring the whole house of cards tumbling down around us as repeater owners, packet operators, ATV operators, and narrowband mode proponents scramble over whatever spectrum is left of our finite frequency resources.

There are two sides to every coin. Is there a real need for amateur radio to keep in lockstep with every new advancement in the communications field? CW communications is as old as radio itself, and after nearly 100 years of existence it is rapidly fading from use. The Coast Guard has recently suspended all monitoring of the CW distress frequencies. However, many hams continue to find CW to be a relaxing and enjoyable mode. It remains the mode of choice for EME, scatter, and other communications needs—and is spectrum efficient. Amateur radio is still a hobby.

Only twenty-five years ago, who would have dreamed that VCRs, cellular phones, fiber optic technology, or powerful home computers would be so commonplace? Twenty-five years ago, most of our VHF and UHF spectrum was an unused wasteland. Now there doesn't seem to be enough to go around. FCC rule changes led to the influx of cheap and readily available wideband communications equipment that changed the face of and populated the amateur VHF/UHF bands in the early '60s, just as the latest proposed FCC rules will sow the new technologies that will change are amateur bands at some future date. One other thing is sure—*Communications Quarterly* will be here to guide are readership through the latest advances in communications technology as we head into the next quarter century.

**Peter J. Bertini, K1ZJH**  
Senior Technical Editor

*And lest you repeater owners and packet operators feel singled out for chastisement, you need to know that K1ZJH is a trustee and originator of the 146.94 and 443.20 repeaters on Mt. Tom in Holyoke, Massachusetts. This is probably one of the most used and oldest repeater systems in New England. Peter also operates a K-node packet digipeater on Soapstone Mountain in Somers, Connecticut on 145.09.*

**KA1STC, Editor.**

### 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, W6SAI

### 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

**Editorial Offices:** P.O. Box 465, Barrington, NH 03825. Telephone/FAX: (603) 664-2515.

**Business Offices:** 76 North Broadway, Hicksville, NY 11801. Telephone: (516) 681-2922. FAX: (516) 681-2926.

Communications Quarterly is published four times a year (quarterly) by *CQ Communications, Inc.* Subscription prices: Domestic—one year \$29.95; two years \$56.95. Foreign—one year \$39.95; two years \$76.95. Foreign Air Mail—one year \$60.95; two years \$118.95. Contents copyrighted CQ Communications, Inc. 1993. Communications Quarterly does not assume responsibility for unsolicited manuscripts. Allow six weeks for change of address. 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.



# Put yourself in the winner's circle with the PK-232MBX multi-mode controller.

The PK-232MBX sets the pace for multi-mode controllers. While the competitors are racing to catch up, the PK-232MBX surges into the lead, winning more contests and providing the best value of any multi-mode controller on the market today.

Leave the competition in the dust—with the highest quality filters available (seventeen-pole effective filtering), the PK-232MBX pulls signals out of the noise better than any other controller.

Satisfy your drive for variety with PACTOR, Packet, AMTOR/SITOR, Morse, RTTY (Baudot/ASCII), WEFAX, NAVTEX, and TDM modes.

You won't spin your wheels trying to identify a signal. Our revolutionary SIAM™ technology automatically identifies the incoming signals, switches the PK-232MBX to the recognized mode, and starts displaying the data.

PACTOR, Packet and AMTOR mailboxes—with selective control of third party traffic—are standard to the PK-232MBX.

Our extensive Host Mode control gives you unmatched power and versatility from computer support programs for DOS, Windows™, Macintosh, and Commodore computers.

Our highly trained service crew keeps you on track with helpful advice and in-depth knowledge of amateur radio.

When you're ready for a multi-mode controller with a proven track record, the PK-232MBX is the choice of champions.

For more information on the PK-232MBX, call AEA's Literature Request Line at (800)432-8873 or contact your favorite ham radio equipment dealer.



*Connect with us*

# TECHNICAL CONVERSATIONS

*At the request of our Consulting Technical Editor Bob Wilson, WA1TKH, Jerry Sevick, W2FMI, explains why he didn't use polarity dots in his recent articles on baluns that appeared in Communications Quarterly.*

## Dear Editor:

On several occasions I have been asked why I don't use the polarity dots common in the schematic diagrams of conventional transformers. They are used to show the polarity of the output voltage with respect to the input voltage or the direction of the currents for mutually aiding flux in the core. These dots are very popular in the amateur radio literature.

A few years ago, I would have given the following two reasons: 1) My formal training didn't include this technique as part of the curriculum, which was known as Engineering Science and Applied Physics,\* and 2) I learned about these broadband transformers from a colleague at Bell Labs, Clyde Ruthroff,<sup>1</sup> who treated these devices with transmission line theory. Therefore, his classic 1959 paper, which became the industry standard, didn't use the dots. Furthermore, Guanella<sup>2</sup>—who really started the concept of transmission line transformers with his classic 1944 paper—didn't use the dots either. In fact, his coiled transmission lines didn't even use cores.

But lately I would add another reason for not using polarity dots and that is *perception*. From contacts "on the air" and from reading the amateur radio literature, I see that practically everyone still perceives these devices as conventional transformers that transmit the energy from input to output by flux linkages. They don't look at these devices (as Ruthroff and Guanella did) as chokes that isolate the input from the output and then as transmission lines that transmit the energy by an efficient transmission line mode. In fact, very accurate insertion loss measurements<sup>3</sup> on devices using low permeability ferrite and powdered-iron cores, have shown losses as low as 0.02 to 0.04 dB. A graphically proof of this is seen when small transformers using transmission lines of no. 18 wire wound on 1-inch OD toroids or 1/4-inch diameter rods handle 500 watts of power without any sign of thermal runaway.

I have also been told lately that this class of matching devices known as *transmission line transformers*, is not an emerging technology. It is a technology that came of age quite a few years ago. But, from my vantage point, I disagree with this view. This class of transformers has not reached its potential in power, high-frequency performance, and applications because: 1) This class of transformers has not been included in the electrical engineering curriculums, and 2) companies are reluctant to publish important technical papers for fear of aiding their competition. In fact, this latter reason is also prevalent in other technologies as is evidenced by the lack of application papers in the professional literature. It has been reported that, in the last fifteen years, 85 percent of the technical papers are generated by the colleges and universities, not industry!

And finally, maybe a name-change would be helpful in understanding these devices. Instead of *transmission line transformers*, why not use *broadband*

*transmission line matching networks*? It would certainly help in doing away with the polarity dots.

**Jerry Sevick, W2FMI**  
Basking Ridge, New Jersey

## REFERENCES

1. C.L. Ruthroff, "Some Broad-Band Transformers," *Proceedings IRE*, Volume 47, August 1959, pages 1337-1342.
2. G. Guanella, "Novel Matching Systems for High Frequencies," *Brown-Boveri Review*, Volume 31, September 1944, pages 327-329.
3. J. Sevick, *Transmission Line Transformers*, 2nd edition, Amateur Radio Relay League, Newington, Connecticut, 1990, Chapter 12.

## FOOTNOTE

\*Harvard University Graduate School.

*Mason Logan, K4MT, had these comments on the use of transformers in noise bridges.*

## Dear Editor:

Existing noise bridges use a transformer whose input winding couples the unbalanced noise source to a center-tapped floating winding. This center-tapped floating winding forms the 1:1 ratio arms of the noise bridge. However, while transformers are very good, they are not ideal. For instance, lack of perfect equality in the two arms of the bridge and the effect of the uncontrolled parasitic winding capacitances result in built-in bridge errors.

Because the transformer has been so crucial, almost every noise bridge article has a new transformer design.<sup>1,2,3</sup>

A transformer is not necessary. The circuit schematic for a noise bridge without a transformer is shown in **Figure 1**. A double-shielded floating noise source and a pair of matched metal film resistors replace the transformer. The matched resistors are an added load on the noise source, so more noise power is needed than for a transformer bridge.

For illustration, the circuit arrangement in the lower part of the bridge is the same as that discussed in detail in **References 1 and 3**. Its operation is unaffected by the substitution of the double-shielded noise source.

The small size of a noise source<sup>1,3,4</sup> on its perforated board (smaller than its 9-volt battery) is what makes this circuit practical. The inner shield box need not be much larger than just the battery.

The inner shield is insulated from, and centered inside the outer shield. The outer shield is insulated from the bridge enclosure and grounded to the D corner of the bridge.

The capacitance between (1) the inner and outer box shields, (2) between the battery switch wires and their shield, and (3) between the wire connecting the inner shield to the bridge corner C and its shield, are in parallel at corner C and grounded to corner D. They make up part of the center-balancing capacitor,  $C_O$ .

The capacitance from the wire connecting the noise source output to the bridge corner A, to its shield, is in parallel with the variable calibrated capacitor  $C_S$  and increases its minimum capacitance.

The selection of the fixed capacitor  $C_O$  takes all the shield capacitances into account. It is chosen to center the calibrated variable capacitor  $C_S$  when a resistor is measured as the unknown impedance.

The floating double-shielded noise source does



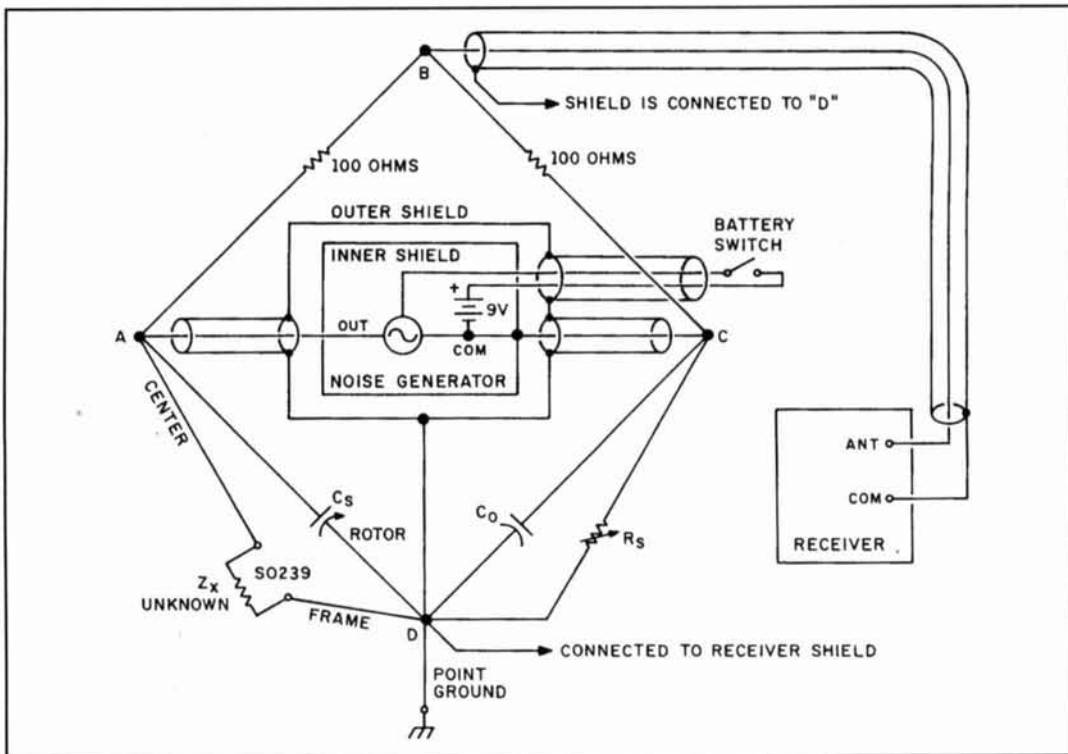


Figure 1. RF 1:1 impedance bridge with a double-shielded noise generator.

eliminate the bridge transformer errors. It will not eliminate problems in the lower part of the bridge caused by inductance of the calibrated variable resistor,  $R_s$ . But all development effort now can be directed toward solving the problems of the lower part.<sup>4</sup>

Mason A. Logan, K4MT  
Sun City Center, Florida

#### REFERENCES

1. Forrest Gehrke, K2BT. "A Precision Noise Bridge," *Ham Radio*, March 1983, page 50.
2. Wilfred N. Caron. "A Simple and Accurate Admittance Bridge," *Communications Quarterly*, Summer 1992, page 44.
3. A.E. Popodi, AA3K/OE2APM. "Building the Perfect Noise Bridge," *Communications Quarterly*, Spring 1993, page 55.
4. Forrest Gehrke, K2BT. "Technical Conversations," *Communications Quarterly*, Winter 1993, page 92.

## When we talked to technicians about servicing radios, they told us they were too busy to talk.

Now the new HP 8920A RF Communications Test Set solves test problems in less time, so you have more.

- Easy-to-use front panel
- Optional full-feature spectrum analyzer
- Rugged portable package
- Built-in computer with optional software for FM, cellular mobile and base station testing.
- Price \$13,800



**Keeping you ahead in communications.**

For more information call: 1-800-344-3802 and ask for Charlie or Ken.

 **HEWLETT  
PACKARD**

### The Quad Antenna

Hams love antenna books and this book is no exception. Written by world renowned author Bob Haviland, W4MB, *The Quad Antenna* is the authoritative technical book on the design, construction, characteristics and applications of Quad Antennas. Discover how to easily build a quad antenna for your station that will help you fill your logbook with rare DX that you have only dreamed about before.

Order No. QUAD..... \$15.95



NEW

### The Packet Radio Operator's Manual

This book is written by CQ columnist and Amateur Radio Packet authority Buck Rogers, K4ABT. An all new introduction and guide to packet operation, it is the perfect single source, whether you're an advanced user or just starting out. Learn about packet radio without all the technical jargon. Also included are detailed hookups for dozens of radio/packet controller/computer combinations, making this book the definitive resource for the active packet user.

Order No. PROM..... \$15.95



NEW

### Ham Radio Horizons: The Book

Written by Peter O'Dell, WB2D, this is a book about ham radio that every beginner can enjoy! If you want to get in on the fun and excitement of Amateur Radio, *Ham Radio Horizons* is the perfect way to get started. HRH is full of tips from expert hams in: DXing, Contesting, Serving the Public, Ham Radio in Space, Experimenting, Digital Communications — you name it! This exciting book is an excellent gift to a prospective ham or for use in your club's licensing classes and library.

Order No. BHOR..... \$12.95



NEW

### The Vertical Antenna Handbook

Take advantage of the 20 years of research and practical experience of naval communications engineer Capt. Paul H. Lee, USN(ret), N6PL. Learn the basic theory, design, and practice of the vertical antenna. Discover easy construction projects such as a four-band DX vertical or a broadband array for 80 meters. Ever wonder how to build a functional directive vertical system? Paul Lee can get you started today!

Order No. VAH..... \$9.95

### The CQ 1994 Amateur Radio Almanac First Edition

The most exciting ham radio book in years, here you have it all at your fingertips—thousands of facts and figures, maps, graphs, photos of hams and stations, a beautiful color insert, and more—in one handy volume you'll be sure to refer to over and over again. For the History... the Records... the Rules, order the Source!

Order No. BALM .....\$19.95



FIRST NEW

## Get Started with CQ's New Video Library!

Also Available in PAL format

### Getting Started in Packet Radio

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

#### Getting Started in Packet Radio



Getting Started in Packet Radio shows you how to set up the necessary equipment and actually get on the air on Packet. No theory... just the nuts and bolts of how to do it.

Order No. VPAC.. \$19.95

### Getting Started in Ham Radio

This is a fast-paced video introduction to the fascinating world of ham radio. CQ's experts show how to select equipment and antennas; which bands to use; 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.

Order No. VHR .....\$19.95

#### Getting Started in Ham Radio



Getting Started in Ham Radio walks the viewer through setting up the ham station, including the antenna, and gets you on the air.

### Getting Started in Amateur Satellites

Learn with this video how veteran operators set up their satellite stations. Find out how to locate and track ham satellites with ease. Watch as operators 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 Amateur Satellites



A newcomer's guide through the equipment, the techniques and the jargon of satellite communications.

Order No. VSAT..... \$19.95

### Getting Started in DXing

Top DXers share their experience with equipment, antennas, operating skills, and QSLing. You'll see hams work rare DX around the world. If you're new to DXing, this video is for you! All this valuable information may well give you the competitive edge you need to master the exciting world of DXing.

Order No. VDX..... \$19.95



Getting Started in DXing shows you how to set up the necessary equipment and actually get on the air on DX. No theory... just the nuts and bolts of how to do it.

### Getting Started in Contesting

For the newcomer to contesting or an experienced veteran, this video is for you! You'll get advice and operating tips from some of contesting's most successful competitors, including Ken Wolff, K1EA, Dick Newell, AK1A, and CQ's own contest columnist, John Dorr, K1AR. Here's just a sample of what you'll see: what contesting's all about, explaining contest jargon, tips for beginners, how to compete from a small station, operating secrets from the "pros", live QSOs from world class stations, VHF/UHF contesting.

Order No. VCON .....\$19.95

#### Getting Started in Contesting



Learn how to compete in the exciting world of contesting. You'll see what contesting is all about, explaining contest jargon, tips for beginners, how to compete from a small station, operating secrets from the 'pros', live QSOs from world class stations, VHF/UHF contesting.

NEW

### Ham Radio Horizons: The Video

This introduction to Amateur Radio is an excellent complement to the *Ham Radio Horizons* book. Enjoy seeing all aspects of ham radio ranging from what it takes (and costs) to get started to how you can get your ham license. Designed for the general public, HRH is ideal for public events, presentations to community groups and as an opening to your club's licensing courses! There's no better way to introduce someone to ham radio.

Order No. VHOR.....\$19.95



NEW

### 1994 Radio Classics Calendar

Imagine ham radio history being displayed on your wall with CQ's new **1994 Radio Classics Calendar**.

If you enjoy nostalgia, you'll want CQ's **1994 Radio Classics Calendar**. Each month you'll reminisce about radio history with striking photography of rare morse keys, antique radios and radio tubes.

And there's more . . .

You'll refer to your **1994 Radio Classics Calendar** time after time as you search for the schedules of upcoming ham events and conventions. Public holidays and useful astronomical data will be at right by your side, too!

Receive 15 months of use (Jan '94 - Mar '95) with custom photography that has never been published anywhere else. And all at the low price of only \$9.95 each. CQ's **1994 Radio Classics Calendar**—a great product for every nostalgia buff.

Order No. CCAL.....\$9.95



NEW

### 1994 Amateur Radio Calendar

There's no better addition to your shack or office than CQ's new **1994 Amateur Radio Calendar**.

If your interest is towering antennas with stunning scenery and shacks that go beyond belief, then you'll want the **1994 Amateur Radio Calendar** is for you. Envision seeing professional photographs of some of the biggest stations in the world!

CQ's **1994 Amateur Radio Calendar** contains dates for major operating events and conventions. In addition, it's chock-filled with public holiday and commemorative dates plus valuable astronomical information.

With CQ's **1994 Amateur Radio Calendar**, you actually receive 15 months of use (Jan '94 - Mar '95)—and with all-new photography that can't be beat! At only \$9.95 each, this is a real bargain from CQ.

OrderNo. HRCAL.....\$9.95



NEW

# New From **CQ** . . . Radio T-Shirts!!

High-Quality T-Shirts, made in U.S.A., pre-shrunk 100% cotton.

Special Introductory Offer  
**\$2 off** each shirt!

**\$15.95** per shirt plus shipping & handling regularly \$17.95



Item # 13

**Yes! Rush my CQ T-Shirt(s) to me right away! Here's my order below**

Name \_\_\_\_\_ Callsign \_\_\_\_\_

Address \_\_\_\_\_

City \_\_\_\_\_ State \_\_\_\_\_ Zip \_\_\_\_\_

Please allow 4-8 weeks delivery

Available sizes: L, XL, XXL (add \$2) Please add \$2 shipping and handling per item ordered. Order 4 or more and receive free shipping.

Make your check or money order payable to: CQ Communications, Inc., 76 North Broadway, Hicksville, New York 11801

Please charge my:

MC/VISA/AMEX/Discover # \_\_\_\_\_ Expires \_\_\_\_\_ Signature \_\_\_\_\_

Qty	Item #	Size	Price	S & H	Total
				<b>Grand Total</b>	



**Order Toll Free (800) 853-9797 or FAX (516) 681-2926**

# HAL DOES IT *FASTER*



ARQ Link To K9GWT		CLOVER		MOD	SNR	FRQ	PHS	ECC	TPR
Tune For Equal Amplitudes		MAX							
T1	████████████████████	MY	16P4A	53	0	11	0	100	
T2	████████████████████	MY	16P4A	54	0	11	0	100	
T3	████████████████████	HIS	BPSM	59	0	7	0	100	
T4	████████████████████	HIS	BPSM	58	0	7	0	100	

Now you can send and receive data at higher rates on HF with a modem that plugs into your PC. The PCI-4000/ PC-CLOVER system plugs into your 286 or higher PC and brings you a new era in data transmission over HF. The PC-CLOVER system includes binary data transfer and data compression for even higher efficiency. The CLOVER system is adaptive, automatically adjusting for differences in propagation conditions. CLOVER uses Reed-Solomon error correction to insure error-free operation and improve efficiency. The PC-CLOVER system comes with everything you need to add CLOVER to your PC.

CLOVER AUTO-ARQ NORMAL TXT T 58 R 2 DIS WORD R001 T001 21:04

File Name: DATA.WKS	Byte Count: 10476
File Size: 12160	Xfer Time: 00:01:33
Compressed: 5200	Xfer Rate: 136
Protocol: PKLIB	% Complete: ██████████

HAL PC-CLOVER Press ESC Key To Stop Binary File Transfer

Rev up your HF data with the PC-CLOVER system from HAL. CLOVER is the revolutionary modulation technique designed for the HF band. It uses 4 tones, up to 16 phases and 4 amplitudes to really SCREAM on HF. And, it does all this in a 500 Hz bandwidth.

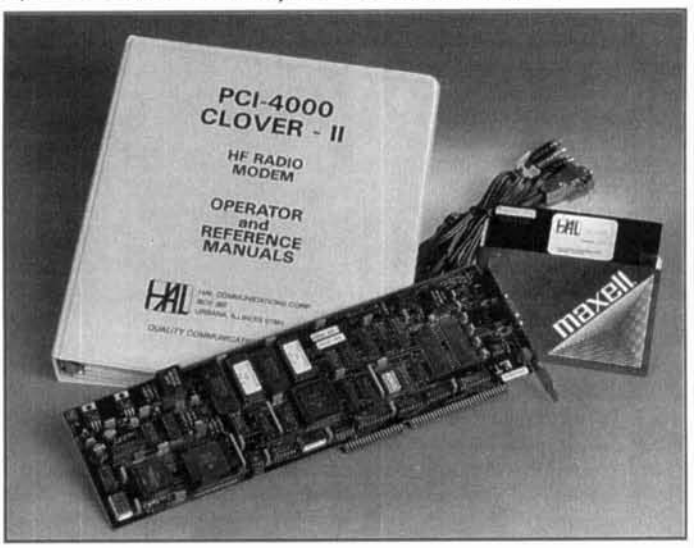
Supercharge your station with CLOVER. All you need is your HF radio, 286 or higher PC, and the PCI-4000/PC-CLOVER system from HAL.

Don't be left in the dust. Call HAL and order your PC-CLOVER system and get ready for SPEED.

Order Factory Direct and Save.  
 Call HAL at (217) 367-7373  
 List Price \$995.  
 Amateur Net \$895.



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



# R-2600

*Bob Wendt of Canyon State Communications*

## "We take it to the mountain."

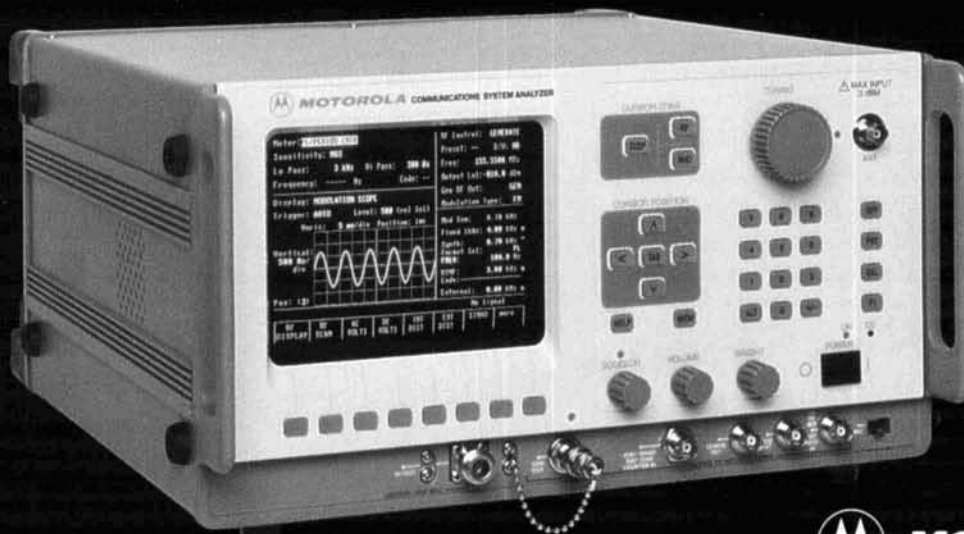


" We bought our Motorola R-2600s for features, ease-of-use, and quality. As an independent two-way service business we are always looking for equipment that will help us get the job done right. Computerized, digital accuracy, with an analog feel. Dependable on the job. I particularly like its software upgrades. Our regular use of the R-2600 is on the bench, but when necessary ( just about every day) we take it to the mountain."



*Bob Wendt*

*Bob Wendt, Manager  
Phoenix Technical Operations  
Canyon State Communications*



For more information call 1-800-235-9590



**MOTOROLA**  
Test Equipment CBU



# 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

# CONNECTING COMPUTERS TO RADIOS

## *A PC interface for the Ramsey 2-meter transceiver*

---

In a previous issue of *Communications Quarterly*,<sup>1</sup> I talked about why I think amateur radio equipment should be more closely integrated with personal computers. A computer provides more power and is more flexible than the control circuitry inside a radio. This article discusses the hardware and software needed to operate a radio from a computer screen/keyboard. I'll illustrate these concepts with an example: using a PC to set the frequency of a Ramsey FX-146 2-meter FM transceiver. Although the project specifically targets IBM-compatible PCs running either DOS or Windows, many of the ideas presented here should be adaptable to other radios, computers, and operating systems.

Interfacing a computer to a radio involves the connectors, cabling, and circuitry to link the necessary control and data signals. Also necessary is the software to get the computer to generate these signals, and to provide a user interface to communicate information between the computer and the operator. The hardware required depends on the radio. Many commercial radios and instruments these days contain a serial data port for computer access. Some of them use signals at "RS-232" voltage levels. For these radios no additional hardware is needed because just about all PCs also have an RS-232 serial port. Many popular transceivers use interface signals with the same functions as those in a PC serial port, but at "TTL" voltage levels. In this case, simple level shifting circuits are needed for the interconnection. A comprehensive discussion about interface circuits for commercial transceivers was presented recently

by Blackburn in *QST*.<sup>2</sup> Software of various quality levels is available from many sources, both commercial and public domain. I'll concentrate more on equipment capable of digital control, but without a data port. Here, additional hardware is required. Typically, this is just a circuit to extract and format the requisite number of data signals from information available in the computer. The Ramsey transceiver described in the following paragraphs needs up to 19 signal I/O lines to determine its frequency.

The techniques I'll present here also apply to homebrew equipment. There is a prevalent notion that the use of digital circuits is best left to commercial designers. Hams are reluctant to incorporate PLLs, DACs, counters, and so on, in circuits they design and build because they assume that getting digital data in or out involves excessive complexity. I will try to show that any concerns about the difficulties involved are misplaced.

I think another reason that the connection between radios and computers has developed slowly is that much of the software presented for amateur radio use has been fairly primitive, particularly regarding the "user friendliness" of these programs. For this project, I will describe a Microsoft Windows™-based graphical user interface (GUI) created with newly available programming language tools—Microsoft's *Visual Basic*™ or *Visual C++*™. These tools can generate professional looking results and don't require advanced software skills to use.

The user interface I suggest for the transceiver follows the model of a "virtual instrument."<sup>3</sup> The idea is to approximate, on the computer

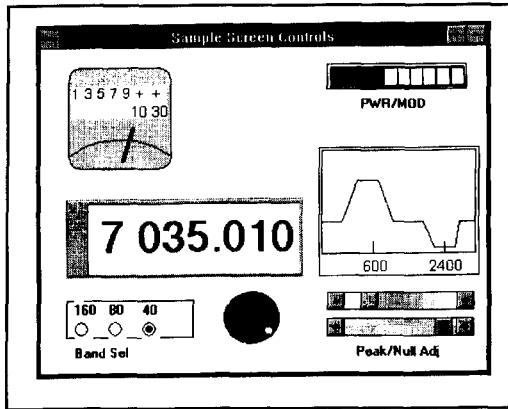


Figure 1. Sample controls.

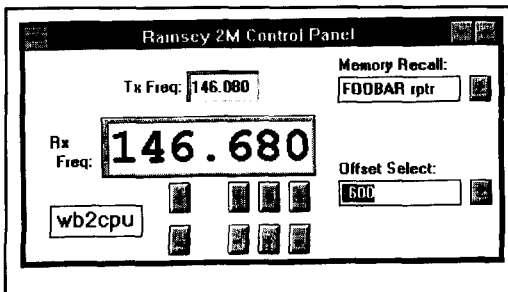


Figure 2. Ramsey 2-meter control panel.

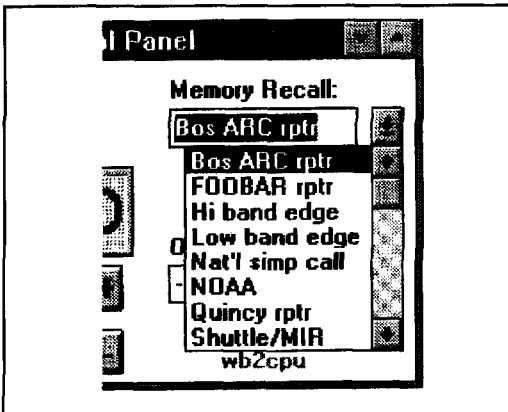


Figure 3. Memory recall box.

screen, the functions normally found on the front panel of an actual piece of equipment. Options are selected using the keyboard and/or the mouse to manipulate “controls” present on the screen. For Windows-based systems, some of these controls, as adapted for representative radio functions, are shown in **Figure 1**. Here you see slider controls used to adjust passband characteristics, button controls used for band selection, and a “knob” to set the frequency. I’ll use the term “virtual radio” for this type of display. There are

\*For Microsoft Windows applications see “The Windows Interface: An Applications Design Guide”, Microsoft Press, 1992.

many more controls available to use. The color and dynamic character of the controls are lost in print. On the actual screen the knob turns, the passband graph changes to reflect its settings and the red S-meter “LEDs” vary with the signal. While some will find it hard to give up a real tuning knob, these new tools can actually be more efficient and convenient.

In creating the graphical user interface for a particular piece of equipment you select the controls that seem most appropriate for their intended functions. To promote consistency, consider the standards and conventions used by other programs in the same operating system environment, so what you, or anyone else, learned using those other programs can be applied to this one.\* I’ll describe later how easy it is to create these displays with *Visual Basic*. Because there are no parts to buy or holes to drill, an initial “prototype” can be modified repeatedly based on the experience gained by using it. Through this process the screen can be optimized to suit your own preferences.

## The FX-146 GUI

For the interface to the Ramsey transceiver, I’ve created the virtual radio panel shown in **Figure 2**. It provides the operator with several ways to change frequency. You can use the “+” and “-” buttons under each digit to increment/decrement that digit. If you click on the arrow to the right of the “Memory Recall” box, it expands as shown in **Figure 3**. This presents a list of frequencies you’ve previously programmed in to select from. Typically, you’ve associated names with these frequencies; for example, the FOOBAR repeater on 146.68. Just select the name from the list with the mouse—no more having to remember the actual frequency or even a memory number. Also associated with the name is a repeater frequency offset, if any. For frequencies not already programmed, the “Offset Select” box, in the lower right of the window and in **Figure 4**, selects the transmit/receive frequency split. You can also enter information directly from the keyboard. If you type “Ctrl-F 145.50,” that frequency will be sent to the transceiver and displayed in the center box. If the offset default is set to “Auto,” the correct transmit offset will be used, following standard conventions. Typing “Ctrl-M” followed by enough of the memory location name to be uniquely distinguishable is equivalent to selecting that name with the mouse. Using the names shown in **Figure 3**, you’ll note that “Ctrl-M B” is enough to pick the Boston Amateur Radio Club, but “Ctrl-M Na” is needed to select the National Simplex Calling frequency. “Ctrl-O” and “Ctrl-S” are used to select the offset default and to



add a new frequency to the "Memory Recall" list, respectively.

## The transceiver

The Ramsey FX-146 is a popular and inexpensive kit that, along with its predecessor, the FTR-146, has been described favorably in several ham publications.<sup>4</sup> Frequency generation is performed using a PLL synthesizer providing the potential for covering any frequency between 140 and 180 MHz. As supplied, however, the user can only select among any twelve specific frequencies using a front panel switch that enables one of the user-programmed diode groups that determine the frequency. Ramsey used this somewhat limited method to hold down the price and give users the flexibility to add whatever additional frequency selection system they choose. Ramsey made it easy to access the signals necessary to program the PLL; this makes the transceiver a good candidate for computer control.

The PLL-based synthesizer section of the transceiver is interesting and instructive in its own right and deserves description. This is particularly true of the method used to handle repeater frequency splits. **Figure 5** shows the synthesizer portion of the transceiver. It's based on the Motorola MC145152 parallel input PLL frequency synthesizer, U6 on the schematic. Part of the 145152 is the equivalent of a 16-bit counter used, along with U3, to divide a 2-meter signal generated by a voltage controlled oscillator (VCO) down to 5 kHz. This is compared to a 5-kHz reference, and the difference produces a signal used to keep the PLL locked on frequency. The 16 inputs to the counter determine its divider ratio. For example, a 145.00-MHz signal would need to be divided by 29000 (decimal), which is 0111 0001 0100 1000 when expressed as a 16-bit binary word, to produce a 5-kHz output. Therefore, this is the value the counter is programmed with to generate 145.00 MHz. The schematic shows several rows of the diode frequency selection matrix that programs the divider value. Connecting a diode in the appropriate position causes a "1" to be programmed in that position of the binary word. Positions without diodes are 0's. A 12-position switch selects one of the rows of diodes.

The situation is slightly more complicated because of two factors. First, for operation at a given frequency, the synthesizer must be able to generate two frequencies: one at the actual transmit frequency and one at the receiver local oscillator (LO) frequency. Because the first IF of the receiver is 21.4 MHz, using the 145.00 MHz example, the synthesizer must also generate 123.60 MHz for the LO when the transceiver

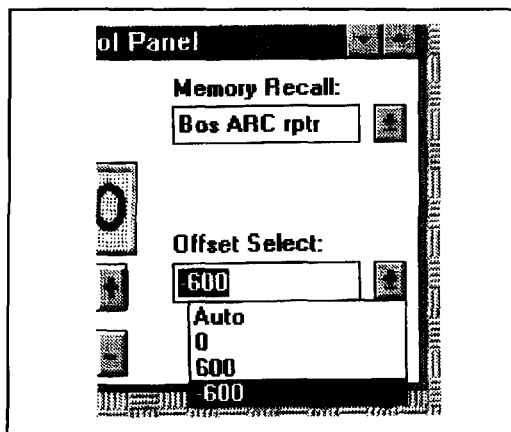


Figure 4. Offset select box.

er is receiving. Second, is the need to accommodate frequency offsets for repeater splits. The FX-146 handles the standard offsets of  $\pm 600$  kHz and one additional user-defined, non-standard frequency split. If you're using a  $-600$  kHz transmitter offset, you still need 123.60 MHz from the LO to receive 145.00 MHz, but now you also need 144.40 MHz for the transmitter. Ramsey's clever solution for keeping track of both the offset required by the receiver LO and the transmit offset is to use a second diode matrix, seen on the left-hand side of the schematic. Four 74283 4-bit adders are used to combine the two matrices. You may have to stare at it a bit to figure out what's going on, but eventually you'll see that the diodes representing the  $-21.4$ -MHz LO offset are switched in on receive, and diodes for plus or minus 600 kHz can be activated for transmitting. In addition to the 16 bits to program the receive frequency, the second matrix requires three more bits to select the correct offset. Thus, a total of 19 bits form the frequency control word necessary to completely determine the frequency of the PLL over its full frequency range.

## Interface hardware

Programming the frequency of the FX-146 from an external computer requires only that the frequency control word be generated by the computer instead of by the diode matrix, and be sent to the transceiver using an appropriate interface technique. Within the transceiver, the signals are applied at the locations I've marked with small circles on the schematic. The transceiver pc board conveniently has a row of holes on 0.1" centers connected to these points to allow you to do this via a pin header and ribbon cable. If you leave one of the matrix rows free of diodes and select it with the front panel switch, the computer can set the frequency. A simple program can be written to calculate the



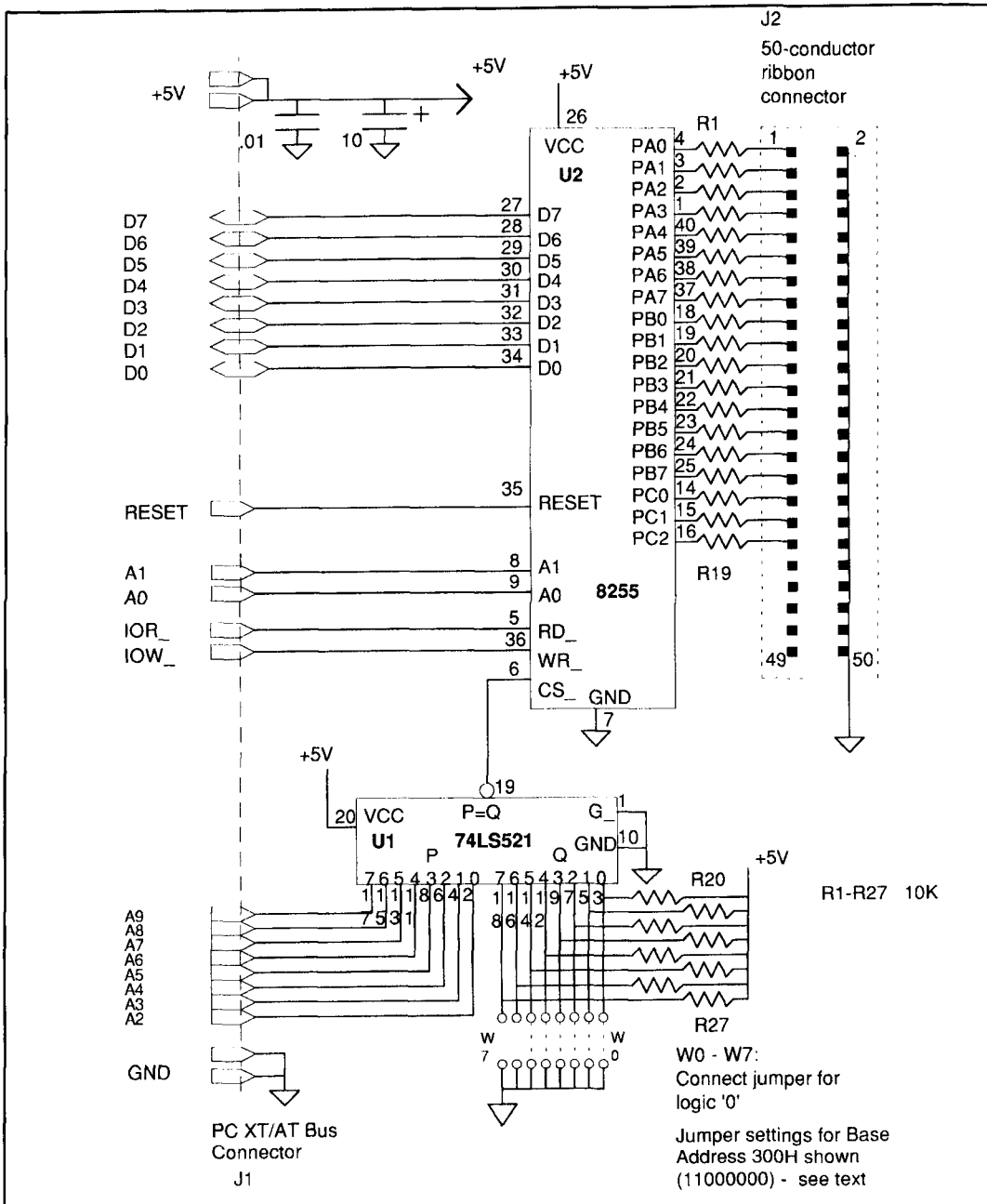


Figure 6. 8255-based bus interface board.

value of the frequency control word from the frequency and offset information supplied by the operator. The question is: What type of data interface should you use to get this word from the computer to the transceiver?

There are three general mechanisms available: using a computer serial data port, using a parallel data port, or taking the data directly from the computer bus. Each is appropriate for different situations. I've summarized the three methods, along with some of their advantages and disadvantages, in the sidebar "Serial, Parallel, Or Bus?." For this project, I'll describe two alternative circuits. One approach

is a board that plugs into a PC bus slot. This avoids having to add circuitry inside the transceiver box and doesn't tie up a port on the computer. The second technique takes data from the computer parallel port. This saves a bus slot and avoids going into the computer.

Suitable bus interface boards are available commercially. I've used the Prairie Digital, Inc. DAS Model 30,\* which costs about \$80. This board not only contains 24 general purpose digital I/O lines, but also an A/D converter and a

\*Prairie Digital, Inc., 846 Seventeenth Street, Industrial Park, Prairie du Sac, Wisconsin, 53578. (608)643-8599.

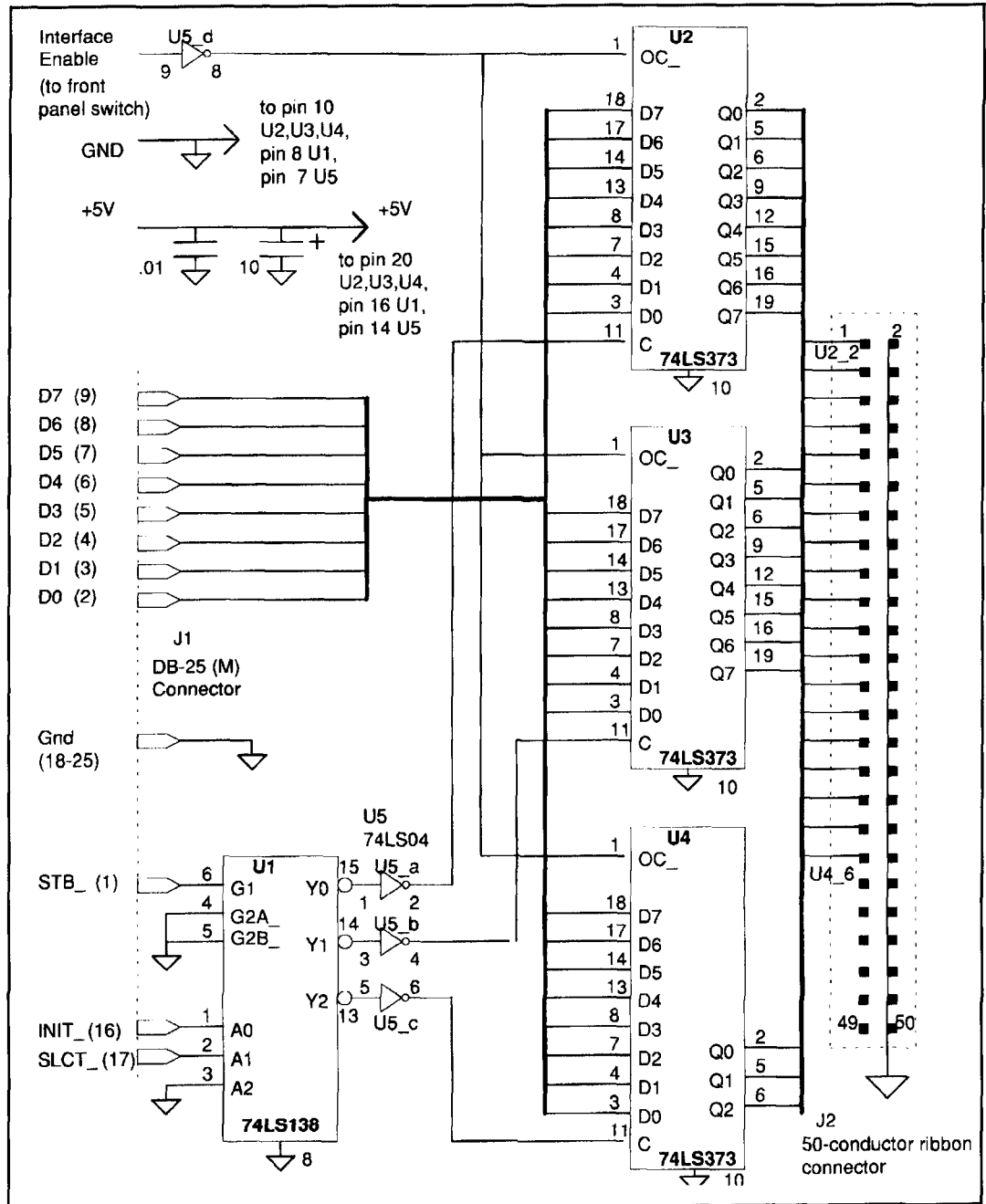


Figure 7. Parallel port interface board.

12-bit counter. Other boards, such as the Computer Boards, Inc. CIO-DIO24\* for about \$60, also seem adequate. If you want to build your own, a similar circuit is given in **Figure 6**.

The bus interface in **Figure 6** consists of an 8255 parallel peripheral interface<sup>5</sup> chip and an address comparator. The 8255 is capable of inputting or outputting 24 lines of binary data to or from the computer. It's organized as three 8-bit read/write data ports: A, B, and C, plus a

write-only control port. The four ports can be accessed as sequential locations by using the chip's two address lines, A1 and A0. A 74521 8-bit address comparator is used to put the 8255 ports at a specific place, the base address (BA), in the machine's I/O address space. Port A is at the base address, port B at an address location one byte higher, port C at BA + 2, and the control port at BA + 3. The base address is determined by eight user-programmed jumper wire positions on the board that connect to eight inputs on the 74521 (the P inputs). Each P input has a pull-up resistor and will be at logic

\*Computer Boards, Inc., 44 Wood Avenue, Mansfield, Massachusetts, 02048, (508)261-1123.

"1" if nothing else is connected to it, so a jumper is only needed to program a "0." The jumper positions represent the most significant bits of the desired 10-bit base address. Eight address lines, A9 to A2, from the PC bus are connected to the 74521 *Q* inputs. If the value of each *P* input matches the value of its corresponding *Q* input, the bus address matches the base address set by the jumpers and an output of the 74521 goes TRUE. This signal is used to activate the 8255 using its chip select (*CS\_*) input. Any of the 1024 I/O locations that is a multiple of four can be used as the base address, as long as it doesn't conflict with an address already in use. Addresses between 300H and 320H (768 to 800 decimal) are often available.\*

To output a byte of data, the computer is programmed to send it to the of the desired port. In BASIC this is done with the "OUT" statement. For example, assume base address 768H has been set with jumpers (11000000) and the 8255 has been set up so that its 24 signal lines are all outputs. When the BASIC statement "OUT 769,170" is executed, the machine's ten lowest address lines will go to "11 0000 0001" (=769 decimal). This matches the jumper settings, so the 8255 will be enabled. Port B will be selected by the lower two bits, "01." Simultaneously, the data bus will go to "1010 1010" (=170) and the bus write signal, *IOW\_*, will go TRUE. This will cause the value of "170" to be written and stored in Port B of the 8255 and sent out over lines *PB7* through *PB0*. The full frequency control word is composed of the outputs of Port A, Port B, and the least significant 3 bits of Port C.

If you want to connect to the computer parallel port instead of taking data from the bus, you can use the circuit in **Figure 7**. I've designed it using discrete TTL logic circuits, instead of the more integrated 8255, although either approach is fine for either circuit. The 8255 is replaced by a 74138 address decoder and three 74373 octal latches. As mentioned in the sidebar, the parallel port is really three 1-byte registers: a data register, a status register, and a control register. These are located at three consecutive addresses in the PC's I/O space. The interface circuit uses only the data and control registers. The parallel port has an address in the machine's I/O space. The port is often at base address 378H and is called "LPT1:." Verify that this is the address used by consulting your computer's instruction manual. The circuit is wired so information written to the data register of the parallel port can be written into one of the 74373s. Two lines from the control register, *INIT\_* (=A0) and *SLCT\_* (=A1) are decoded by the 74138 to select one of the three 74373 ports. Control port signal *STB\_* connects to the common latch enable line, *C*, to latch the eight data bits into the selected latches. The board

can either be mounted in the transceiver or placed in a box between the computer and the radio.

For those using an 8255-based interface board, there's an additional consideration I should mention. If you turn the front panel frequency switch to a position that connects one of the programming diodes to a port pin when the interface is attached to the transceiver, you could destroy the 8255. If that bit is programmed to be LOW, there is a low resistance path from +5 volts through the diode and a port driver output transistor to ground. The resulting excessive current could damage the port/parts. The series resistors shown on the 8255 outputs in the schematic (**Figure 6**) will prevent this from happening, but the problem still exists for commercial I/O boards. Even with the limiting resistors, if the interface board is connected, the transceiver can only be controlled by the computer; none of the other switch positions will work correctly. To restore this function, the interface must be disabled by setting the 8255 port pins to a high impedance state. I'll describe software to do this later.

For interface boards built using separate TTL ICs, the problems described in the previous paragraph can be bypassed easily. The 74373s include an output enable (*OE\_*) pin, pin 1. Because the Ramsey frequency select switch enables a diode row by connecting +5 volts to it, you can wire the switch so it also enables the interface—but only when the appropriate switch position (corresponding to a diode row intended for use with this interface) is selected. The "interface enable" input shown in **Figure 7** is used for this. If you want to permanently enable the interface, either tie this pin HIGH, or, ground the *OE\_* pins and omit the inverter (IC\_5d).

Whether you choose the bus or parallel interface, test it independently of the transceiver to make sure the computer can control the output lines, before complicating things by connecting the radio. The simplest way to send data to a location in I/O address space is by using a BASIC interpreter in its "immediate" mode. In immediate mode, commands are executed as soon as the line is typed in, without having to run a complete program. Recent versions of MS-DOS have included the program QBASIC. The following example uses QBASIC commands to send the three bytes "01010101," "10101010," and "00000000" to ports A, B, and C of the interface board, respectively.

Enter QBASIC and activate the lower portion of the screen with the mouse to enter immediate mode.

---

\*The PC environment has been called a "nonstandard standard." Many conventions, like port address assignments, are adhered to by some, but not all, manufacturers.

For a bus interface with the 8255 at a base address of 772 (=304H) type:

out 775,128	(write 10000000 to control port to set all ports as outputs)
out 772,85	(write 01010101 to Port A)
out 773,170	(write 10101010 to Port B)
out 774,0	(write 00000000 to Port C)

For a parallel printer port interface at address 888 (=378H) type:

out 890,0	(disable latch inputs using control port pin <i>STB_</i> )
out 888,85	(set up data for Port A: 01010101)
out 890,1	(latch into A)
out 890,0	(disable latch)
out 888,170	(set up data for Port B: 10101010)
out 890,3	(latch into B)
out 890,2	(disable latch)
out 888,0	(set data for Port C: 00000000)
out 890,5	(latch into C)
out 890,4	(disable latch)

You can verify correct operation by looking at frequency control signals at the appropriate 4 port pins of the 8255 or 74373s, or, at the transceiver end of the interface cable with a scope or voltmeter. The pins you set to "1" should be greater than 3 volts (8255 pins 38, 40, 2, and 4 in port A for the example above). The pins set to "0" should be less than 1 volt. You may want to build a test circuit with LEDs to confirm the output of a port visually.

To use the interface board to control the transceiver, you could just expand these test instructions into a standard BASIC program. With just a little more effort, you could create the much more powerful program I'll describe.

## A graphical user interface?

One of the major trends in computers in recent years, and one of the reasons personal computers have become so widespread, is the use of graphical rather than command-line program interfaces. The graphical user interface (GUI) first became popular on the Apple Macintosh and is now available on almost all machines. Graphical interfaces receive user input from a pointing device, such as a mouse, along with the keyboard, and present information via visually oriented screen elements like menus and icons.

GUIs are usually easier to learn and use. You can appreciate this if you've ever started up a non-graphical program that you've never used

before and found yourself facing a nearly blank screen, containing perhaps some cryptic instructions, with no idea what to do or how to exit the program. GUI-based programs try to ensure this doesn't happen. Within an environment like Windows, OS/2, or the Macintosh System 7 operating systems, there is a high degree of consistency among different application programs. There are standard mechanisms for doing things such as opening files, scrolling through text, selecting options or getting help. You should only have to learn to do these things once. GUIs are more intuitive to use because they are based on models that correspond to everyday activities. A commonly used metaphor is the "office desktop," with pictorial representations of checkbook registers to fill out, "trash cans" for disposing of data, or file cabinets to "open." Granted, these may seem cartoonish, but they are effective because they trigger behavior that has already been learned in a noncomputer context.

Similarly, GUIs are often more efficient because they have a higher information bandwidth than command-line only interfaces. There are more "channels" of communications between the computer and the user: pictures, controls, menus, the keyboard, the mouse, etc., and the one appropriate for the information to be conveyed can be used. For instance, keyboard entry is usually the most efficient way to input text, but a mouse is a better way to draw a line. There are fanatics on both sides; some want to do everything from the keyboard and some want to do everything using the mouse, but a compromise that gives the user several options is probably optimal.

For radio-related use, a better paradigm than the office desktop is the "virtual instrument/virtual radio," that I described earlier. By recreating the look of a familiar piece of equipment there should be almost no "learning curve." The major objections to GUIs have been that they are harder to create, have tended to run too slowly, and need more powerful computers to run them. However, these obstacles are quickly disappearing.

## Visual programming tools

Until recently, the tools available to create graphical interfaces were limited. Software packages like National Instrument's LabView™, described in **Reference 3**, produce elegant looking results and are easy to use. Unfortunately, they typically cost as much as an average HF transceiver and are aimed at VXI or IEEE bus interfaces found mostly in laboratory or industrial instrumentation. The only other alternative previously available, using standard language compiler packages, requires a low-

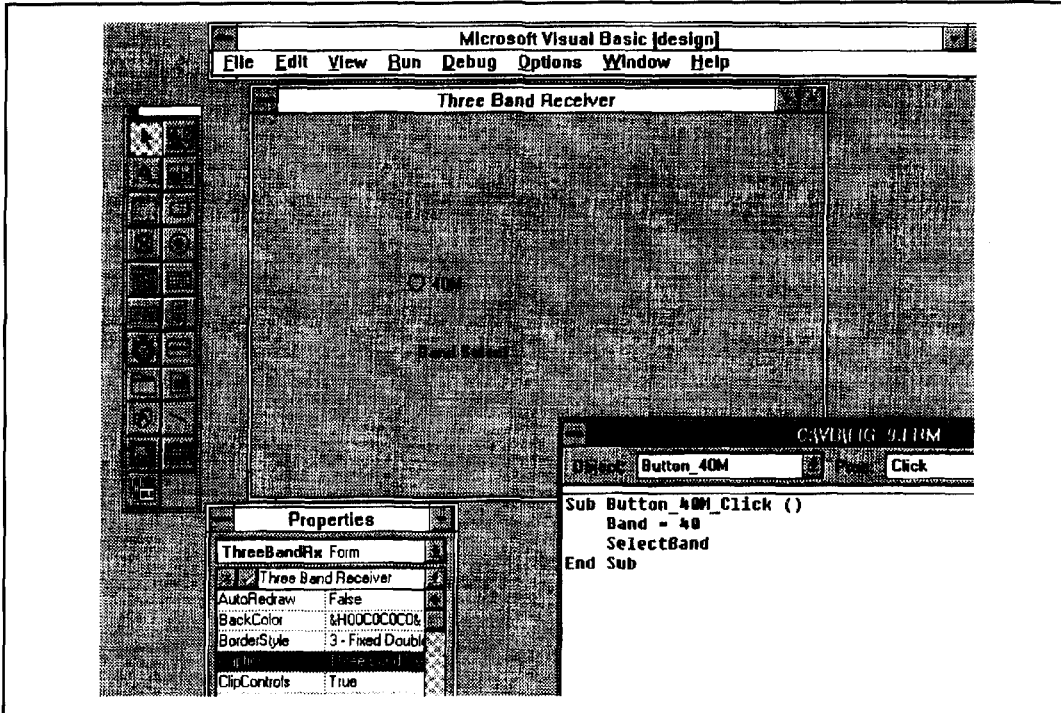


Figure 8. Visual Basic design environment.

level knowledge of the operating system. Fortunately, new products have appeared that make things much easier. The most popular of these is a family of tools from Microsoft: *Visual Basic for DOS*, *Visual Basic for Windows*, and *Visual C++*. The two *Visual Basic* products are almost identical in function. The output from the Windows version looks better because it's able to take advantage of Windows' higher screen resolution. *Visual C++* is more powerful, but harder to use. By taking advantage of "competitive upgrade" offers that are fairly easy to qualify for, each package can be purchased for a street price of under \$100. The description of the process for creating the Ramsey interface that follows uses *Visual Basic for Windows*, but largely applies to *Visual Basic for DOS*, and, to a lesser extent, *Visual C++*. However, I have only implemented and tested the *Visual Basic for Windows* version.

## Programming in Visual Basic

Using traditional programming methodology, called procedural programming, the coding to handle the operator inputs necessary to support GUIs is usually messy. One of the strengths of *Visual Basic* is that it provides an intuitive and powerful means for creating program code to respond to these user inputs. Conventional programming is based on using algorithms to manipulate data. You define variables that represent pieces of information and write formulas

to process this data. *Visual Basic* employs an "event-driven" programming model. Two elements are added to the data/algorithms paradigm: forms and events. A "form" is merely the collection of the visual components of the window seen by the user. The form contains the buttons, list boxes, menus, scroll bars, etc., mentioned earlier. The characteristics of each control, such as its type font, color, or border are defined using its "Properties Box." Even if you've never programmed before, it should only take an hour or so to master the process and create a rough version of the form you want. One or more "event" is also associated with each control or menu item. Examples of events are: "Click" (pressing and releasing the primary mouse button while the mouse cursor is on the control), "DoubleClick" (as the name implies), or "MouseDown" (holding down the primary mouse button over a control). *Visual Basic* provides a very easy-to-use mechanism for connecting an event to the specific section in the program that you want to execute when the event happens.

All programming languages have ways to break up the program code into modules. These units are variously called subroutines, functions, or procedures, with minor differences among them, depending on the language. In addition *Visual Basic* contains a specific type of procedure called an "event procedure". An event procedure is a program module that is automatically executed when its corresponding event takes place on its associated control. For

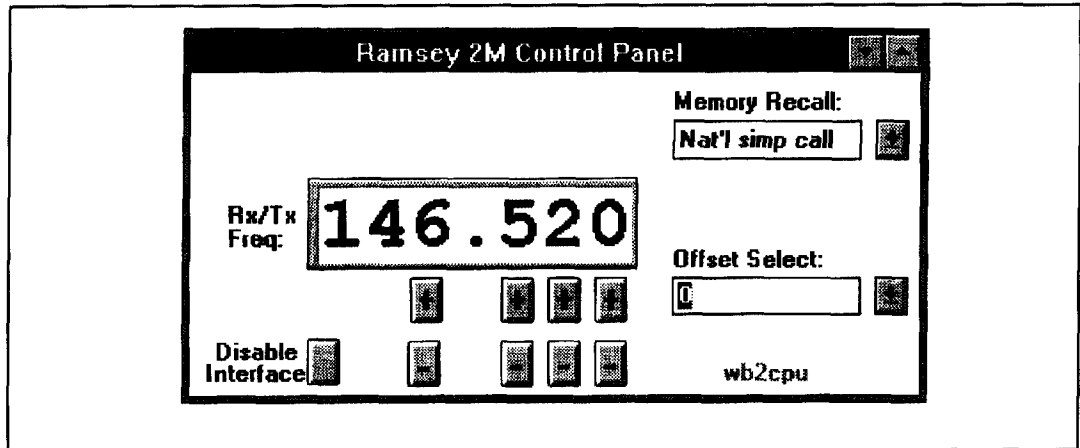


Figure 9. Control Panel with Disable Button and Zero Offset.

example, if your form has a button switch called "Preamp\_On" there would be an event procedure "Preamp\_On\_Click" that would be activated whenever you mouse clicked on that control on the form. There would also be an event procedure for every other event defined for that control. The code portion of a *Visual Basic* program consists of the event procedures for all the controls plus as many general (non-event) procedures as you want to add. *Visual Basic* sets up templates for all the event procedures. A template is an outline of the procedure containing statements that define the start and end of the procedure. All you have to do is supply whatever additional BASIC code you want to run when the event occurs.

Figure 8 shows the *Visual Basic* design environment. The largest box is the form that's being created. I've given it the title "Three-Band Receiver." So far, I've placed a button control named "40M" and a label with the caption "Band Select" in it. To the left of the form you'll find "Tool Bar." Each of the boxes in the Tool Bar represents different controls that can be added to the form. Scanning down from the upper right box in the Tool Bar, the first five boxes are for the controls: "Picture Box," "Text Box," "Command Button," "Option Button," and "List Box." You use the mouse to select the box corresponding to the control you want and drag it onto the form. Once the desired control is placed in roughly the correct location on the form it can be adjusted with the Properties Box, part of which is shown in the lower left of the figure. Properties are altered by editing the line in this box for the characteristic you would like to change. The illustration shows the "Caption" property being modified. Finally, to the lower right of the whole screen is the Code Window. Any procedure in the program can be viewed and edited here by selecting that procedure using the control at the top

of this box. *Visual Basic* defined much of the procedure "Button\_40M\_Click" for me. The only things I've added are an assignment statement to set the global variable "Band" to "40" and a subroutine call to the general procedure "SelectBand." Obviously, I can't cover all the information given in the *Visual Basic* reference manuals, but I will try to at least outline the steps that are involved in creating the Ramsey transceiver interface.

## Designing the Ramsey GUI

The first step in creating the interface program is designing the form. I've chosen to deviate in at least one respect from what might typically be found on a physical piece of equipment. One often-used method of frequency selection involves the use of a pair of buttons—one to increase the frequency and one to decrease it—together with a switch to select the frequency step. For a VHF radio these steps might be 1 kHz, 10 kHz, 100 kHz, and 1 MHz. The system I've chosen to implement has a pair of buttons (up and down) for each of the step sizes. After trying both approaches on my virtual radio panel, I've found I can change the frequency faster this way. I used small "Command Button" controls for these eight buttons. The buttons are named with the step size (in kHz) prefixed with either "Plus" or "Minus;" e.g., "MinusTens." Because the FX-146 has a minimum 5-kHz frequency resolution, the right-most buttons have this step size too. "ComboBox" controls are used for the Memory Select and Offset Select boxes. Both frequency display boxes are "Label" controls. If there is a non-zero offset, the larger one shows the receive frequency and the smaller one shows the transmit frequency. If the offset is zero, there's no need for both displays, so I



make the smaller box disappear (property "Visible = False") and the remaining display shows the common receive/transmit frequency as seen in **Figure 9**. The whole virtual radio panel takes up only about an eighth of an SVGA screen.

Program data variables are either "global" (available to all program modules) or "local" (available only within the program unit in which they are defined). Three significant global variables are: "RxFreq," "Offset," and "MemIndex." "RxFreq" and "Offset" store the receive frequency (in MHz) and transmit offset frequency (in kHz) as their names imply. "MemIndex" is used to store a number corresponding to the user-programmed memory assignment currently selected. "MemIndex" is given the value of "-1" if none of the preassigned channels is currently chosen. The data stored for each user-defined frequency consists of the frequency, the offset, and the name given to the frequency. These are stored in three arrays: "MemFreqs," "MemOffsets," and "MemNames," respectively, with a common access index, "MemIndex." The arrays are local to a procedure called "MemStore."

In the event procedure associated with the "Click" event for each of the frequency step buttons, you increment/decrement the variable "RxFreq" by the amount associated with that button—for example, "RxFreq = Rxfreq + .1" for the "PlusHundreds\_Click" procedure. These event procedures, and anything else that changes the frequency, call the procedure UpdateFrequency. This procedure centralizes many of the support functions required whenever the frequency changes, such as updating the frequency display(s) and sending the encoded frequency control word to the transceiver. The parallel port interface version of the procedure "UpdateFrequency" is shown in **Figure 10** and illustrates what a typical *Visual Basic* module looks like.

When testing the interface board, you used the BASIC "OUT" statement to transmit data to the transceiver. Near the bottom of "UpdateFrequency" (**Figure 10**) you see the section of the transceiver interface code that performs a similar output function. Instead of an OUT statement, there is a subroutine called "port\_out." The "port\_out" subroutine is used because, at least in the current version of *Visual Basic for Windows* (v2.0), there is no OUT function in the language. The only way to send data to a location in the I/O address space is to create a routine to do this using another programming language and call it as a subroutine from the Visual Basic program. The mechanism by which external modules can be invoked by a *Visual Basic for Windows* pro-

gram requires placing them in "Dynamic Link Libraries" (DLLs). If you have access to a Windows C or PASCAL compiler you may be able to create the necessary DLL routine by encapsulating whatever system call is provided to perform this I/O operation. If not, I'll provide it to anyone who needs it—either by posting it to an accessible BBS, or, by mail. OUT is included in *Visual Basic for DOS* so there's no problem in that environment.

Several other loose ends remain. A method is needed for the computer to disable the interface hardware for boards using the 8255 IC. As mentioned, the port pins interfere with normal (noncomputer controlled) use of the transceiver if they are left on. An additional button, labeled "Disable Interface," shown in the lower left of the form (**Figure 9**), can be added. The associated event procedure "Disable\_Click" toggles the interface between enabled and disabled states. Disabling simply consists of sending a code to the 8255 that makes all the port pins inputs rather than outputs.

For Windows (8255-based bus interface board):

```
*assume I/O bus address of 768 (=300H)
  port_out(768, 155)
*155(=9BH) is the code to make all 8255
  pins inputs
```

For DOS, substitute "OUT" for "port\_out."

This guarantees the port pins will be in a high impedance state and not conflict with anything they are connected to. When disabled, I change the color of the frequency readout from

Frequency	Divider	Binary
140.000	28000	0110 1101 0110 0000
142.000	28400	0110 1110 1111 0000
143.000	28600	0110 1111 1011 1000
<b>143.360</b>	<b>28672</b>	<b>0111 0000 0000 0000</b>
144.000	28800	0111 0000 1000 0000
145.000	29000	0111 0001 0100 1000
146.000	29200	0111 0010 0001 0000
147.000	29400	0111 0010 1101 1000
148.000	29600	0111 0011 1010 0000
<b>148.475</b>	<b>29695</b>	<b>0111 0011 1111 1111</b>
149.000	29800	0111 0100 0110 1000
154.000	30800	0111 1000 0101 0000
162.000	32400	0111 1110 1001 0000
<b>163.835</b>	<b>32767</b>	<b>0111 1111 1111 1111</b>
166.000	33200	1000 0001 1011 0000

Table 1. PLL divider ratio.

```

Sub UpdateFreq ()
Dim value As Integer
Dim temp As Integer
Dim CompareFreq As Double

' H. Cahn, WB2CPU; 6-Jul-1993
'
' This version for 13-bit frequency control word
' and parallel printer port I/O ("LPT:" at 378H = 888 decimal)
'
' First, calculate frequency offset if present;
' ...use offset from memory if a memory item is selected;
' else, if 'auto' selected, calculate it,
' else, use value from offset select box
If MemIndex >= 0 Then
    Offset = MemOffsets(MemIndex)
    OffsetSel.Text = MemOffsets(MemIndex)
Elseif (OffsetSel.Text = "Auto") Then
    EvalAuto
Else
    Offset = DefOffset
    OffsetSel.Text = DefOffset
End If

' Figure out Tx value and setup to display both frequencies
FreqDisp.Caption = Format$(Rx_Freq, "###.000")
Tx_Freq = Rx_Freq + (Offset / 1000)
TxFreq.Caption = Format$(Tx_Freq, "###.000")

' Show both Tx and Rx displays unless told otherwise
TxFreq_label.Caption = "Tx Freq:"
Rx_label.Caption = "Rx"
TxFreq.Visible = True

If Offset = 0 Then
' Hide Tx offset display
TxFreq_label.Caption = ""
Rx_label.Caption = "Rx/Tx"
TxFreq.Visible = False
End If

' Send it to xcvr here...
' 13 bits, 10 bits for main frequency, 3 bits for offset
' ... 1st, lower byte (low 8 bits of frequency)

CompareFreq = Rx_Freq / 5000
value = (CompareFreq * 1000000) Mod 256
' set up data
temp = port_out(888, value)
' select LS register
temp = port_out(890, 1)
' latch it in
temp = port_out(890, 0)
' disable latch
temp = port_out(890, 1)

' ... then, upper byte of frequency code.
' Upper 6 bits are hard-wired in xcvr to "011100"
' so mask out all but lower 2 bits.
value = ((CompareFreq * 1000000) \ 256) And 3
' Compute offset code to add; set only 1 of the 3 bits
' for 600, -600, or simplex, respectively
If Offset = 600 Then
    value = value + 4
Elseif (Offset = -600) Then
    value = value + 8
Else
    value = value + 16
End If

' set up data
temp = port_out(888, value)
' select MS register
temp = port_out(890, 5)
' latch it in
temp = port_out(890, 4)
' disable latch
temp = port_out(890, 5)
End Sub

```

Figure 10. Module: "UpdateFreq."

its normal green to gray by changing its "Color" property.

One of a ham's most important responsibilities is to not interfere with other radio services. All equipment is capable of failures that can cause operation on an unintended frequency: a shorted inductor, an open capacitor, etc. With computer control, though, additional failure modes are possible as a result of programming bugs, interface faults, or control cable faults. Therefore, take precautions to ensure legal operation. An external means of verifying the

transmitter output frequency, such as a counter or a separate receiver, is one way to confirm legal operation. In addition, there is a way of modifying the interface that will almost guarantee that any problems with it will not cause out-of-band operation. **Table 1** shows the PLL divider ratio required to produce the listed frequencies. As explained, this value is just the frequency divided by 5 kHz. The ratio is given expressed both as a decimal number and its 16-bit binary representation.

As you can see, over the entire range of 140

to 163.835 MHz the most significant (left most) 3 bits are always "011." In fact, over a range slightly larger than the 2-meter band, 143.360 to 148.475 MHz, the left most six bits are always "011100." This means, if you confine yourself to the ham band, these six bits can be hard-wired to this value and only 10 of the 16 bits have to be determined by the computer. Instead of selecting a row completely free of diodes in the transceiver, you can program a row with "011100" in the upper six positions for use with the computer interface. If this is done a total of only 13 instead of 19 interface lines are needed to control the transceiver; ten for the main frequency plus three to program the offset. This has two advantages. First, it frees up I/O bits so you may be able to use them for other functions,\* and second, there's now little possibility that any error or failure in the program or interface can take the transceiver out of the frequency range of 143.360 to 148.475 MHz. You can, of course, dedicate two of the twelve transceiver switch positions for interface control, one for the full range and one for two meters only, and only use the former for non-ham receiving. As an additional precaution the virtual radio can be programmed to display the frequency in red, instead of green, when the frequency is out of the band.

## Bells and whistles

The virtual radio interface presented here has several advantages over a physical radio front panel:

- Expensive front panel components, such as knobs, switches, readouts, meters, and associated circuitry, are eliminated.
- The user can configure their equivalents, the controls on the virtual panel, as desired.
- The resulting panel is often easier and faster to use.

Perhaps most impressive, however, is the ability to add totally new features without adding new hardware.

I'd like to speculate on some extensions to the Ramsey transceiver that could be added in this manner. Although I have not attempted them, they should be feasible and may stimulate additional ideas about what can be done. Assume that the transceiver interface was implemented using the Prairie Digital bus interface board mentioned earlier. The first thing you might do is use the built-in A/D converter to report received signal strength. This reading could be fed to a virtual panel signal strength meter that looks like a conventional analog meter or an LED bar graph. The transceiver can now be used as a high-powered, completely configurable scanner. It could cycle through your already defined list of preprogrammed fre-

quencies, or scan a specific range with any desired frequency step, or you could define another preprogrammed list just for scanning. Using the A/D output, you could add a squelch function to determine what channels to stop at. *Visual Basic* contains software timers. Consequently, you can easily program in when the computer is to turn the radio on and off, or tell it how long to stop at each channel. You can incorporate whatever type of control you want for any of these functions. One method for selecting the squelch level might be to click on the desired point on the S-meter. For unattended monitoring of a particular frequency or frequencies, the system could automatically record the signal level into a log file or to a simulated chart recorder graph at periodic intervals. This feature would be useful for tracking stations as part of a propagation study or for monitoring how heavily a frequency is used.

The interface program can exchange data with other programs on the computer. Windows programs communicate with each other through a mechanism called Dynamic Data Exchange (DDE). With this facility, the user interface can be combined with Windows logging programs, database programs, packet control programs, etc., for an integrated operating environment. Entering a repeater's frequency could trigger the display of a list of its active stations or a map of the area. Windows' high resolution display permits several windows of information to be displayed on the screen simultaneously.

I've tried to show how a powerful radio control interface can be made with simple hardware and easy-to-create software. A computer-based programmable interface gives amateurs the ability to modify their equipment that has been lost in recent years with fixed programmed, closed transceiver boxes. I hope this article encourages others to undertake similar projects with this radio or other equipment. Please feel free to contact me with any comments, questions, or descriptions of these projects.

## Acknowledgment

Thanks to Mike Ardai, N11ST, for help with several aspects of this project.

## Conventions used in this article

Three different number systems, decimal, hexadecimal, and binary are used. A hexadecimal number ends with an upper-case "H" and a binary number consists of only 1's and 0's. All other numbers are decimal.

Digital signal names are in italics; a trailing underscore ("\_") indicates that the signal is

---

\* The TTL-based parallel port interference circuit described can be built in with two, instead of three, 74373 octal latches now because fewer than 16 output lines are needed.

active when it is in the LOW state. For example the TTL based signal "RESET\_" will perform the reset function when driven to zero volts.

Keyboard entry with the "Ctrl" key involves pressing two keys simultaneously, for instance, "Ctrl-A." Alphabetic keys are case-insensitive.

## Software available

I will be happy to send interested readers a disk containing the program code, including the required "DLL," necessary to create the user interface described, for a cost of \$5. It requires Microsoft Visual Basic to use. Please specify the PC disk size required (3.5" supplied if not

specified). Contact me at the address given at the beginning of the article.

An interface board kit and control software similar to what is presented here will be offered by Ramsey Electronics, Inc. 793 Canning Parkway, Victor, New York 14564, (716)924-4560.

## REFERENCES

1. Howie Cahn, "Bringing Amateur Radio into the Computer Age," *Communications Quarterly*, Winter 1993, page 94.
2. Wallace R. Blackburn, "Everything You Always Wanted to Know About Hardware for Computer-controlling Modern Radios," *QST*, February 1993, page 37.
3. Bryan Bergeron, "Micro-Computer-Based Instrumentation," *Communications Quarterly*, Winter 1992, page 36.
4. Rick Littlefield, "The Ramsey 2-Meter Transceiver Kit," *73 Amateur Radio Today*, February 1992, page 18.

## Serial, Parallel, or Bus??—Three Data Interface Methods

PCs are equipped with two ways for exchanging electronic data with the outside world, the serial port and the parallel port. They are typically used to connect the computer to printers, modems, etc. A third method for external interfacing involves communicating directly with the computer's internal bus. Information about these three techniques is presented below.

### Serial Port

#### Description:

Serial transmission methods send data one bit at a time. The serial data port needs only three signals for two-way transfers: Transmit Data (TXD), Receive Data (RXD), and ground (GND). Several other signals, like Data Set Ready (DSR) or Clear to Send (CTS) are also provided for additional status and control functions. The data is grouped into characters of five to eight bits and sent using an asynchronous format. A Start bit is sent before a character, and the data line is held low for one clock period. After each character, one or more Stop bits (data line HIGH) are sent. The Start and Stop bits are all that's needed to keep the source and destination circuits synchronized.

The PC Serial Port is usually based on an 8250, or similar IC. The 8250 takes parallel data from the computer bus and puts it into a serial format for transmission, and likewise, converts incoming serial data into parallel form.

The serial port's transmit and receive data lines are at "RS-232" voltage levels (logic "0" = -12 volts, logic "1" = +12 volts) for high noise immunity.

**Connector:** A DB-25 (Male) or DB-9 (M) connector on the computer rear panel. The DB-25 is more common.

**Signals:** The signals on the DB-25 or DB-9 connectors are defined in **Figure 11**.

**Logical Name:** "COM1:", for the first serial port in a system. Three additional serial port names, "COM2:", "COM3:", and "COM4:" are defined, but usually only the first two are typically used.

**Base Address :** 3f8H (COM1:)

**Register Addresses:** 3f8H – 3fCH

#### Advantages:

- Provides the highest noise immunity and allows the longest cable length of the three methods.
- Fewest signal lines required. Serial data can be transmitted as tones over telephone lines or as light using fiber optics.
- Bi-directional data transmission supported.

#### Disadvantages:

- Most complicated interface circuit and support software.
- Slowest transmission rate of the three methods.
- Uses up a serial port.

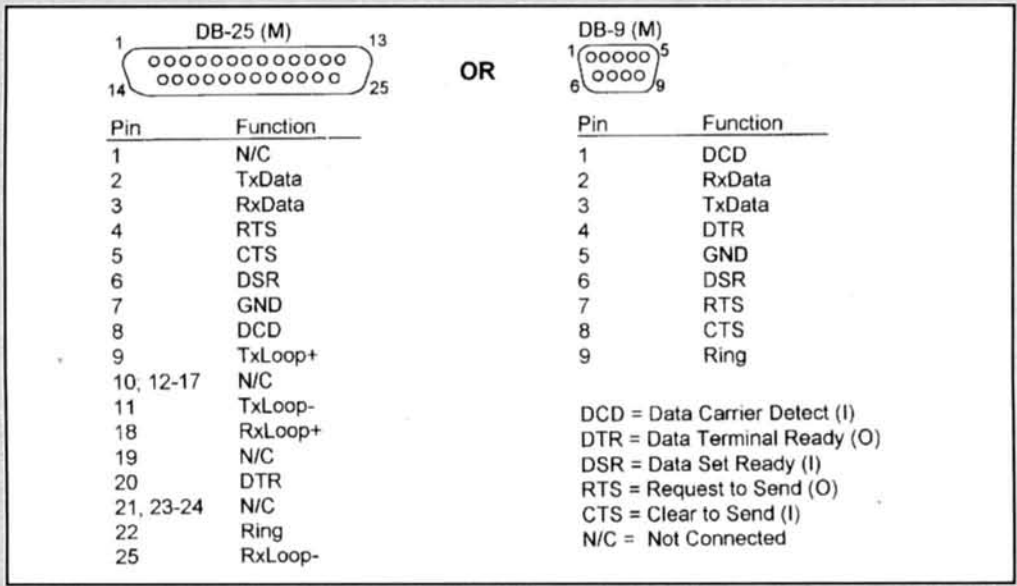


Figure 11. Serial port connector(s) signal definitions.

**Parallel port** (also called Printer or Centronics port)

**Description:**

Eight of the parallel port signals are parallel data lines. When used with a printer, additional port signals are used to implement a handshaking system for data transfer. Transmission consists of data being presented by the computer on the port data line, the data being latched into the printer by the strobe signal (STB<sub>-</sub>), and the printer replying with an acknowledge signal (ACK<sub>-</sub>) to verify receipt. For low-speed, nonprinter interface applications, such as those described here, the handshaking isn't required and the control and status register signals are available for other functions. In the Ramsey interface, they are used to select which register of the interface circuit receives the data.

**Connector:** A DB-25 (Female) connector on the computer rear panel. (When connected to a printer, the printer end of the interface cable typically has a 36-pin "Centronics" type connector.)

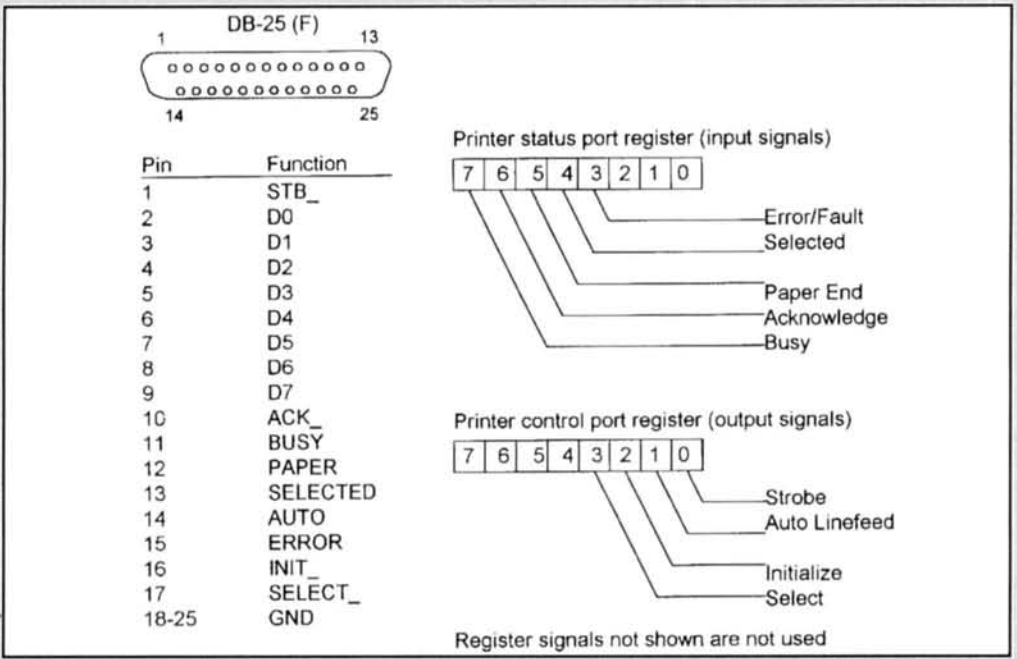


Figure 12. Parallel port connector and register signal definitions.

**Signals/Registers:** The parallel port is organized as three registers: Data, Status, and Control. The register signals are brought out to the DB-25 connector—see **Figure 12**.

Data register, outputs;  
8 lines: D7-D0  
Status register, inputs;  
5 lines:  
ERROR, SELECTED,  
PAPER, ACK\_, BUSY  
Control register, outputs;  
4 lines:  
INIT\_, SLCT\_, AUTO, STB\_

**Logical Name:** "LPT1:" (for 1st parallel port), then "LPT2:", LPT3:"

**Base address:** One of the following: 378H, 278H, 3bcH

**Register Addresses:** Data @base address (BA),  
Status @BA+1,  
Control @BA+2

**Advantages:**

- Higher speed data transfers possible than with a serial port.
- Relatively simple interface circuitry.
- Unlike bus interface, doesn't require a computer bus slot.

**Disadvantages:**

- Uses up a parallel port.
- Cable length limited.
- Poorer noise immunity than serial port.
- Only single direction (output) data transfers are generally possible.

## Bus interface

### Description:

Bus signals are the means used by the PC CPU to communicate with its memory and internal peripherals. They can also be used for interfacing to external devices. A typical data transfer machine cycle is shown in **Figure 13**. To transfer data, the desired source or recipient of the data is selected using the address bus, and the data is presented on the data bus. The control signal MEMR\_, MEMW\_, IOR\_, or IOW\_ is used to initiate the transfer. The data is latched into the destination by the rising edge of one of these signals. The address bus actually represents two buses that share the same physical signal lines—the memory address space and the I/O address space. If MEMR\_ is asserted, data will be read from memory by the CPU. If MEMW\_ is asserted, data will be written to memory by the CPU. Similarly, IOR\_ or IOW\_ will cause a read or write from/to an I/O device.

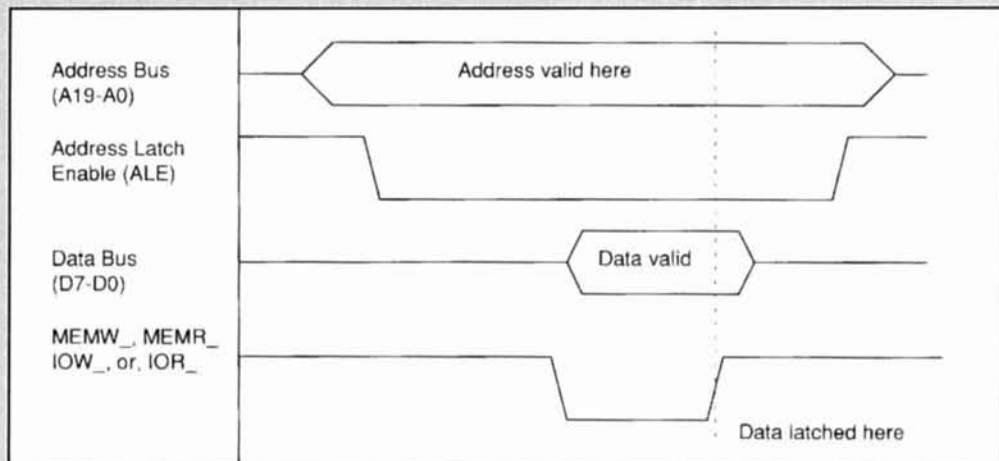


Figure 13. CPU data transfer cycle.

Pin	Signal	Pin	Signal	
A1	IOCHK_	B1	GND	
A2	D7	B2	Reset	
A3	D6	B3	+5V	
A4	D5	B4	IRQ2	
A5	D4	B5	-5V	
A6	D3	B6	DRQ2	
A7	D2	B7	-12V	
A8	D1	B8	OWS_	
A9	D0	B9	+12V	
A10	IOCHRDY	B10	GND	
A11	AEN	B11	MEMW_	
A12	A19	B12	MEMR_	
A13	A18	B13	IOW_	
A14	A17	B14	IOR_	
A15	A16	B15	DACK3_	
A16	A15	B16	DRQ3	
A17	A14	B17	DACK1_	
A18	A13	B18	DRQ1	
A19	A12	B19	DACK0_	
A20	A11	B20	CLK	
A21	A10	B21	IRQ7	
A22	A9	B22	IRQ6	
A23	A8	B23	IRQ5	
A24	A7	B24	IRQ4	
A25	A6	B25	IRQ3	
A26	A5	B26	DACK2_	
A27	A4	B27	T/C	
A28	A3	B28	ALE	
A29	A2	B29	+5V	
A30	A1	B30	OSC	
A31	A0	B31	GND	

Pin Definitions	
Ax	Address line x
AEN	Address Enable
ALE	Address Latch Enable
Dx	Data line x
DACKx	Data Acknowledge x
DRQx	Data Request x
IOCHK_	I/O Channel Check
IOCHRDY	I/O Channel Ready
IOR_	I/O Read
IOW_	I/O Write
IRQx	Interrupt Request x
MEMR_	Memory Read
MEMW_	Memory Write
T/C	Terminal Count

"\_" indicates signal is active LOW

Figure 14. PC/XT bus connector definitions.

**Connector:** Bus interface circuitry is added to the computer on a board with an integral 62-pin connector that plugs into the PC motherboard. There are 31 signals on each side of the double-sided board. The signals are labeled A1-A31 and B1-B31. "AT/ISA," "EISA," and "VL" bus extensions have additional connector(s). See **Figure 14** for bus connector description.

**Signals:** The basic PC/XT 62-pin bus includes an 8-bit data bus, a 20-bit address bus, plus power, control, and status signals. All PCs include this subset of the bus. The bus extensions mentioned above contain additional signals.

- Advantages:**
- Fastest of the three data transfer methods.
  - Simplest interface software.
  - Interface circuit can typically be powered directly from the computer; usually, no additional power supplies are required.

- Disadvantages:**
- Requires a board be installed within the computer and, a free bus slot (not suitable for laptops).
  - Poorest noise immunity—where there are many signal lines to be brought out of the computer, multiple-conductor cable is required; typically unshielded ribbon cable.
  - Only short cable lengths are possible.

# GRAVITY WAVE COMMUNICATIONS

*The ultimate frontier?*

---

**T**ired of QSOs cut short by poor propagation conditions? Frustrated with increasing congestion on the HF bands? Looking for a new challenge and a fresh area of technology to explore? The answer, in theory, could come some day from communications using gravity waves. Okay, so there are still a few practical details to be ironed out. But if present day researchers are successful, you may want to switch communication mediums—in a century or two.

Like electromagnetic waves, gravity waves propagate at the speed of light. Unlike radio waves, gravity waves travel through matter with negligible attenuation—making global communications a theoretical possibility, independent of satellites and the ionosphere. More significantly, gravitational waves exist over a frequency spectrum totally empty of users, free of international regulations, and completely independent of the electromagnetic spectrum. Much like the early radio experimenters and SWLs of long ago, the first gravity wave listeners (GWLs) are currently experimenting with the reception of signals from space using this new medium.

Sound great? Are you ready to build your own gravity wave transceiver? Well, there are a few not so minor complications associated with gravity wave communications that I'll discuss momentarily; but first, let's cover a little historical background.

## Einstein's bright idea

Back in 1873, James Clerk Maxwell showed mathematically that radio waves exist across a wide electromagnetic frequency spectrum. Within fifteen years, Heinrich Hertz had turned

Maxwell's theory into reality by transmitting and receiving electromagnetic energy in his laboratory. The world of communications hasn't been the same since.

Less than thirty years after Hertz first detected radio waves, Albert Einstein's general theory of relativity showed theoretically that gravity waves should exist. When Einstein published his theory in 1915, he showed that a stationary mass contains an amount of energy equal to  $E=mc^2$ . This simple formula states that mass can be converted into energy and energy can likewise be converted into mass. If we put this equation into the form  $E/c^2=m$ , energy itself has a small but calculable mass. One of the implications of Einstein's theory is that because mass and energy are basically the same, then mass and energy can both produce gravity, and the acceleration of mass will generate waves in its gravitation field.<sup>1</sup>

## Evidence of gravity waves

Astronomers first discovered firm physical evidence of gravity waves more than a decade ago. Binary stars are believed to be the largest sources of gravity waves in the universe. These astronomical objects consist of a pair of stars rapidly circling around and around each other with a rotational rate measured in years, days, or even just a few hours.

If a binary star is radiating gravity waves as Einstein predicts, then the pair of stars should gradually be losing energy and spiraling in toward each other. In early 1979, Joseph Taylor and other radio astronomers operating the 1000-foot radio telescope at Arecibo, Puerto Rico reported a significant discovery. Specifically, their discovery involved measure-



ments of the eight-hour orbital period of binary pulsar PSR 1913 + 16. (A binary pulsar consists of a pulsar and a compact star spinning around each other.) Their measurements showed that this binary pulsar's rotation rate was gradually increasing, which meant that the star and pulsar were losing kinetic energy and slipping closer together. More significantly, this decrease in kinetic energy was equal to the amount of energy that the astronomers calculated was being radiated in gravity waves.<sup>2</sup>

Much like the atmospheric noise and static we all endure on the HF bands, the gravitational frequency spectrum has plenty of naturally occurring noise sources. These range from orbiting planets and spinning galaxies to supernovas. Just as with deep space microwave energy, traces of gravity wave radiation from the original Big Bang are estimated to still be present today.<sup>2</sup> **Figure 1** summarizes some of the major sources of gravity waves in our galaxy and where their predominant frequencies are believed to be on the gravitational spectrum. While some sources like binary stars and pulsars are continuously radiating gravity waves, others such as collapsing galaxies and supernovas occur infrequently and, as a result, are believed to release only occasional surges of gravitational radiation. As is apparent from the relatively low frequencies involved, many of these signals are the gravity wave equivalent of ELF radio transmissions with wavelengths of hundreds of thousands of kilometers. And weaker gravity waves being radiated by our spinning solar system cover a band of frequencies between  $10^{-7}$  and  $10^{-10}$  Hz. (Obviously, the gravity wave bit rate possible if we could modulate the speed of the earth's orbit around the sun would be extremely slow!)

Physicists today are working to predict the unique signal characteristics of mass being swallowed by black holes and spinning neutron-star binaries. A collapsing binary is expected to generate a unique "chirp" of gravity waves sweeping upward from 10 to 1000 Hz during the last three minutes before collapse.

Such signals should reveal details about the masses, spin rates, and orbital paths of previously unseen objects.<sup>3</sup>

There is a major catch to communicating with gravity. Although any accelerating object radiates gravity waves, these waves are by their very nature extremely weak and difficult to detect. The electromagnetic force between two charged particles is far more powerful than their gravitational force. In fact, the electric force pushing on two protons is  $10^{36}$  greater than their gravitational attraction.

## Communicating with gravity

A gravity wave communication link can be thought of in the same terms as a radio link.<sup>5</sup> Energy is used by a transmitter to generate a gravity wave that is later intercepted by a gravity wave antenna connected to a receiver.

Whether it is a radio wave or gravity wave, the ratio of the power,  $P_{ro}$ , out of the receiving antenna divided by the power,  $P_{ti}$ , is given by the transmission formula:

$$\frac{P_{ro}}{P_{ti}} = \frac{D_t E_t A_r E_r}{4\pi r^2} \quad (1)$$

where:

- $P_{ro}$  = power out of the receiving antenna, W
- $P_{ti}$  = power into the transmitting antenna, W
- $D_t$  = directivity of the transmitting antenna
- $E_t$  = efficiency of the transmitting antenna (between 0 and 1)
- $A_r$  = aperture of the receiving antenna,  $m^2$
- $E_r$  = efficiency of the receiving antenna (between 0 and 1)
- $r$  = distance between antennas, m

One of the major differences between the radio and gravity wave versions of this equation is the efficiency values of the antennas. In most radio communication links, the ohmic losses in the antennas are small and the anten-

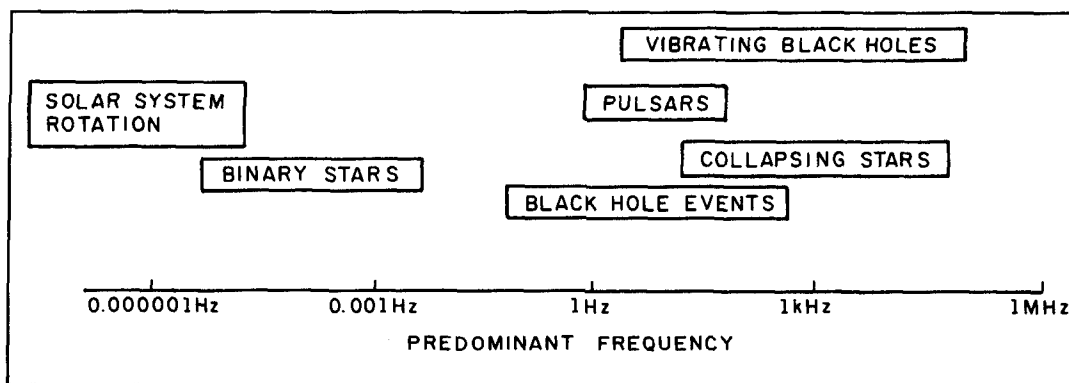


Figure 1. Major signal sources in the gravitational frequency spectrum.<sup>3,4</sup>

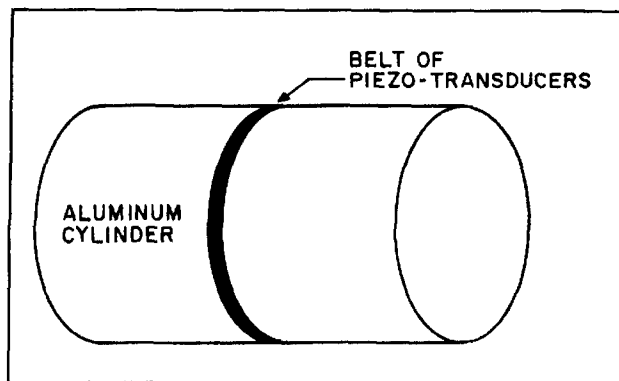


Figure 2. A Weber-Bar gravity wave receiving antenna.

nas are fairly efficient. In contrast, any gravity wave antenna presently conceived of is highly inefficient with an efficiency value close to zero. This antenna efficiency problem is the biggest obstacle preventing gravity-wave communications from occurring today.

Another striking difference between radio and gravity waves has to do with the intrinsic impedance of space. While the impedance of free space for radio waves is 377 ohms, the comparable gravitational impedance is closer to  $10^{-19}$ . This means that for gravity waves, space can be thought of as a transmission line with a characteristic impedance very close to a short circuit. John Kraus, W8JK, points out that because of this incredibly small impedance, we would have an immense mismatch to contend with between the transmitting and receiving ends of a gravity wave communications link.<sup>6\*</sup>

### First you build a receiver

If Hertz was detecting radio waves 25 years after Maxwell's theory of their existence, then why aren't researchers building gravity wave receivers today—over 80 years after Einstein's theoretical proof of their existence? In fact, physicists and astronomers have been listening for gravity waves since the 1950s.

Gravity waves induce tiny vibrations in all sorts of objects. Gravity waves in the kilohertz range may be causing extremely subtle ringing in everything from metal girders to drinking glasses. The first person to seriously pursue the detection of gravity waves was Joseph Weber, a professor at the University of Maryland. His initial approach to gravity wave reception was to construct a gravity wave antenna out of a large aluminum cylinder designed to resonate at a gravitational frequency of about 1 kHz. In order to detect extremely small distortions in the shape of the cylinder, a series of piezoelectric strain transducers was mounted on the surface of the cylinder as shown in Figure 2. Each transducer (containing a highly sensitive piezoelectric crys-

tal) was connected to an amplifier and recorder. (One of Professor Weber's elastic solid receiving antennas has been on display at the Smithsonian Institution in Washington, D.C.)

A more advanced version of the Weber-Bar antenna constructed in 1968 used a pair of 1400 kilogram (3000 pound) gravity wave receivers at Argonne National Laboratory near Chicago and at the University of Maryland. By comparing the signals being measured by the two separate receivers, Weber's team filtered out all the random vibrations in each (due to sound waves, floor vibrations, cosmic rays, and electromagnetic fields) and could look for only those cylinder distortions that occurred simultaneously at both receiving sites. This spatial diversity receiving system could then discriminate between local noise sources and gravity wave signals from space. Between 1968 and 1982, Weber's team detected a number of gravity wave pulses using this receiving system in addition to a receiver pair located at the Universities of Maryland and Rome.<sup>7,8</sup> These initial detections have been highly controversial and may in theory have been caused by ELF or VLF effects instead of gravity waves. However, researchers operating the Maryland/Rome gravity receivers recently reported detecting similar signals that coincided with a major supernova.<sup>9</sup> This supernova signal suggests that the Weber-Bar design does in fact work.

We've already discussed the fact that gravity waves are very weak. There are two ways to increase the sensitivity of a gravity wave receiving antenna. The first is to reduce the thermal noise caused by naturally occurring molecular vibrations. Thermal noise reduction can be achieved by cooling the antenna and by increasing its mass. Gravity antennas based on Weber's metal bar design have used aluminum cylinders weighing up to 6 tons and cooled to temperatures approaching absolute zero—not the kind of antenna an amateur is likely to install in his basement. The best of these can measure vibration-induced deflections as small as one part in  $10^{18}$ . (The reported supernova signal produced a deflection of 2 parts in  $10^{17}$ .)

The second way to boost receiver sensitivity is to make the antenna significantly longer. A new antenna design based on this idea uses interferometric detectors and test masses linked by a laser beam to achieve an antenna up to several kilometers long. Gravity wave-induced vibrations in the test masses result in phase shifts in the laser beam. The effective length of the antenna can be further multiplied several orders of magnitude by reflecting the laser beam hundreds or thousands of times between the test masses. This type of antenna is sensi-

\*Many of the ideas about gravity wave communications discussed in this article were initially described by John Kraus.

tive to signals over a much wider frequency range than Weber's bar antenna and is estimated to have a bandwidth extending from a few hertz to several kilohertz.

A working interferometric gravity antenna prototype tested recently at Caltech had a total length of 40 meters and was sensitive to displacements as small as  $3 \times 10^{-16}$  meters (equal to one-third the diameter of a proton).<sup>10</sup> The United States recently appropriated \$23.5 million to begin work on two 4-kilometer gravity-wave detectors, and France and Italy are drawing up plans for an interferometric gravity antenna of their own.<sup>3,11</sup> The U.S. design should be capable of detecting strain values of one part in  $10^{22}$  and is scheduled for completion during this century.

Russian researchers have even proposed a receiving device capable of detecting gravity waves in the gigahertz range.<sup>6</sup> Their design, called a Braginsky-Mensky toroidal-waveguide cavity detector, is based on a resonant waveguide cavity in the shape of a ring. An extremely stable RF signal generator would introduce a microwave signal into the cavity at its resonant frequency (say 1 GHz). Any small modulation of the waveguide walls by a passing gravity wave with a frequency of 2 GHz would cause a slight phase shift in the frequency of the resonating radio wave. This design still would not be a simple project because it would require superconducting cavity materials cooled to near absolute zero to get the cavity's Q high enough. Even then, a  $1 \text{ W/m}^2$  gravity wave would only cause a phase shift of  $10^{-8}$  Hz.

## Then you build a transmitter

If building a gravity wave receiver sounds daunting, the transmitter is an even bigger challenge. Rotating a conventional Yagi antenna does produce a gravity wave signal—though an incredibly weak one. The key to gravity wave transmitting is that the larger an object's gravitational pull and the more rapid the acceleration, the more intense the resulting gravity waves will be. The amount of gravitational energy radiated by a rotating object is related to its length, mass, rotation rate, and how unevenly its mass is distributed.

Even so, the amount of energy emitted in gravity waves often appears insignificant when compared to other more commonly discussed types of radiation. While the sun itself radiates 400,000 quadrillion megawatts of light, the motion of Jupiter orbiting the sun radiates only about 5.2 kilowatts in gravity waves. The most intense known source of gravity waves in the universe, a binary star called UV Leo, only radiates 0.15 percent as much energy in gravity waves as does our sun in sunlight.<sup>5</sup> Compared

to these galactic gravity wave sources, any gravity wave transmitter we might fabricate on earth would be truly QRP.

Let's suppose you have completed construction of a 10 km long interferometric receiving system and decide to use a rotatable 550-ton steel beam for your transmitting antenna. Let's also assume your steel bar is 20 meters long and you have a motor capable of rotating it at speeds approaching 270 rpm—very close to the speed where the beam will literally explode from the centrifugal force. Even at this rotation rate, your antenna would only be radiating a mere  $10^{-29}$  watts.<sup>6,12,13,14</sup>

What would the radiation efficiency of this monster be? As I mentioned earlier, most ham antennas are fairly effective radiators with efficiencies ranging from, say, 0.5 to 1 (50 to 100 percent). If we assume that the motor requires about 100 kW to bring the steel beam up to speed, then our antenna's efficiency would be approximately:

$$\begin{aligned} \text{Transmitting Efficiency} &= \frac{\text{Gravity Output Power}}{\text{Input Power}} \\ &= \frac{10^{-29}}{10^5} = 10^{-34} \end{aligned} \quad (2)$$

Clearly, this wouldn't be the kind of transmitter you would want to power with solar panels. To make matters worse, a gravity wave amateur (GWA) attempting to pick up your  $10^{-29}$  watt signal would need a receiving antenna capable of detecting displacements no larger than one part in  $10^{-40}$ . (I didn't say gravity-wave communications were going to be easy!)

There is one other complication to be resolved—the bit rate. Because our whirling dervish would have a carrier below 10 Hz, the maximum bit rate possible would be extremely low. There's also the nontrivial question of how to modulate the rotation rate of such an awesome device.

## The bottom line

If theoretical physics says that radio waves and gravity waves both exist, why is radio wave communication practical and gravity wave communication not? Within the past century, efficient methods have been developed for transmitting and intercepting electromagnetic waves—efficient methods that do not yet exist for gravity waves.

People hundreds of years ago were familiar with static electricity, much as we are familiar with the "static" gravitational field of the earth. Yet if Isaac Newton had been presented with Maxwell's equations and asked to produce an electromagnetic signal (i.e., a radio wave), the

only method available to him would have been to put a static charge on a piece of amber or rubber and to rapidly wave the charged material back and forth over his head. Primitive as it may sound, the acceleration of the charged material by the back and forth motion would radiate an incredibly weak ELF signal.

At present, our methods of producing gravity waves are as inefficient and impractical for communication as a charge waving radio transmitter would have been in Newton's day. But even though such a method is currently unworkable, the possibility exists that future breakthroughs in physics will make some form of gravity-wave communications feasible.

Are gravity-wave communications likely to happen any time soon? *No way.* Is the concept of gravity-wave communications a theoretical possibility? *Most definitely.* While gravity-wave communications may not look feasible today, scientists are continually surprised by breakthroughs in areas like superconductivity, antimatter, supergravity, and subatomic physics. New gravitational radiation detection techniques ranging from higher Q torodial cavities and pulsed-chirped generators to quantum-mechanical squeezed-state detectors may help pave the way.<sup>6</sup>

When Henrich Hertz first started transmitting

and receiving radio waves in 1888, the thought of global radio communications must have seemed totally implausible. Is it possible that hams (gravity wave amateurs, that is) of the future may look back on present gravitational radiation research in the same way? ■

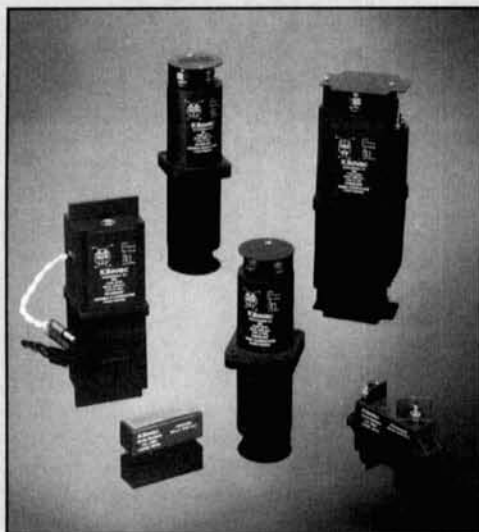
#### REFERENCES

1. D.R. Stinebring, et al., "Cosmic Gravitational Wave Background: Limits from Millisecond Pulsar Timing," *Phys. Rev. Lett.*, Volume 65, July 16, 1990, page 285.
2. J.H. Taylor and J.M. Weisberg, "A New Test of General Relativity: Gravitational Radiation and the Binary Pulsar PSR1913+16," *Astrophysics Journal*, Volume 253, 1982, page 908.
3. Ivars Peterson, "The Last Three Minutes: Computing the Shape of Gravitational Waves to Come," *Science News*, Volume 143, June 26, 1993.
4. J. Hough and R. Drever, "The Hunt for Gravitational Waves," *Observing the Universe*, N. Henbest, editor, August 17, 1978, page 190.
5. W.J. Kaufmann III, *The Cosmic Frontier of General Relativity*, Little, Brown & Company, 1977, page 249.
6. John D. Kraus, "Will Gravity-Wave Communication Be Possible?," *IEEE Antennas and Propagation Magazine*, Volume 33, No. 4, August 1991.
7. J. Weber, "Gravitational Radiation Antennas," unpublished paper.
8. J. Weber, "The Search for Gravitational Radiation," *General Relativity and Gravitation*, Volume 2, A. Held, editor, Plenum Pub., 1980.
9. M. Aglietta, et al., "Analysis of the Data Recorded by the Mont Blanc Neutrino Detector and by the Maryland and Rome Gravitational-Wave Detectors During SN1987A," *Il Nuovo Cimento*, Volume 12C, No. 1, January-February 1989.
10. A.D. Jeffries, et al., "Gravity Wave Observations," *Scientific American*, Volume 256 No. 6, June 1987.
11. R. Ruthen, "Catching the Wave," *Scientific American*, Volume 261, No. 3, March 1992.
12. Michael Disney, *The Hidden Universe*, Macmillan Pubs., 1984, page 183.
13. C.W. Misner, et al., *Gravitation*, Freeman, 1971, page 979.
14. P.C.W. Davies, *The Search for Gravity Waves*, Cambridge, 1980.

## PRODUCT INFORMATION

### New generation of high voltage DC relays and contactors from Kilovac

Kilovac Corporation announced the introduction of a new generation of high voltage (up to 1000 volts DC) relays and contactors for industrial applications including electric



vehicles, light rail transit, back plane power for computers, utility and factory control power, distributed power systems, circuit protection, and transfer switching.

Kilovac's new generation "PD series" relays and contactors are designed to switch up to 1000 volts DC, with current carry ratings of 5A, 10A, 90A, 150A and 250A, and overload capabilities as high as 1500A. All are available from stock, except the 250A model that will be available in production quantities in the fourth quarter of 1993. Published prices in 100 piece quantities range from \$84 each for the 5A relay to \$637 each for the 250A contactor.

For more information on Kilovac's vacuum-sealed technology and new generation "PD series" industrial relays and contactors; or to receive a copy of Kilovac's new 8-page DC relays and contactors product guide, contact Pat McPherson, Vice President Sales and Marketing, at 800-253-4560 or write to Kilovac Corp., P.O. Box 4422, Santa Barbara, CA 93140.

# THE UNI-DIRECTIONAL LONG WIRE ANTENNA

*A look at this useful alternative  
to the Yagi*

An end-fed antenna has pattern distortion; the lobe in the direction of the far end is stronger than the lobe in the direction of the feed end. The effect is most noticeable in antennas several wavelengths long—those known as the long-wire antennas.

I will examine this phenomena in detail, with an eye to determining the amount of directivity available. I'll then show how the directivity can be increased by introducing resistance loading at the end away from the feed point. Antennas with these characteristics are members of the class called traveling-wave antennas. The Beverage is the most common antenna in amateur usage.

Because the antenna analysis program MININEC isn't designed for true end or voltage feed, my analysis is based on current point feed, at one-quarter wavelength from the end. I'll show how this feed technique can be used to give low impedance feed while retaining the broadband capability of the end-fed long wire.

Due to the space needed, the long wire isn't an appropriate antenna for everyone. For those with space, it is a useful alternative to the Yagi—providing good performance from a simple antenna. It's also a useful antenna for portable or emergency use.

## The effect of frequency on antenna radiation patterns

When the frequency of a signal fed to a half-wave dipole is increased above its resonance value, the antenna lobe at right angles to the

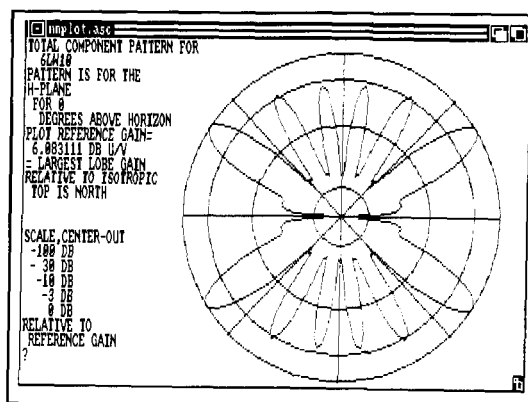


Figure 1. Free-space pattern of a center-fed antenna 3 wavelengths long.

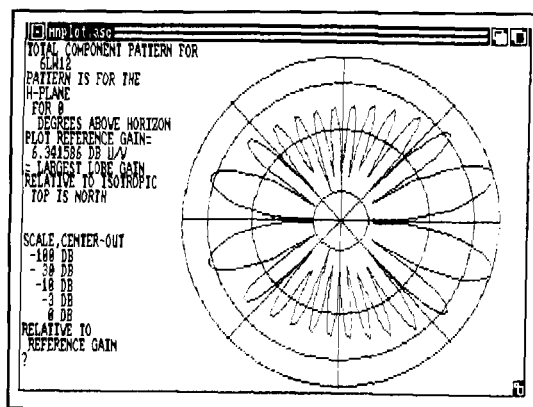


Figure 2. Free-space pattern of an end-current-fed antenna 6 wavelengths long. The feed is  $1/4$ -wavelength from the left end.

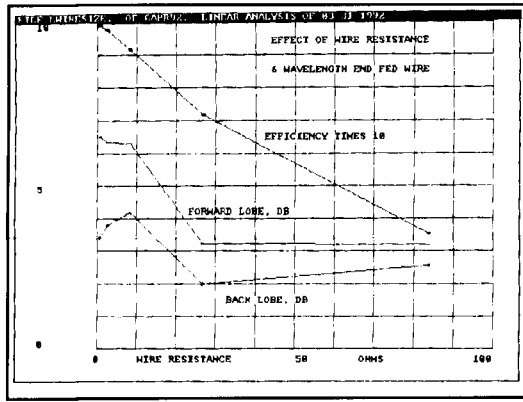


Figure 3. Effect of radiator resistance on a long-wire antenna. Efficiency, forward gain, and F/B ratio all decrease as wire resistance increases. Calculated by MININEC with resistance load at each wire segment.

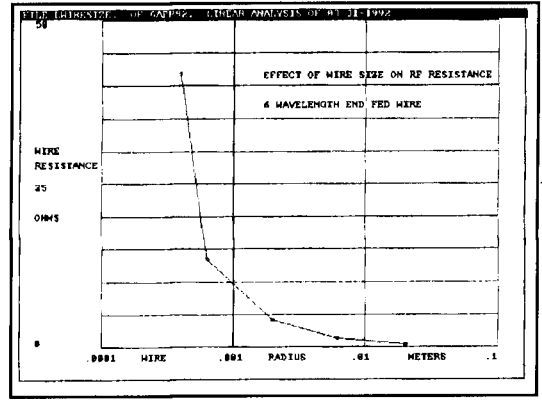


Figure 4. RF resistance of a wire as a function of wire size at 20 MHz. Multiply size in meters by 80,000 for size in mils. Resistance varies as the square root of frequency.

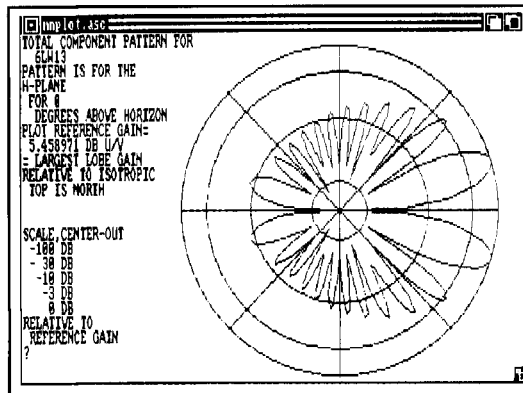


Figure 5. Free space pattern for the antenna of Figure 2, but with a 1000-ohm resistance located  $1/4$  wavelength from the end opposite the feed. Note the improvement in F/B. There is a 0.8 dB loss in gain of the main lobe.

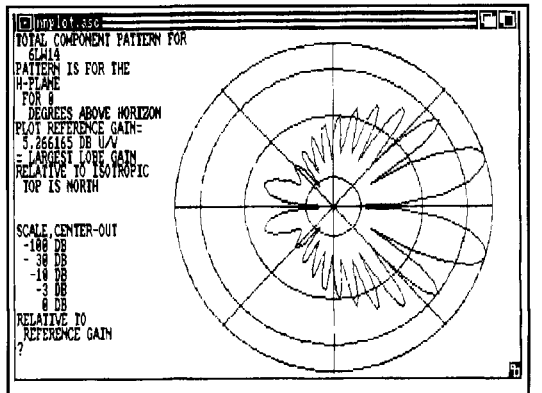


Figure 6. Same as Figure 5 but with 333-ohm resistors at  $1/4$ ,  $1/2$ , and  $3/4$  wavelengths from the end opposite the feed. Note the further improvement in F/B, and the added 0.2 dB loss in gain.

antenna pattern first develops a dimple at the 90-degree point. With an additional increase, the lobe splits into two lobes that are narrower than the original, each at about 45 degrees to the original lobe.

With a further increase in frequency, additional lobes are formed. A step in this process is shown in Figure 1 for a frequency at which the antenna is essentially three wavelengths long. The general rule is that the number of lobes between 0 and 90 degrees is equal to the antenna length in wavelengths, for integer lengths. If there are fractional parts of a wavelength, there is also a lobe at 90 degrees. This is fully formed if the fraction is  $1/2$ , and shows the dimple leading to two lobes for fractions between  $1/2$  and unity.

The pattern created appears to be somewhat different from the construct usually shown in textbooks. These show a line parallel to the 0 to 180 degree axis, tangent to the end lobes—illustrating that the intensity of the minor lobes is equal to gain of the end lobes in the direction

at right angles to the antenna. The difference is due to the presentation, which is usually linear in textbooks, and is logarithmic here.

### Effect of end feed

When the feed is off center, the pattern changes, becoming asymmetric. Figure 2 shows what happens to a 6-wavelength antenna, fed  $1/4$  wavelength from the left end. Note that there are six lobes between 0 and 90 degrees and between 90 and 180 degrees. However, the two halves are different; the lobes away from the feed are the largest.

The reason for the asymmetry lies in the radiation from the antenna. A wave starting from the feed point and traveling towards the far end is losing energy by radiation. As a result, the wave reflected from the far end back to the source is smaller than the exciting wave. The antenna is acting as a combination of a traveling-wave antenna and a normal antenna with a

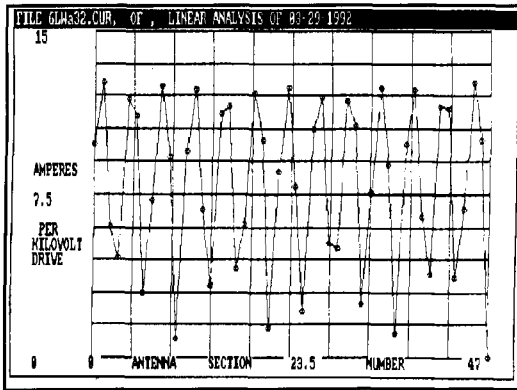


Figure 7. Current distribution of the antenna of Figure 2, with current value shown for each MININEC calculating segment. The actual curve would have more rounded points, as for a sine wave.

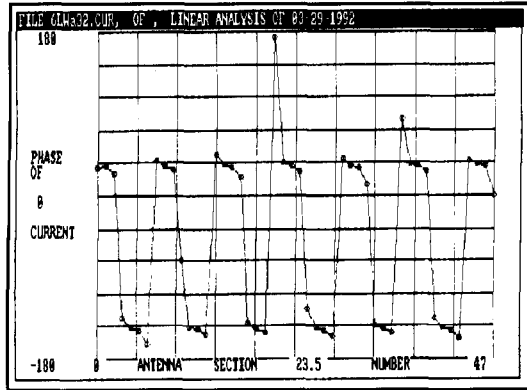


Figure 8. Phase of the current of Figure 7, showing that the standing wave predominates. The apparent spike is a phase of just over 180 degrees at segment 21. Compare to that at segment 44.

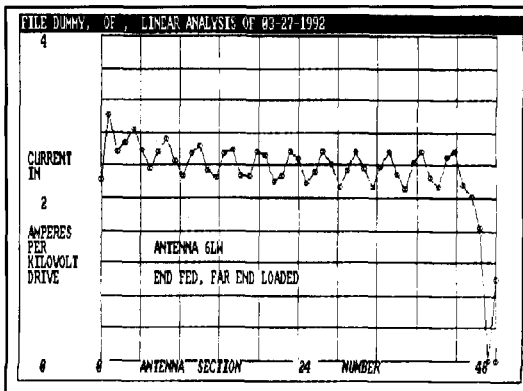


Figure 9. Current distribution for the antenna of Figure 5. The dominant component is now that of a traveling wave progressing from the feed to the right. There is a small standing wave component impressed on this.

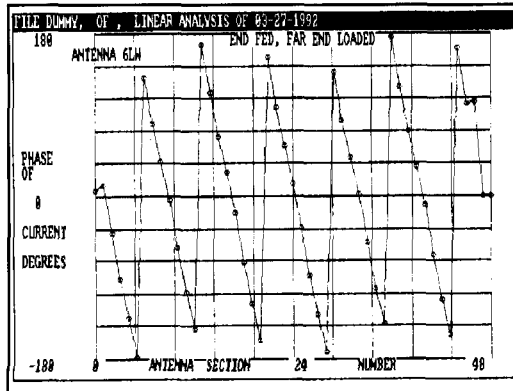


Figure 10. Phase of the current of Figure 9. Primarily, the phase increases steadily along the antenna. The small standing wave component shows in the detail variation.

standing wave present. I will consider the currents later.

This feed location is convenient, as it gives a low-impedance feed point. On the other hand, nominally, it can be used only at one frequency—losing the multiple frequency capability of a true end-fed antenna like the Zepp. It's also convenient from an analytical view. The program used for analysis, MININEC, has no provision for direct voltage feed. This could be modeled by adding a transmission line section to form a Zepp, but at the expense of more complexity and time. I chose the simple approach here.

## Effect of antenna resistance

The above patterns are for ideal conditions, for antenna conductors so large that their resistance can be ignored. When conductor resistance is appreciable, there are additional pattern changes.

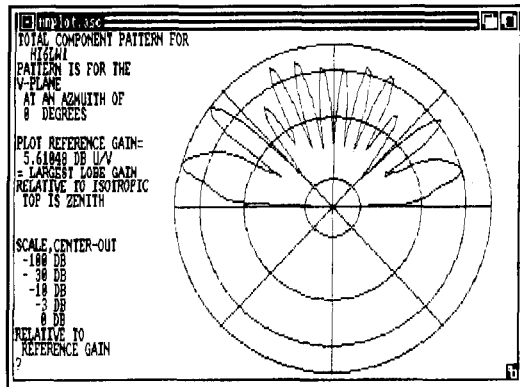


Figure 11. Vertical plane pattern for the antenna of Figure 2, installed relatively high above earth. The azimuth choice makes the feed at the right.

The magnitude of these changes is shown in Figure 3. One direct effect is a decrease in antenna efficiency, due to the current squared loss. Another is fluctuations in lobe structure. The main effect of this change is a reduction in

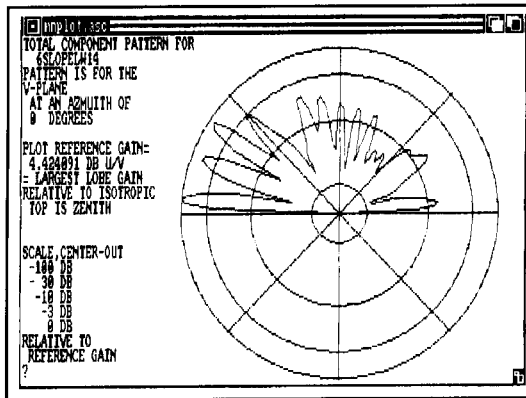


Figure 12. Same as Figure 11, but with the antenna installed as a sloper (see text).

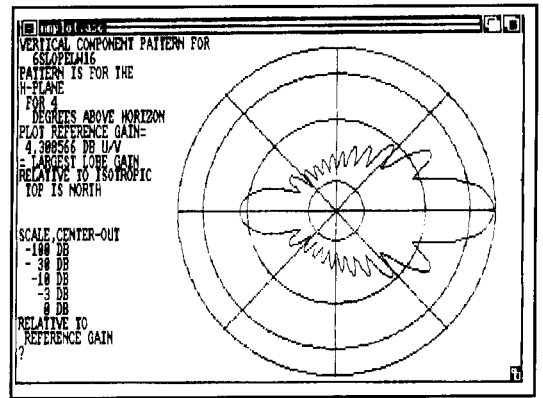


Figure 13. Horizontal plane cut pattern at 4 degrees elevation for the sloper antenna (see Figure 14).

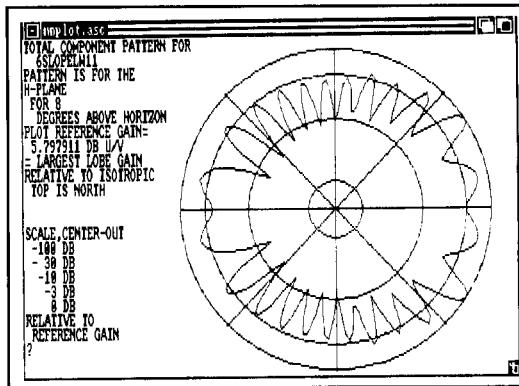


Figure 14. Same as Figure 13, but for the cut 8 degrees above the horizon. Note that maximum gain at this elevation is at 40 degrees to the antenna axis.

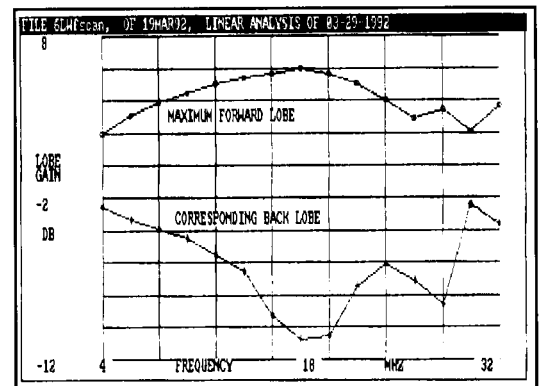


Figure 15. Lobe gain versus frequency for the 6 wavelength antenna at 20 meters shown in Figure 5.

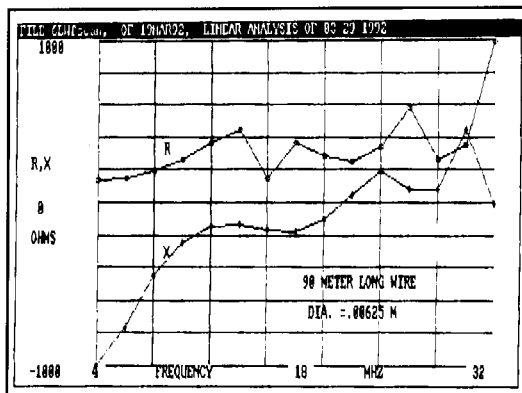


Figure 16. Drive-point resistance and reactance as in Figure 15. The antenna is easily fed with 300-ohm twin-lead or open-wire line.

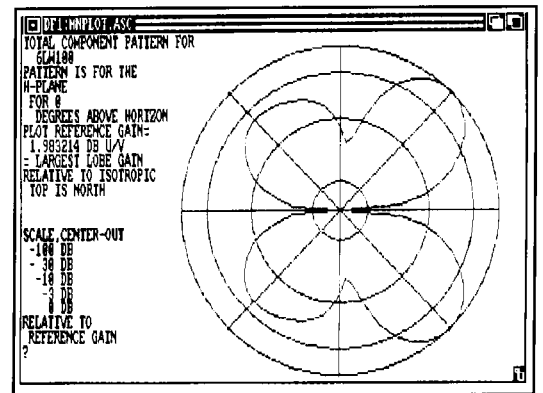


Figure 17. Free-space pattern for the antenna of Figure 5 at 2 MHz.

intensity of the main lobe—the one off the far end of the antenna. At the same time, the F/B ratio degrades. These effects are the reason very fine wire “invisible” antennas don’t always perform well. And, of course, iron and stainless steel wire are poorer performers than copper.

You can get a “feel” for wire size effect by looking at Figure 4. This shows the calculated

RF resistance as a function of wire size, at a frequency of 20 MHz. Multiply the radius in meters by 80,000 to get the wire diameter in mils. For example, no. 12 wire has a radius of 0.001 meters.

### Resistance loading

Instead of using small wire to make the lobe changes, you can obtain superior results by



placing a concentrated resistance at a point where it will affect the wave reflected from the far end of the antenna. A convenient place is the first far-end current maximum, at  $1/4$  wavelength from the end.

**Figure 5** shows the pattern with a 1000-ohm load resistance at this point. The far lobe gain is essentially unchanged, but the lobe at the feed end has decreased, giving a F/B of about 10 dB.

Instead of a single loading resistor, you can use several. **Figure 6** shows the effect of three 333-ohm resistors, one at each high current point from the far end. Gain has decreased slightly, and the F/B has increased further, to about 14 dB.

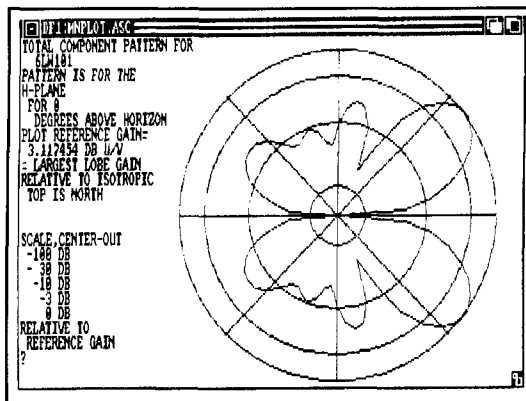
The value of resistance giving best F/B is a function of antenna length and diameter, and also varies with height. The best ratio obtained in many trials is a F/B of 28 dB, with a resistance of 400 ohm, for a free space antenna.

It's not necessary to place the resistance at the antenna. It can be at the end of a transmission line, which allows for placement of the resistance at ground level for convenient adjustment. Such a line can be placed at both ends of the antenna, giving voltage feed at one end and resistance loading at the other. The far end line can be made of resistance wire, canceling the reflected wave as is common in rhombic installations. (I haven't studied this, but it would seem to eliminate the need for resistance adjustment when maximum F/B is desired.)

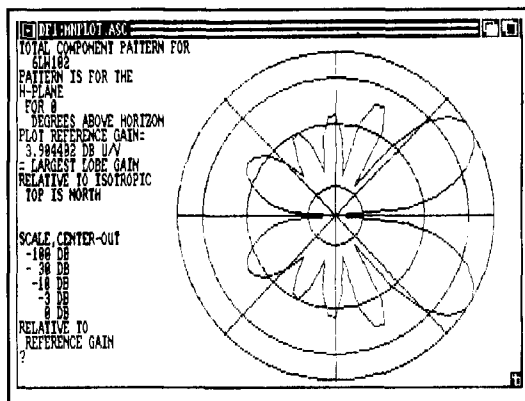
## Current on the antenna

**Figure 7** shows the current along the 6-wavelength end-fed antenna without loading; **Figure 8** shows the phase of the current. There are essentially twelve half-waves of current along the antenna, alternating in phase by 180 degrees, as illustrated in **Figure 8**. Note that the antenna isn't exactly at the 6-wavelength resonant point. This is shown by the fact that the current at the feed point isn't exactly at 0 degrees phase, and by the changes in current magnitude along the antenna. The apparent "spike" in the phase plot comes from the plot routine's interpretation of slightly more than 180-degree phase shift as compared to slightly less.

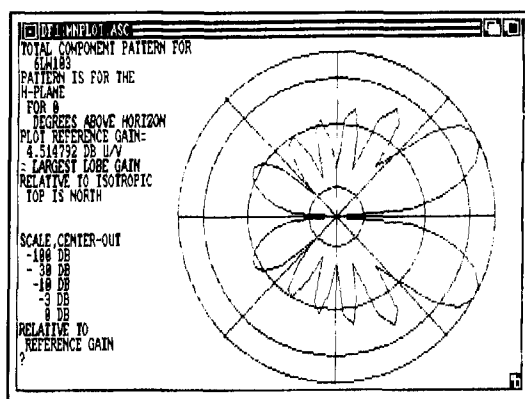
**Figures 9** and **10** show the corresponding plots for the same length antenna with a loading resistance at the far end. Now the current along the antenna is a combination of a major component, the wave traveling down the antenna, and a standing-wave component created by the interaction of the small component reflected from the far end of the wire. The reduction in current along the wire is due to loss by radiation and resistance. As is true in terminated transmission lines, the phase of the current



**Figure 18.** Free-space pattern for the antenna of *Figure 5* at 4 MHz.



**Figure 19.** Free-space pattern for the antenna of *Figure 5* at 6 MHz.



**Figure 20.** Free-space pattern for the antenna of *Figure 5* at 8 MHz.

increases linearly along the antenna, to a good approximation. The antenna is acting as a traveling-wave antenna, akin to the Beverage and others of this class.

## Effect of ground

**Figure 11** shows the vertical plane pattern for a 6-wavelength long antenna placed 20 meters, or 66 feet above ideal earth. Because

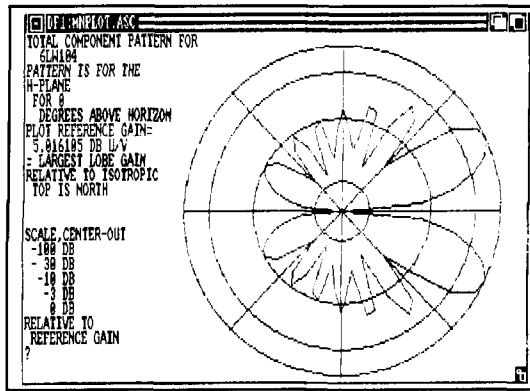


Figure 21. Free-space pattern for the antenna of Figure 5 at 10 MHz.

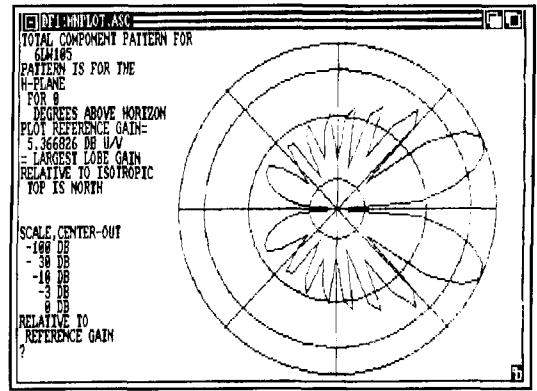


Figure 22. Free-space pattern for the antenna of Figure 5 at 12 MHz.

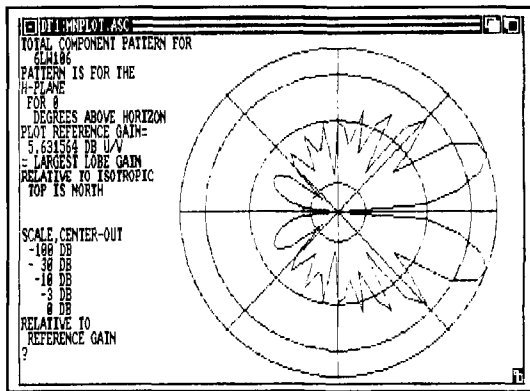


Figure 23. Free-space pattern for the antenna of Figure 5 at 14 MHz.

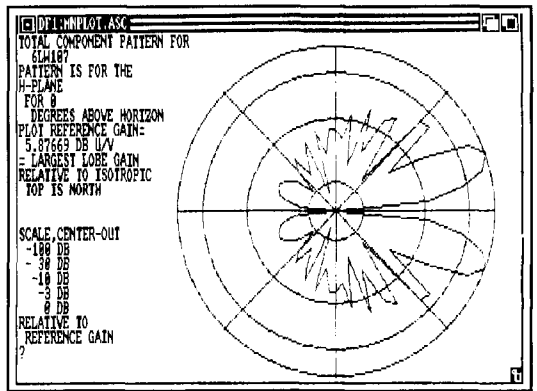


Figure 24. Free-space pattern for the antenna of Figure 5 at 16 MHz.

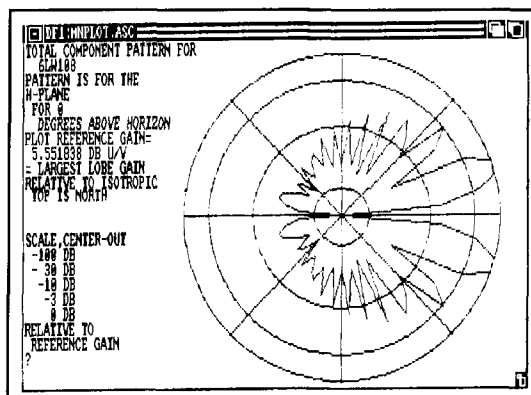


Figure 25. Free-space pattern for the antenna of Figure 5 at 18 MHz.

there is no loading, the F/B ratio is poor—about 3 dB. As a result of the height, the main lobes are relatively strong. The pattern at lower elevations is similar, but with the low angle lobes reduced in size.

Figure 12 shows the vertical plane pattern for this antenna installed as a sloper, with the high end at 35.8 meters or 117.4 feet, and the low end at 8 meters or 26.25 feet. While this antenna isn't loaded, the F/B ratio in the plane of the antenna has improved, to about 9 dB.

This improvement is also shown in Figure 13, a horizontal cut at 4 degrees above the horizon. Note that the reference angle is 180 degrees from that of the previous plot.

However, this apparent improvement in F/B doesn't occur for other vertical planes or elevation angles. This is illustrated by Figure 14—the same antenna but for the horizontal cut at 8 degrees above the horizon. The maximum lobe intensity appears at 40 degrees to the line of the antenna.

It's likely that other combinations of length and slope angle will provide better patterns, especially if two of these slopers are used back-to-back. Such an antenna becomes a half rhombic, in the vertical plane, and rhombic design factors apply. I won't explore these combinations further here.

## Broadband operation

I've already mentioned use of these antennas at other frequencies. Figure 15 summarizes the main pattern features for a 90-meter long wire in free space, nominally designed for 20 MHz. Feed is a quarter wavelength from one end (at 20-meter wavelength excitation), with a 400-ohm load resistance a quarter wave from the other end.

Both maximum gain and maximum F/B occur for a frequency about 10 percent below the 20-MHz design value. Forward lobe gain equals or exceeds that of a dipole over the range 4 to 32 MHz. The lobe in the opposite direction is from 3 to 16 dB smaller.

**Figure 16** shows the drive-point impedance for the range of frequency. Drive resistance is close to 300 ohms for the range 8 to 30 MHz, and becomes large only above 30 MHz. Drive reactance varies less than 200 ohms from zero over the range from 11.5 to 30.5 MHz, becoming large below 8 MHz. The values are within range of an open-wire line antenna matcher of normal design.

**Figures 17** through **25** show the free-space patterns at frequencies of 4, 6, 8, 10, 12, 14, 16, 18 and 20 MHz. As frequency increases, the number of lobes increases, and the main lobes move closer to the axis of the antenna. At the same time, the F/B increases from about 4 to about 14 dB.

This arrangement makes a useful multi-frequency antenna. It's probably not quite as good as a true end-fed, end-loaded design—the double-ended Zepp—but I haven't investigated this design.

## Other designs

The data shown here is for a single antenna dimension, 90 meters or nominally 6 wavelengths in length. To a close approximation, the performance of other lengths can be obtained by scaling. For example, for a 3-wavelength design, patterns and drive resistances are those for a frequency just twice the value for **Figures 15** through **25**. This approximation neglects the effect of wire size, but this will be small unless "invisible antenna" wire sizes are involved.

For more detail, calculate the performance of the specific design with MININEC. As no small angle wire intersections are involved, any version should give adequate answers.

## Concluding note

There is little practical experience with distributed resistance loading to increase F/B and, as far as a literature search revealed none with lumped far-end loading. Perhaps someone will undertake practical trials. It appears that performance approaching that of a rhombic is possible. ■

## PRODUCT INFORMATION

### Integrated Real-Time Simulation And Automated Testing

The Real-Time Intelligent Systems (RTIS) Corporation has two new additions to its Activation Framework line of support tools for developers of C, C++, Ada, and Fortran software. These tools, called AFL and VVTEST, enable programmers to rapidly develop simulations and to perform automated test generation and evaluation in support of all phases of their software development.

When used with RTIS's Integrated Development Framework (IDF) these tools can be used for real-time system simulation, test environment simulation, as well as automated test generation and evaluation at the subroutine, subsystem, and system level. They can also be used for user training and for field maintenance and support.

VVTEST is a compiler that takes test specifications and converts these into C language automated test generation and evaluation modules for use with IDF. The test specifications, written in a rules language called TSL, specify the constraints on the input variables such as allowable values, timings, sequencing and rela-



tionships. They also specify how the resultant tests are to be evaluated. The resultant system is then capable of automatically generating, running, and evaluating thousands of random tests.

AFL is a compiler that takes simulation models written in simple declarative rules and converts these into C code modules for use with IDF and VVTEST. These simulation models can include decision making, have statistical properties, and call upon algorithms written in C, C++, or Fortran.

For further information contact Peter Green at the Real-Time Intelligent Systems Corporation, 30 Sever Street, Worcester, MA 01609, Tel: (508) 752-5567; FAX: (508) 752-5491.

Joseph L. Lynch, N6CL\*  
P.O. Box 73  
Oklahoma City, Oklahoma 73101

# THE CASE OF THE INVISIBLE METEOR STORM

*The shower that nobody saw*

---

**B**ecause of the possibility of a storm during this year's *Perseids* meteor shower, interest in meteor generated propagation is on the increase. However, as strong a possibility for a storm that the *Perseids* presents, it may not top the "invisible meteor storm" of June 1975. This shower, which no one is reported to have seen, took place over a ten day period between June 20 and June 30 and produced daily meteoroid rates five to ten times higher than the normally observed rates.

## Historical meteor storms

In the Spring 1993 issue of *Communications Quarterly*, I briefly described the two events that stand out in the history of meteor shower induced propagation. By way of introduction to the "invisible meteor storm," I'll present a quick review of these two meteor storms and their effects on VHF propagation.

The first, the Giacobinid-Tuttle Comet induced meteor storm, occurred on October 9, 1946. In a giant way it ushered in the ham radio operator's relationship with meteor vaporization induced ionization.

The 6-meter ham band had been authorized only since May of that year, and the operators who used it did not know exactly what to expect by way of propagation. They were aware of meteor generated ionization and called it the "shooting star" effect. However, no one knew what would happen the night of the forecasted meteor shower. Nevertheless, because of the huge publicity surrounding the event, hams

were prepared to be on the band.

They were not disappointed. Ed Tilton, W1HDQ, the then editor of *QST*'s "The World Above 50 Mc," described in his December 1946 column,<sup>1</sup> 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."

Twenty years later, on the night of November 16-17, 1946, operators on 2 meters had set skeds in anticipation of a better than average *Leonids* shower for that year. They, in turn, were treated to what Sam Harris, W1FZJ, at that time the editor of the "World Above 50 Mc" column,<sup>2</sup> headlined as "The Shower of a Lifetime." Harris went on to report that: "Hundreds of contacts were made by calling CQ, or by breaking stations when their skeds were completed, as most were (made) in the first minute or two of prearranged calls." The storm, which lasted less than two hours, did for 2 meters what the Giacobinid-Tuttle storm did for 6 meters in the way of VHF propagation generated by meteor vaporization induced ionization. Truly, these were once-in-a-lifetime events for many radio operators.

## The case for the invisible meteor storm

What these two storms had in common were publicity and visibility. However, as was stated, the largest meteoroid storm ever observed was never known to have been seen by anyone.

Detection of the storm first came by way of the moon.

---

\*Mr. Lynch is the VHF Editor for our sister publication *CQ*.

During the late 1960s, Apollo astronauts left seismometers on the moon to measure moonquakes and meteorite impacts on the moon's surface. There were four passive seismometers. Three of them were arranged in somewhat of an equilateral triangle 1100 km on a side. A fourth was located 180 km away from one of the other three—off a corner of the triangle. They gathered data between 1969 and 1977, when they were turned off because of lack of funding.

Between January and March 1976, F.K. Duennebier and his associates at the University of Texas examined data taken from these devices over a period of 924 days from January 1, 1973 to July 17, 1975. The results of their analysis were published in *Science*<sup>3</sup> magazine.

They found several random registrations of seismic activity on the moon's surface over the period analyzed. During that time they found significant registrations for the period of June 20, 1975 to June 30, 1975, with a peak day of June 22, 1975.

After determining that these events were indeed impacts and not moonquakes, these scientists concluded that the impacts resulted from a cloud of meteors that the moon had passed through during that time frame.

Following their conclusion, they set about to measure the density of the cloud. By comparing the measured data with the mass of the moon, they determined the mass of the cloud to be between  $10^{13}$  and  $10^{14}$ g.

They then compared their observations with known meteor showers and found that this event did not match anything known. Their conclusion was that this was a sporadic meteor storm from a cloud of meteoroids that had "previously gone undetected."

## Correlating VLF data

Assuming that this storm must have had a corresponding effect on the earth, a group of Brazilian astronomers, headed by Pierre Kaufmann, examined VLF data for the same period. The results of their investigations were also published in *Science*<sup>4</sup> magazine.

The researchers chose to study VLF records because of the effects meteor vaporization induced ionization have on the D and E-layers of the atmosphere. The D-layer forms a waveguide effect on signals within the VLF frequency range, transporting them for long distances across the earth's surface. At the terminator line between daylight and darkness, the D-layer dissipates. Because of this dispersment, the signals undergo a phase shift due to the changes in altitudes between the D-layer (then disappearing) and the E-layer.

It was known that meteor vaporization gener-

ated ionization could affect the phase shift caused by the propagation of the signals between the disappearing D-layer and the existing E-layer. It was also known that the D-layer could become so ionized that it would decrease in height above the surface of the earth. This shift in altitude would cause the signal traveling along the "waveguide effect" of the D-layer to also appear to have shifted in phase. Additionally, researchers were aware that the E-layer could also be affected by such ionization. Therefore, Kaufmann and his associates concluded that examination of VLF reception records could reveal any detectable meteor induced effects on these layers of the atmosphere.

They examined records from several VLF radio stations around the globe. What their observations showed were that signals that traveled transequatorially were affected by increased ionization of the D and E-layers. However, signals that didn't cross the equator showed little, if any effects.

Their observations also indicated that the ionization was at such a level as to top the Giacobinid-Tuttle comet generated meteor shower by three to nine times the intensity. Additionally, their findings indicated that the radiant of the shower was in the southern hemisphere and its direction was toward the sun. Because of its position, it was a daytime shower, and not likely to be observed visually.

The Brazilian team also determined that peak days of activity were June 22–23 and June 26. The first days correspond to the peak day of the moon seismic activity. However, the third day does not, and they conclude that this peak could be an anomaly related to ionization of the earth's atmosphere from some other source.

The astronomers stated that they detected the meteoroid storm in four ways. First, they observed a daytime phase advance in the time period observed. Second, they observed large anomalous nighttime phase advances similar to ones that would be observed during sunrise. Third, they detected a phase advance of the mean nighttime phase level and a reduction in the diurnal phase variation. Finally, they observed a distinct reduction in sunset phase delay time rates that, to them, indicated an "extra ionizing source." The team summarized it's findings by stating that in 21 years of observations they had never before seen the first three effects.

## Correlating VHF data?

It is my opinion that if such an intense shower did occur, then, based on past records of on-the-air activity during the two previously cited meteor storms, there should be corresponding data in contemporary VHF records.

The most prolific and continuous records of operating data are listed in *QST*'s regular "World Above 50 MHz" column. Therefore, I examined the columns surrounding the event for correlating data.

Indeed, Bill Tynan's September 1975 column<sup>5</sup> showed that sporadic-E type propagation occurred during these days, with reports of especially intense events occurring on June 22 and June 30.

One of the most interesting reports was of a three-way QSO on June 22 that Tynan (then located in Maryland) had with K3AAY and K8CAY—the later being only 280 miles away in West Virginia. Tynan concluded his observations by saying, "The strength of the signals and the duration of the opening, along with the other signals being received at the same time, leaves little doubt as to the propagation medium involved." It's clear that he believed strongly that the mode of propagation had to be sporadic-E, rather than tropo, or some other form of short-distance propagation.

Tynan also quotes other reports of very short skip contacts during the same day. Oddly, this short-distance propagation was also referred to as typical during the Giacobinid-Zinner Comet induced shower.

Tynan goes on to quote a report from W7NFC in Athens, Oregon that indicated contacts with all states in the W1, W4, and W5 call areas, plus numerous contacts in the W9 and WØ call areas during the day of June 22. He also quotes from a report received from Bob Mobile, WA1OUB, concluding that "...the day-of-days was June 22, with QSOs all over the country (being reported)."

Tynan also reports on receptions made on June 30 by Pat Dyer, WA5TYX, (near San Antonio, Texas) of numerous signals indicated—particularly of South American TV stations. Additionally, he quotes a report from Glenn Hauser of Enid, Oklahoma, indicating reception of high VHF TV stations on the same date.

June 22 seems to be a key day for all three sets of data. However, Tynan does not report data that would correlate to the Kaufmann-observed peak on June 26. Could it be that many hams were on the air on Sunday, June 22, and that few hams were on the air on Thursday, June 26? Additionally, he reports tremendous activity on June 30, while neither Kaufmann nor Duenebier report correlating data. Could data on this date be related to a sporadic-E event? Could sporadic-E events be occurring simultaneously with meteor storm induced ionization on any or all of these dates?

For the amount of meteor shower activity, there seems to be little other correlating amateur radio VHF data (absence of 2-meter reports, for example). Could the amateur radio

observations be incomplete because "nobody was on the air?" In conducting unrelated research, I looked back into my 6-meter log for the last three years and found that the band has been open each Memorial Day weekend—no matter that the dates of the weekend have floated. Without exception, the band was open during some time during those days. Was the band open because people were home and on the air, or was the fact that band was open and people were home a coincidence?

This Memorial Day propagation phenomenon was demonstrated to me again this past holiday when I was in Mexico. While operating from my station in Ojinaga, I observed that I had propagation on the days before Sunday but, after working a few stations, I couldn't work anyone else because no one was on the air. Curiously, my best propagation day was Sunday, when most people were likely to be home. However, with the Indianapolis 500 auto race running simultaneously with the time frame of the propagation, I once again ran out of stations to work before I ran out of propagation.

One other factor to consider when looking for correlating amateur radio VHF data is this: Because this storm was in the southern latitudes, the most accurate of the propagation reports would be coming from these areas. Unfortunately, not a lot of amateur radio VHF activity takes place there. To date, no contemporary propagation reports from this area have been recovered. Therefore, as anecdotal as these reports from Tynan are, they must be viewed as inconclusive.

## Conclusion

This article represents a preliminary investigation of the data that are available. I would like to be contacted by anyone who might have correlating data. One of my questions concerns the anomaly of the Kaufmann-reported peak of June 26 and the Tynan-reported propagation on June 30. Was there a sudden ionospheric disturbance (SID) or other solar event that affected Kaufmann's observation, or was there a lack of data from amateur radio operators for that date? Were the events that occurred on June 30 strictly related to sporadic-E induced propagation? Data from readers of this article may provide clues to answers for this and other questions. ■

## REFERENCES

1. Ed Tilton, W1HDQ, *QST*, vol. 30, December 1946, page 43.
2. Sam Harris, *QST*, vol. 51, January 1967, page 61.
3. F.K. Duenebier, Y. Nakamura, G.V. Latham, and H.J. Dorman, *Science*, vol. 192, June 4, 1976, pages 1000–1002.
4. P. Kaufmann, V.L.R. Kuntz, N.M. Paes Leme, L.R. Piazza, J.W.S. Vilas Boas, K. Brecher, J. Crouchley, *Science*, vol. 246, November 10, pages 787–790.
5. Bill Tynan, *QST*, vol. 69, September 1975, pages 75, 136, 138, and 140.

# HIGH-FREQUENCY BYPASS CAPACITORS

*An empirically designed model*

Capacitors are commonly used for a variety of applications including coupling, timing, bypassing, etc. In high-frequency applications, there is specific concern with the parasitics of capacitors used for these various purposes. For example, the equivalent series inductance results in parasitic resonances in the capacitor, and the equivalent series resistance results in a limiting minimum impedance. When a capacitor is used as a bypassing element, it is generally desirable for that capacitor to exhibit a very small impedance. Therefore, in a bypass capacitor, the equivalent series inductance and equivalent series resistance must be minimized. The equivalent series resistance, in part, is a function of capacitor quality. High-Q components will by design exhibit much lower equivalent series resistance than general-purpose components. The equivalent series resistance is primarily a function of the actual capacitor construction and cannot be affected significantly by the external circuit in which the capacitor is used.

The nature of the equivalent series inductance of a capacitor is quite different from the equivalent series resistance. The apparent equivalent series inductance of a capacitor is almost always influenced more by the external circuit than the actual capacitor itself. Often it is thought that the better the capacitor, the lower the equivalent series inductance. That is not necessarily the case. To understand this, we must examine inductance briefly.

## A look at inductance

The parameter inductance is basically a *constant of proportionality* relating magnetic flux and current, but that is perhaps an unusual

description. From Ampere's Law, we know that magnetic field  $B$  (units of Teslas) at some point in the vicinity of a current is proportional to that current. By adding a proportionality constant,  $\mu_0$ , the common form of Ampere's Law is constructed as **Equation 1**. The circle on the integral symbol is the common notation used to indicate the integral is performed around a closed loop.

$$\oint \mathbf{B} \cdot d\mathbf{l} = \mu_0 i \quad (1)$$

Ampere's Law

Using Faraday's Law, it can be shown that the EMF, e.g., the potential, induced in a loop of wire (this is not the same loop as for the integral of Ampere's Law) is proportional to the time rate of change of the total magnetic flux (units of Webers) through the loop. The magnetic field and magnetic flux are related by the area through which the flux penetrates (1 Tesla = 1 Weber/Meter<sup>2</sup>).

$$V = -\frac{d\Phi}{dt} \quad \text{Faraday's Law (for a single-turn 100p)}$$

so,

$$V \propto -\frac{di}{dt} \quad \text{Self induced potential}$$

The subtle but all important point here is that Faraday's Law applies specifically to a closed loop. This closed loop may be of any arbitrary shape, but it must be closed. This is more easily understood if you ask the question: If the loop were not closed, how could current flow? Current must flow around a closed path, although the exact path may be quite obscure in many cases. Just where does the current that is fed into the feed line of an antenna go? If a pro-

portionally constant  $k$  is added, the proportionality may be written as an equality.

$$V = -k \frac{di}{dt}$$

That proportionality constant is called inductance, or  $L$ . So:

$$V = -L \frac{di}{dt}$$

Consequently, from Ampere's Law the current and magnetic field of a circuit are related, and from Faraday's Law the potential induced in a closed-loop circuit is related to the change in magnetic flux penetrating the loop. Combining Ampere's and Faraday's Laws, the self-induced potential developed in a loop is related to the change in current in the loop by the proportionality constant inductance. This very loosely demonstrates that inductance is defined in terms of a closed-loop circuit.

Therefore, it is technically incorrect to speak of the "inductance" of a component, such as an inductor. In contrast, one can correctly define the value of a resistive component as the actual resistance of that component, because the actual resistance of that component is independent of the external circuit. This is also true of a capacitor. However, it is not true of an inductor.

When the value of an inductor is specified, it is assumed that the inductance contribution of the total circuit loop of which the inductor is a part is insignificant compared to the contribution of the inductor. In general, this is a reasonable assumption for comparatively large "inductors," perhaps greater than a few microhenrys. But that is not the case for very small values of inductance, or nanohenrys. For circuit elements exhibiting nanohenry inductances, the physical geometry of the circuit determines the "inductance" of the element.

To demonstrate this, let's consider how one would determine the inductance of a 1 cm length of 22-gauge copper wire. Without a geometry defined, the inductance cannot be computed. So, let this length of wire be placed 1 cm above and parallel to a ground plane separated by an air space. This configuration constitutes a short transmission-line segment. From the properties of a transmission line, the inductance per unit length may be computed.

$$Z_0 = \sqrt{\left(\frac{L}{C}\right)}$$

Characteristic impedance

$L$  = inductance per unit length

$C$  = capacitance per unit length

$$V_p = \frac{1}{\sqrt{(L \cdot C)}}$$

Characteristic Propagation Velocity

Solving for the inductance:

$$L(\text{per unit length}) = \frac{Z_0}{V_p}$$

For an air dielectric, the propagation velocity is the speed of light in air (vacuum),  $V_c = 3 \times 10^8$  m/s. For a wire over ground plane where the height above ground plane is  $h$  and the wire diameter is  $d$ , the characteristic impedance is given by:

$$Z_0 = 60 / \epsilon^2 \ln(4h/d) \quad [h \gg d] \quad \epsilon = 1 \text{ for air}$$

This expression may be found in virtually any reference on transmission lines. Substituting,

$$L [\text{per unit length}] = 60 / (\epsilon^2 V_p) \ln(4h/d)$$

The 22-gauge wire diameter is nominally 0.065 cm. The 1 cm height is much greater than the 0.065 cm wire diameter. Therefore,

$$L = 824 \text{ nH/m}$$

Thus, a 1-cm length of this wire will exhibit an "inductance" of nominally 8 nH. But what if the wire were placed 2 cm above the ground plane? The inductance would then be about 10 nH. Or what if a spacing of 0.1 cm were used? Then the inductance of the 1 cm length of 20-gauge wire would be about 4 nH (here  $h$  is not  $\gg d$  so the expression above is not accurate but still provides a reasonable estimate).

This simple example has shown that we cannot simply specify the "inductance" of the 1-cm length of 22-gauge wire as we can specify the resistance of a resistor or the capacitance of a capacitor. The inductance of the wire segment is directly a function of the external circuit of which it is a part. In the case of comparatively large inductive elements, the relative contribution of the external circuit is negligible. However, for very small inductances, such as the parasitic inductance of a capacitor or resistor, the magnitude of those inductances is highly dependent on the physical arrangement of the elements.

The wire over ground plane was chosen for the example above because it is a good approximation of a circuit component like a bypass capacitor installed in a printed-circuit board or hybrid RF circuit. It should be evident from the



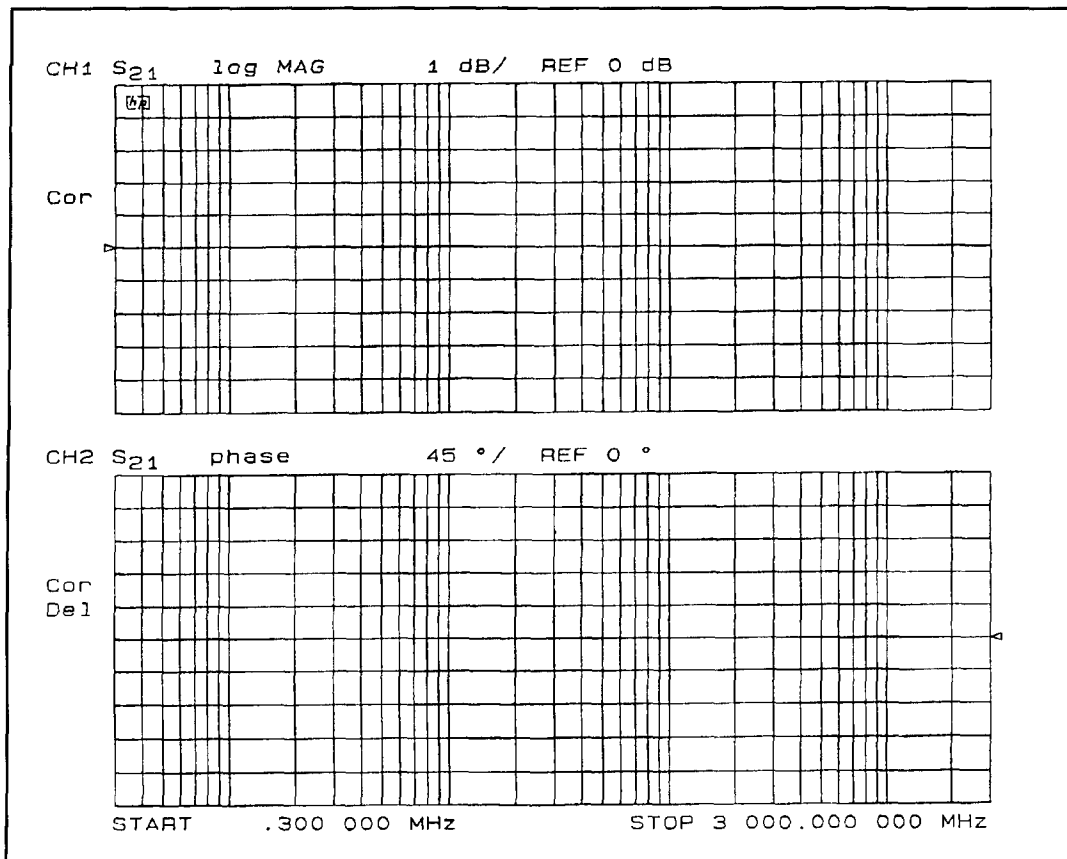


Figure 1. Test-fixture network-analyzer response.

example that the parasitic inductance of a bypass capacitor will be highly a function of how it is mounted. If long leads are used with the capacitor mounted far from the ground plane, the parasitic inductance will be high; but if it is mounted close to the ground plane with minimum-length leads, the inductance will be much lower.

### Capacitors used as bypass capacitors

When a capacitor is used as a bypass capacitor to ground, a closed loop is formed from the grounded terminal of the capacitor, through the capacitor to the bypassed node, through the circuit elements attached to the bypassed node, and back to ground. The smallest that the perimeter of that loop can be made is twice the length of the capacitor—from ground through the capacitor, and then a direct connection back to ground. Because all capacitors have some physical length, the inductance associated with the capacitor cannot be made zero because the smallest closed loop around the capacitor cannot be made zero. This is a property of the physics of the problem, and the actual quality

of the capacitor has little effect on inductance (a very poor capacitor or a very poor installation of a capacitor can certainly increase the inductance above the physical limits).

The effect of the loop around a bypass capacitor can be demonstrated easily with a network analyzer and a length of transmission line. A suitable transmission line for such analyses is 0.141 semirigid coaxial line. For a simple short length of line, the through-line response should simply be a straight line at 0 dB with some small loss with frequency. A small hole is then cut through the outer conductor and dielectric to provide access to the center conductor. If that hole is small enough, it will cause little effect in the through-line response up to some frequency related to the hole dimensions. For example, **Figure 1** shows the through-line response of the fixture used to collect the data presented in this work. The 0.141 semirigid line is used because it is quite easy to fabricate this configuration while preserving the general integrity of the coaxial configuration. This peculiar transmission-line structure is the capacitor test fixture for observing the effect of the equivalent series inductance. As seen in **Figure 1**, the response of the test-fixture line, including the hole in the outer con-

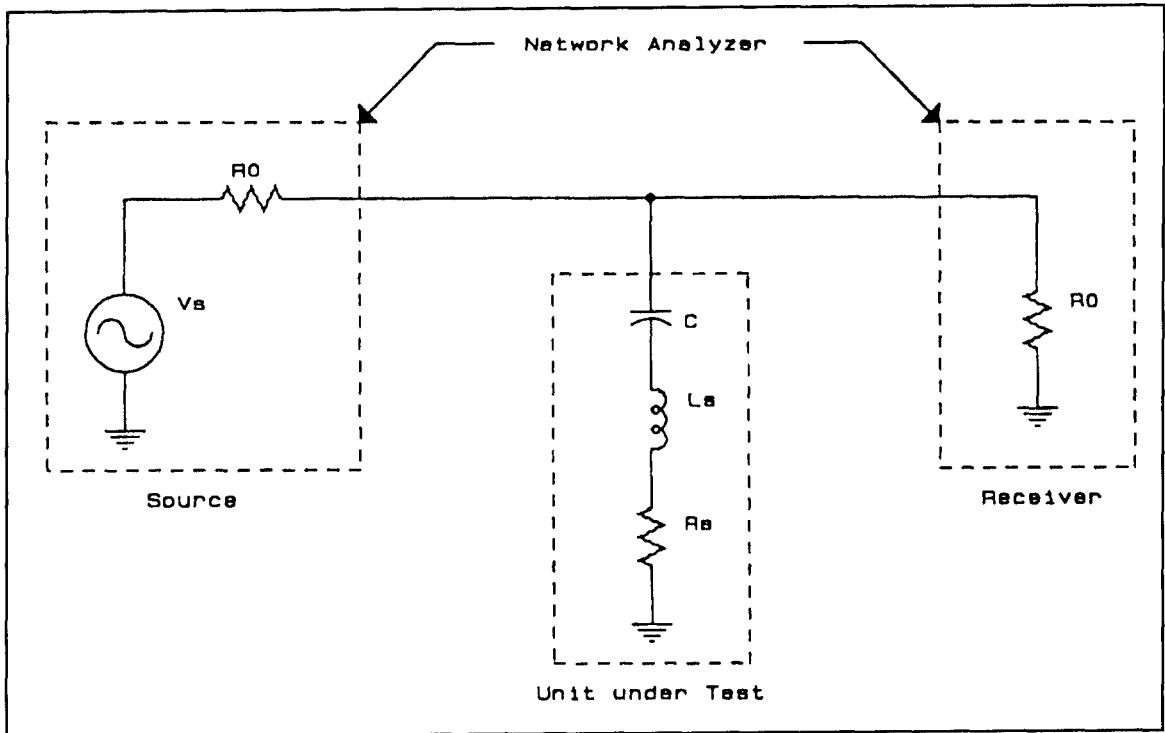


Figure 2. Test configuration equivalent circuit.

ductor and dielectric, is almost ideal in relation to the 3-GHz limit of the measurement.

### Analyzing a capacitor

A capacitor to be analyzed is attached between the exposed center conductor and the outer conductor. The loop formed runs from the outer conductor, through the capacitor to the center conductor, and then back to the outer conductor through one-half the characteristic impedance of the line. The smallest possible loop is then two radii of the line. For very small chip capacitors, the actual line used must be chosen so the distance from the outer surface of the center conductor to the inner surface of the outer conductor is nominally the length of the capacitor to be analyzed. For very small capacitors, if this general method of analysis is to be used, other line configurations, such as a microstrip line, are likely more suitable.

When a capacitor is installed in the test fixture, it acts as a bypass capacitor—bypassing the center conductor to ground. The impedance seen by the capacitor looking into the line is one-half the characteristic line impedance. At the point of attachment of the capacitor, there are effectively two transmission-line segments attached to that node: one to the source and one to the receiver of the network analyzer. As a result, the two line impedances are in parallel and the effective impedance seen by the capaci-

tor is then one-half the line impedance. If the hole in the transmission line is made as small as practical, the physical position of those two impedances lies immediately at the attachment point of the capacitor under test and independent of the length of line on either side of the capacitor—a basic property of a terminated transmission line. This provides a minimum-length return path from the center conductor to the outer conductor, and that is why a transmission-line structure is very convenient for this analysis.

Figure 2 shows an equivalent circuit of the test configuration. The capacitor under test is  $C$ ,  $R_0$  is the characteristic impedance of the test fixture transmission line (50 ohms here), and  $L_s$  and  $R_s$  are the equivalent series inductance and resistance, respectively.

The network analyzer used for the work reviewed in the following paragraphs was limited to an upper frequency of 3 GHz. A rather large silver mica capacitor was chosen for some of the analyses, so the effects could be observed at frequencies well below 3 GHz. Nevertheless, small, high-performance microwave capacitors will exhibit similar performance but at much higher frequencies.

Figure 3 shows a basic block diagram of the test configuration. The two 10-dB pads are used to ensure that the transmission line is well terminated in both directions. In all tests, one lead of the capacitor under test was cut to the minimum practical length, about 1 mm. That

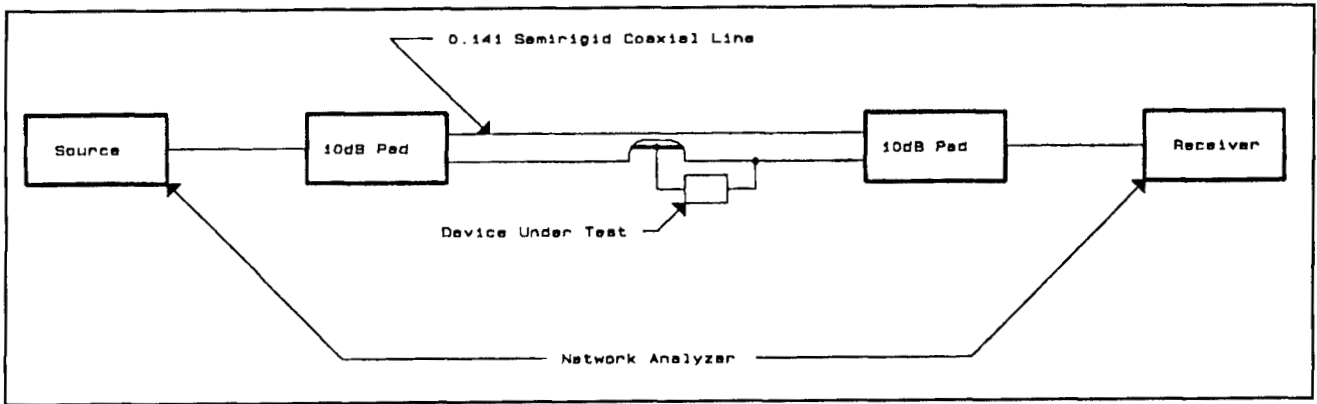


Figure 3. Test configuration block diagram.

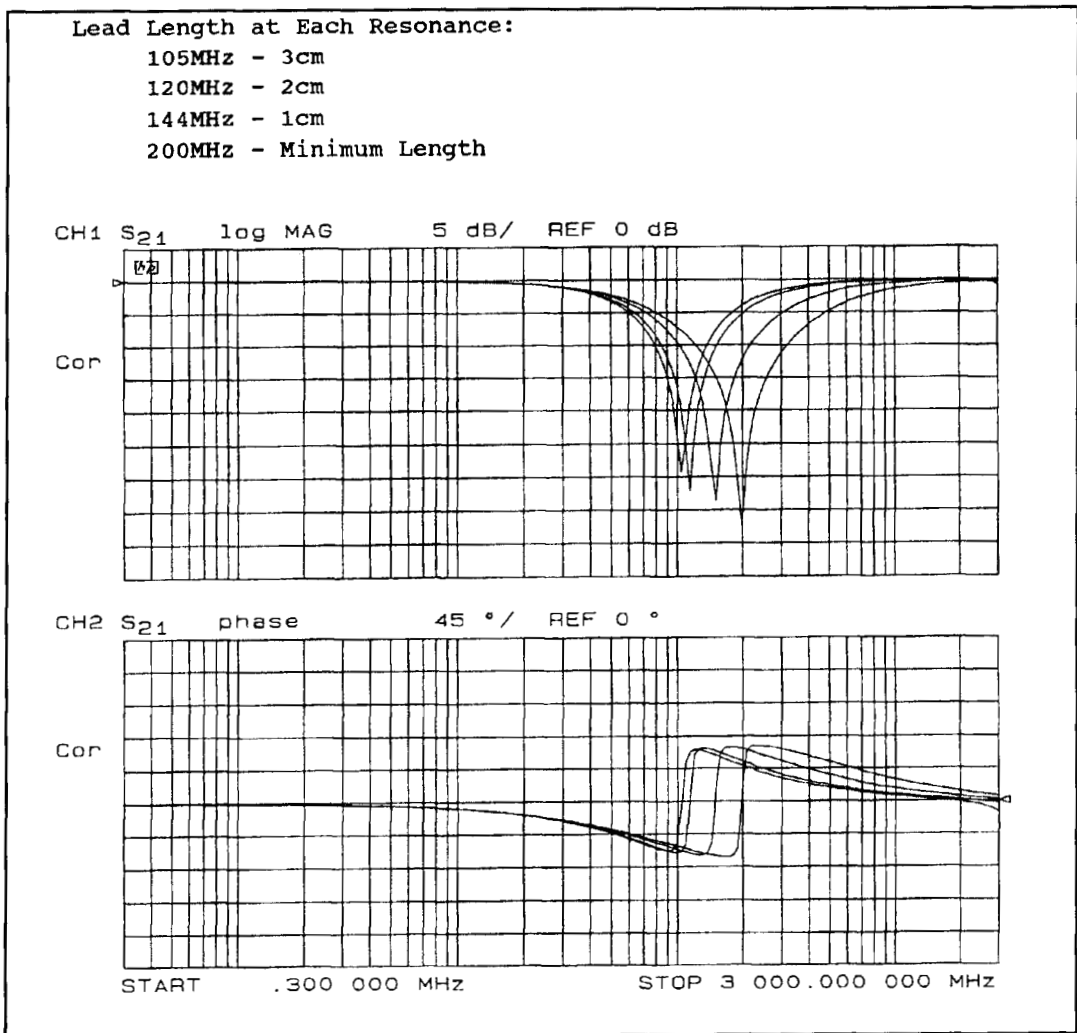


Figure 4. 100-pF CM101 capacitor response with several lead lengths.

minimum-length lead was attached to the test-fixture outer conductor. The other lead of the capacitor was made various lengths as indicated in the particular test. Also, in all cases, the variable-length lead was run parallel with the test-fixture outer conductor separated from the outer conductor by about 2 mm.

A 100-pF CM101 silver-mica capacitor was selected for analysis. The measured capacitance was 102 pF. This capacitor was attached between the center conductor and outer conductor of the test fixture as described above. **Figure 4** shows the response of this capacitor with lead lengths of 3 cm, 2 cm, 1 cm, and min-

imum length. In all cases, a classic series resonance is seen. For the minimum-length case, the resonance is at about 200 MHz and the depth of the response is about 37 dB. Also, at the response minimum, the phase is 0 degrees. The capacitor is therefore totally resistive at that frequency. That resistance is the equivalent series resistance of the capacitor in this configuration. Note that the response minimum is less deep with longer lead length. This is due to the various effects, such as skin effect, and clearly demonstrates the importance of minimum length leads. The capacitive reactance and the source impedance, one-half the line impedance, or 25 ohms for the test fixture, forms a simple voltage divider. At series resonance, the voltage divider is a simple resistive divider. The equivalent series resistance may be easily computed from the response magnitude at that frequency (see Equation 2). Let  $M_f$  be the response magnitude in dB at the series-resonant frequency and  $k_s$  the voltage attenuation ratio at resonance. Also, let  $L_s$  and  $R_s$  be the equivalent series inductance and resistance, respectively, at that resonance, as noted above.

Figure 5 shows the equivalent circuit of the test system at series resonance. From an analysis of Figure 5, the equivalent series resistance is given by Equation 3.

$$k_s = \log^{-1}(M_f/20) \quad (2)$$

$$R_s = (R_0 / 2) \left( \frac{k_s}{1 - k_s} \right) \quad (3)$$

Thus, the CM101 with minimum length leads exhibits an equivalent series resistance of about 0.4 ohms. The equivalent series inductance may be computed from the resonant frequency  $f_r$  and the capacitor value  $C$ .

$$f_r = \frac{1}{2\pi\sqrt{L_s C}}$$

$$L_s = \frac{1}{(2\pi f_r)^2 C} \quad (4)$$

The equivalent series inductance of the CM101 with minimum lead length is then nominally 6 nH. To verify this inductance, a nominal 1 cm length of 22-gauge solid wire (1/4-watt resistor lead) about the length of the capacitor was placed in the test fixture with the lead run parallel to and about 2 mm from the outer conductor. Figure 6 shows the response of the wire segment. As expected, its response is that of a near ideal inductor. Using Equation 3 the impedance  $X_L$  of the wire may be computed at any point. For example, at about 7 MHz the response is -40 dB. The impedance at that point is 0.25 ohms. The inductance is then simply  $X_L/(2\pi f)$ . The inductance of the 1-cm wire is then about 6 nH. This close result

strongly suggests that the actual capacitor contributes little to the inductance.

## Capacitors in parallel

It is often suggested that improved bypassing can be achieved by using several capacitors in parallel. That is not generally true. To demonstrate this, two CM101 capacitors were connected in parallel in the test fixture. The response of the parallel combination is shown in Figure 7. Note that there are two series-resonant frequencies and a parallel-resonant frequency. At about 120 MHz, a series resistance of about 0.2 ohms is seen. However, at about 150 MHz, at the parallel resonance, the effective resistance is only about 20 ohms. Thus, at 150 MHz, very little bypassing is provided by the parallel combination of the two capacitors. This is because one capacitor resonates in a parallel resonance with the inductance of the other capacitor.

The response of parallel capacitors can be more dramatically shown using a small capacitor paralleled by a larger one. This is a common practice where the small capacitor is to provide high-frequency bypassing and the larger one bypassing of lower frequencies. A 0.01- $\mu$ F CK05 capacitor was selected as a typical larger capacitor to be paralleled with the CM101. Again, that value was selected to ensure that the response was well within the range of the network analyzer. Figure 8 shows the response of the 0.01- $\mu$ F CK05 ceramic capacitor. This capacitor also exhibits a series resonance, at about 20 MHz. The equivalent series resistance at series resonance is about 0.14 ohm. The equivalent series inductance is about 6 nH or about the same as for the CM101. This again indicates that the series inductance is indeed more a function of the geometry than the capacitor characteristics.

Figure 9 shows the response of the CM101 in parallel with the 0.01- $\mu$ F CK05 capacitor, both with minimum-length leads. There are two series resonant frequencies, one at about 20 MHz, corresponding to the 0.01- $\mu$ F capacitor, and the other at about 200 MHz, corresponding to the CM101 as demonstrated above. There is also the expected parallel resonance at approximately 150 MHz. That is the parallel resonance of the CM101 and the equivalent loop inductance of the CM101 and the CK05—about 11 nH from Equation 4. In the vicinity of 150 MHz, the parallel combination of the 100-pF CM101 and the 0.01- $\mu$ F CK05 provides virtually no bypassing. If this configuration of these two capacitors were used to bypass the power-supply node of a wideband amplifier, the amplifier would very likely oscillate at about 150 MHz due to coupling between various ele-

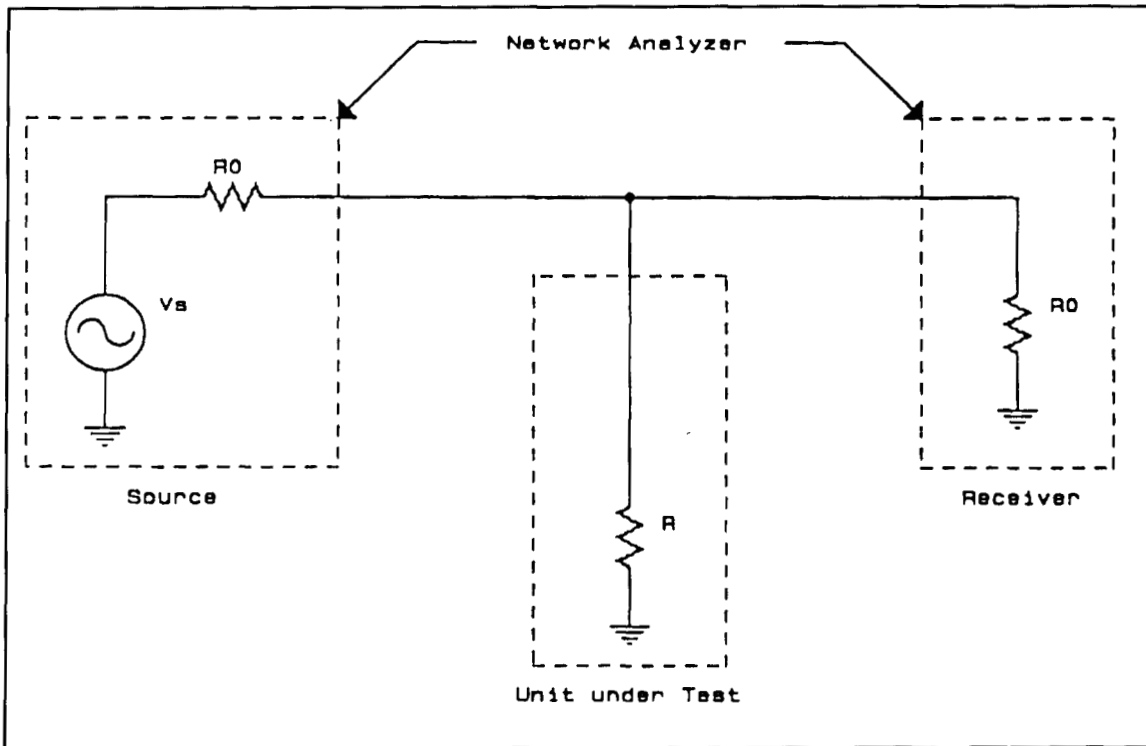


Figure 5. Capacitor equivalent circuit at series resonance.

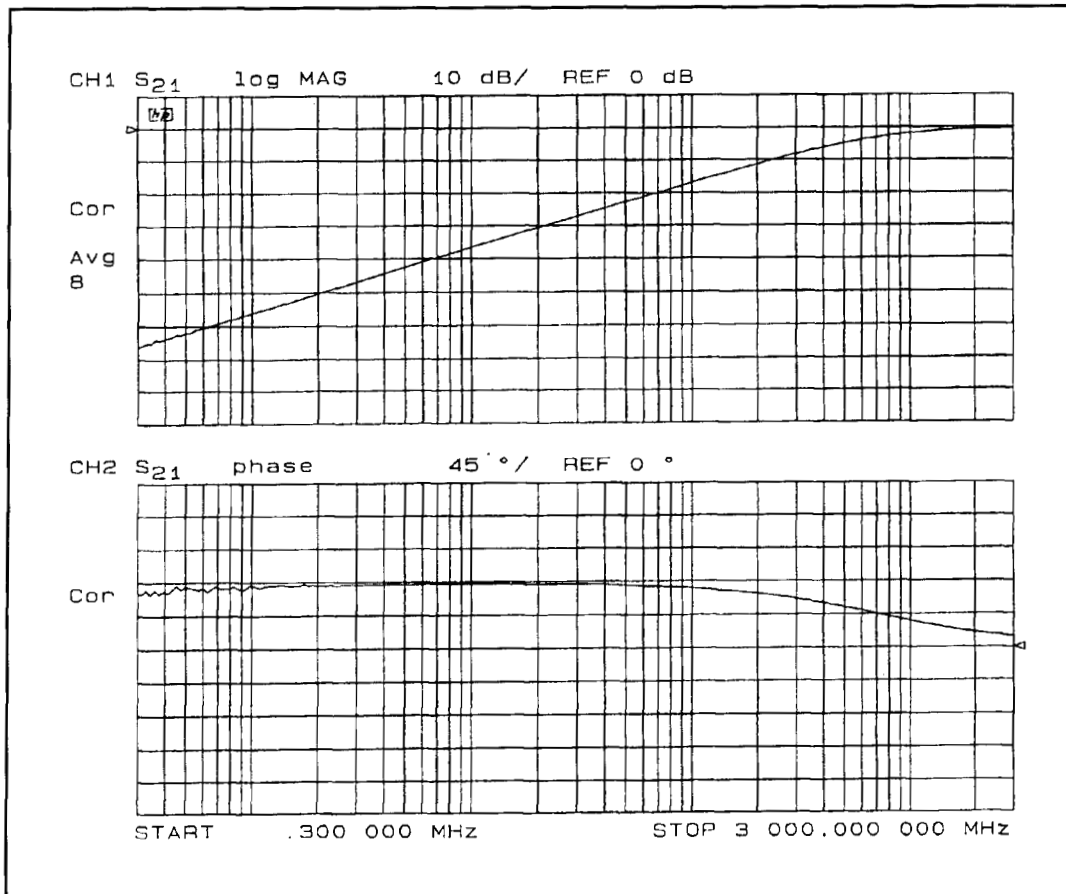


Figure 6. Response of 1-cm length of 22-gauge wire.

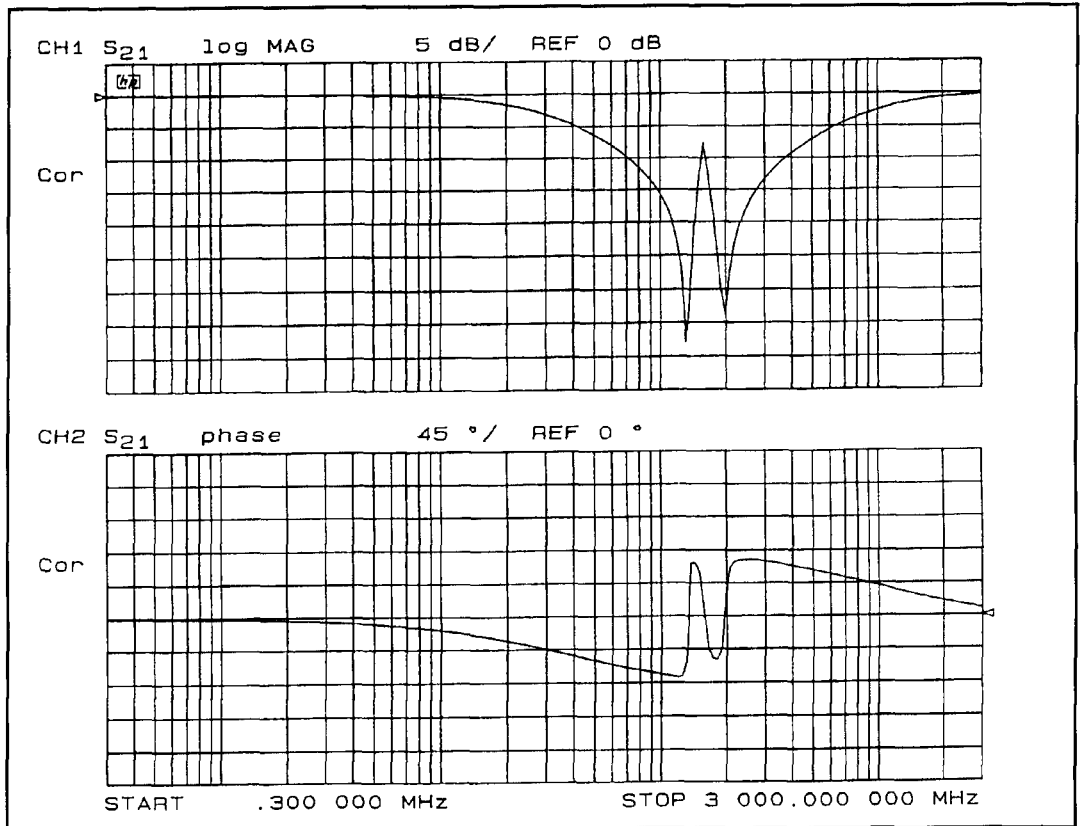


Figure 7. Response of two 100-pF CM101 capacitors in parallel (minimum lead lengths).

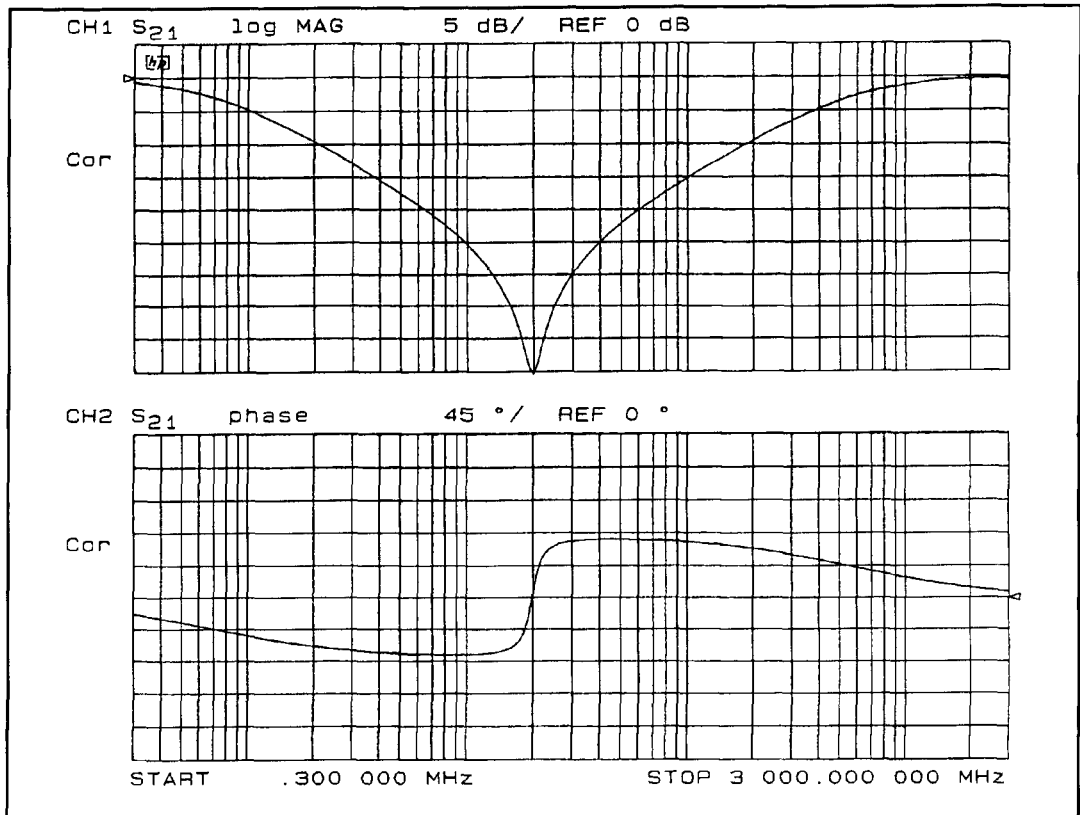


Figure 8. Response of a 0.01-μF CK05 capacitor (minimum lead lengths).

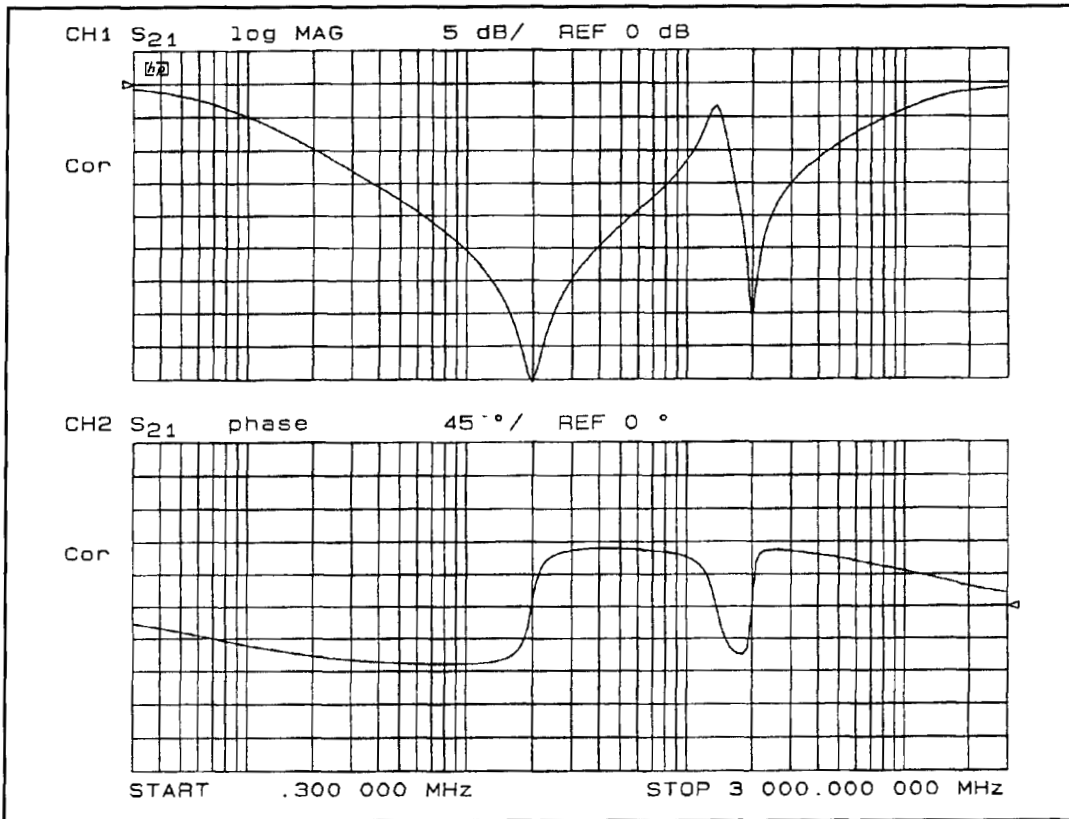


Figure 9. Response of parallel combination of a 100-pF CM101 capacitor and a 0.01- $\mu$ F CK05 capacitor (minimum lead lengths).

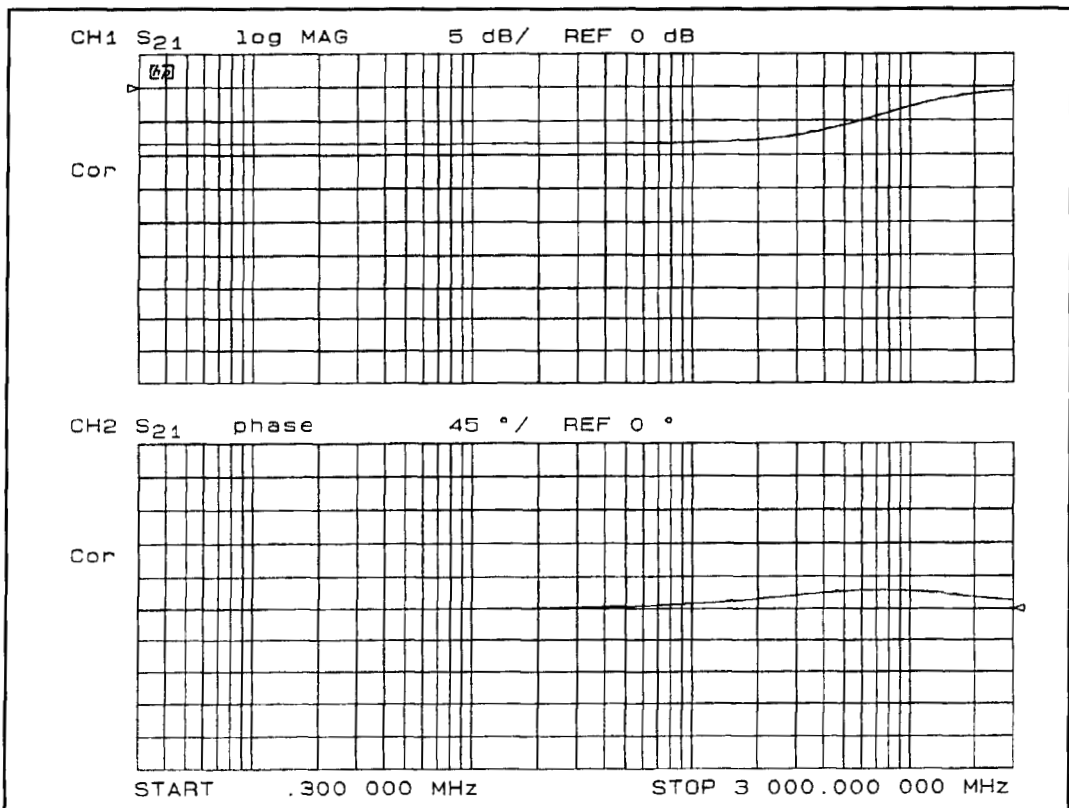


Figure 10. Response of a 15-ohm RCR07 1/4-watt resistor (minimum lead lengths).

ments of the amplifier via the power-supply circuit at that frequency.

The simple paralleling of capacitors results in a severe compromise in the bypassing performance of the combination due to the effective parallel-resonance configurations formed. However, parallel operation can be provided if a little thought is given to the problem. Looking back at the equivalent circuit of **Figure 2**, it is seen that the capacitor is a simple second-order network. The capacitance is fixed by the specific choice of value. The equivalent series inductance is fixed by the physical circuit geometry. The minimum value of equivalent series resistance is generally fixed by the quality of the capacitor. However, the series resistance could be made larger by the addition of more series resistance. But, if one chooses a high-quality capacitor to minimize series resistance, why would one wish to increase the series resistance by adding more resistance?

## Damping

The addition of series resistance can be used to control the response of the second-order R-L-C equivalent network of the capacitor. In other words, if the correct value of resistance is added, the highly resonant networks can be critically damped or even over damped.

From **Figure 2**, the impedance of the equivalent circuit of the capacitor may be directly computed as given in Laplace form in **Equation 5**.

$$Z_c = [s^2 + s(R_s/L_s) + 1/(L_s C)] (L_s/s) \quad (5)$$

The characteristic equation of the capacitor equivalent circuit is classic second order.

$$\omega_n^2 = 1/(L_s C) \text{ Natural Frequency} \quad (6A)$$

$$2\xi\omega_n = R_s/L_s \quad (6B)$$

$$\xi = \frac{R_s}{2\omega_n L_s} \quad (6C)$$

For critical damping, the damping factor  $\xi$  must be equal to  $1/\sqrt{2}$  and for over damping,  $\xi$  must be greater than  $1/\sqrt{2}$ . So, by setting **Equation 6C** greater than  $1/\sqrt{2}$ , solving the inequality for  $R_s$ , and substituting **Equation 4**, the value of series resistance for critical and over damping may be found.

$$\xi > \frac{1}{\sqrt{2}}$$

$$\frac{R_s}{2\omega_n L_s} > \frac{1}{\sqrt{2}}$$

$$R_s \text{ (over damped)} > \sqrt{\frac{2L_s}{C}} \quad (7A)$$

$$R_s \text{ (over damped)} > \frac{1}{\sqrt{2\pi f_r C}} \quad (7B)$$

So, if the series resistance is equal to the value computed in **Equation 7**, the R-L-C network will be critically damped, and for higher resistance it will be over damped. From **Figure 4**, for the CM101 with minimum lead length, a series resistance of about 11 ohms is needed. The capacitor alone is about 0.4 ohms, so if a 10.6-ohm resistance were added, the capacitor response would be critically damped. But, the addition of a resistor will increase the overall circuit path length resulting in a larger inductance. For example, a standard 1/4-watt resistor with minimum lead length is about 1 cm in length. Therefore, the equivalent series inductance of the capacitor in a configuration with one lead about 1 cm long would be the more accurate model. From **Figure 4**, with a 1 cm lead length, the resonance is about 150 MHz—corresponding to an inductance of about 11 nH. The resistance required for critical damping is then about 15 ohms.

Unfortunately, resistors, too, have parasitics. **Figure 10** shows the response of a 15-ohm 1/4-watt RCR07 resistor in the transmission-line test fixture. This resistor has a network zero at about 700 MHz, which corresponds to an equivalent series inductance of about 3 nH. (It is difficult to estimate the precise frequency of the zero from the figure due to the close proximity of the zero at about 700 MHz and the pole at about 1.5 GHz.) **Figure 11** shows the response of a 2.7-ohm 1/4-watt resistor. The network zero at about 150 MHz, again corresponds to a nominal 3-nH inductance. Because the diameter of the actual conducting element in the 1/4-watt resistor is very much larger than that of the 22-gauge wire of **Figure 6**, one would expect a resistor to exhibit a lower inductance than the wire. These results are also consistent with the CM101 inductance.

**Figure 12** shows the response of the CM101 with a 15-ohm 1/4-watt resistor in series, both with minimum lead length. As expected, the response is now damped. **Figure 13** shows the response of two of these capacitor-resistor networks in parallel. Note that the parallel combination no longer shows the parallel-resonant behavior. The parallel combination of the damped capacitors provides improved bypassing over that of a single capacitor, as was originally the goal of paralleling capacitors.

**Figure 14** shows the response of the 100-pF CM101 capacitor with a 15-ohm series resistor



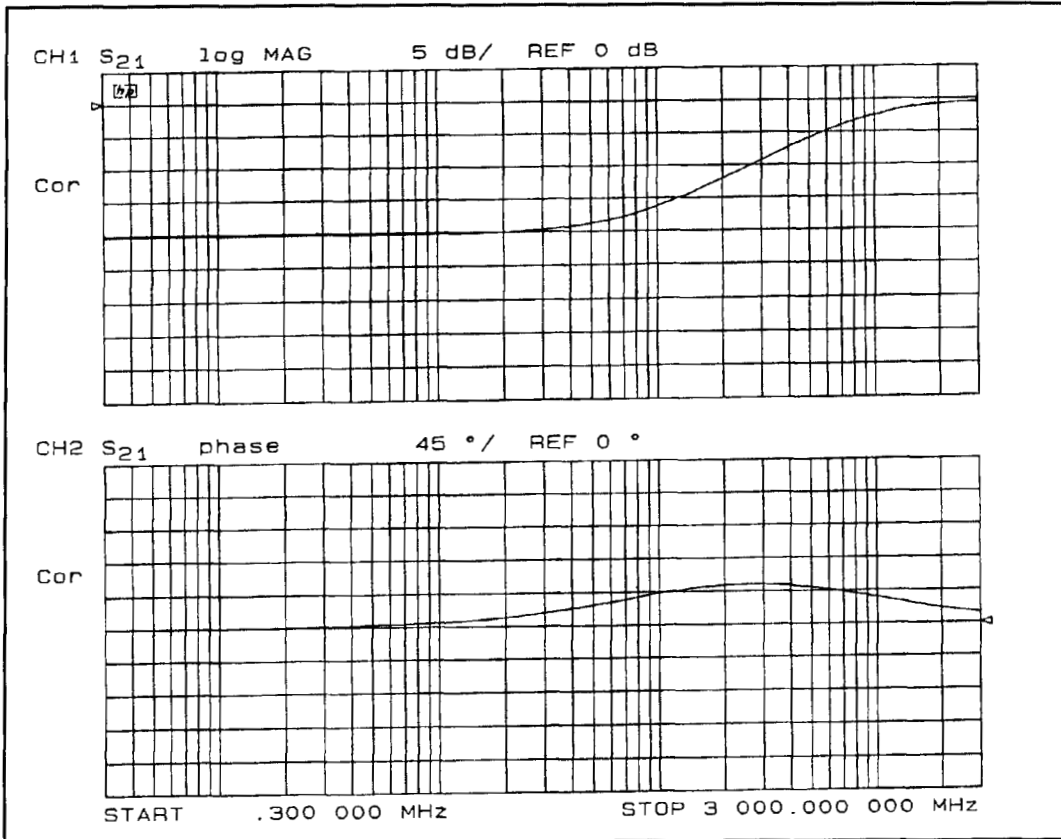


Figure 11. Response of a 2.7-ohm RCR07 1/4-watt resistor (minimum lead lengths).

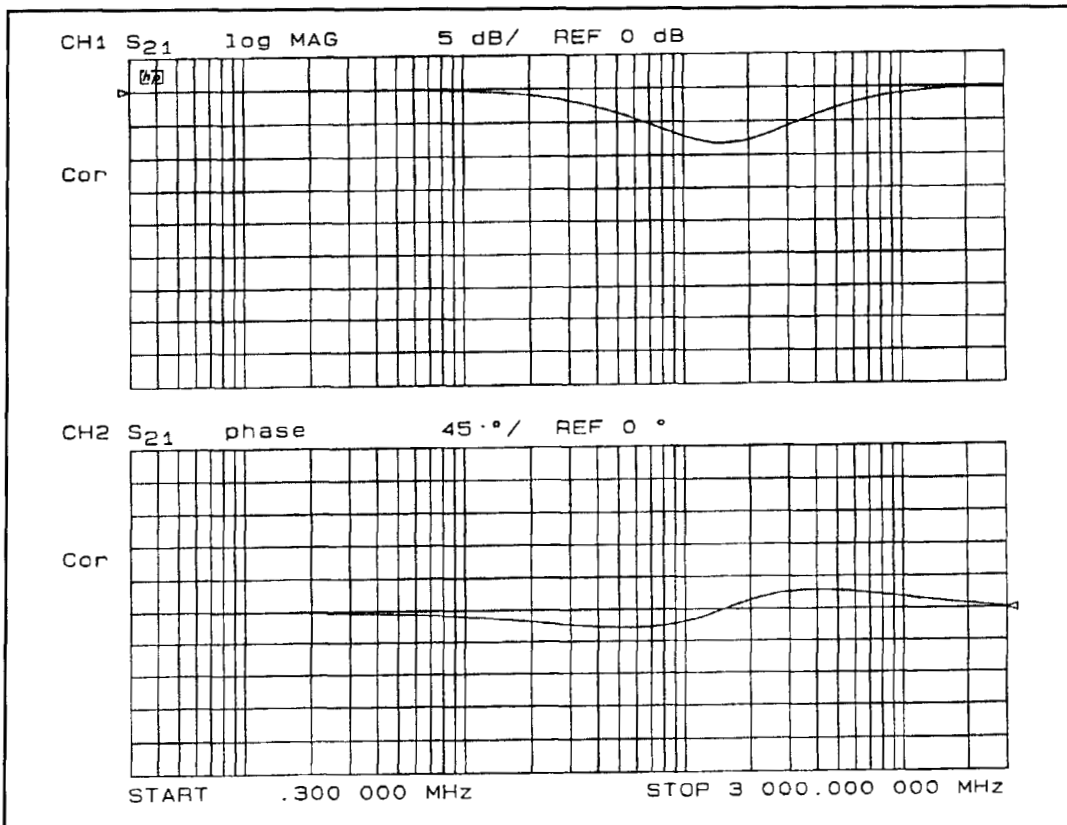


Figure 12. Response of a 100-pF CM101 capacitor in series with a 15-ohm RCR07 resistor (minimum lead lengths).

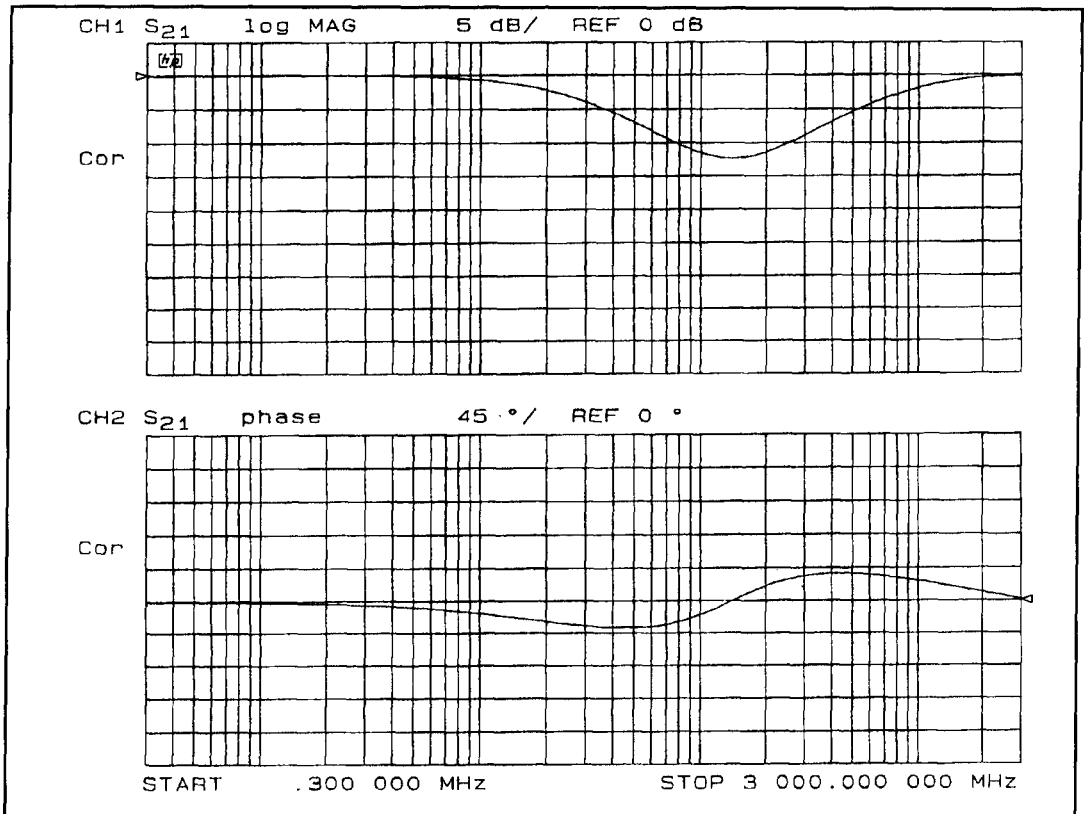


Figure 13. Response of a parallel combination two series networks of a 100-pF CM101 capacitor in series with a 15-ohm RCR07 resistor (minimum lead lengths).

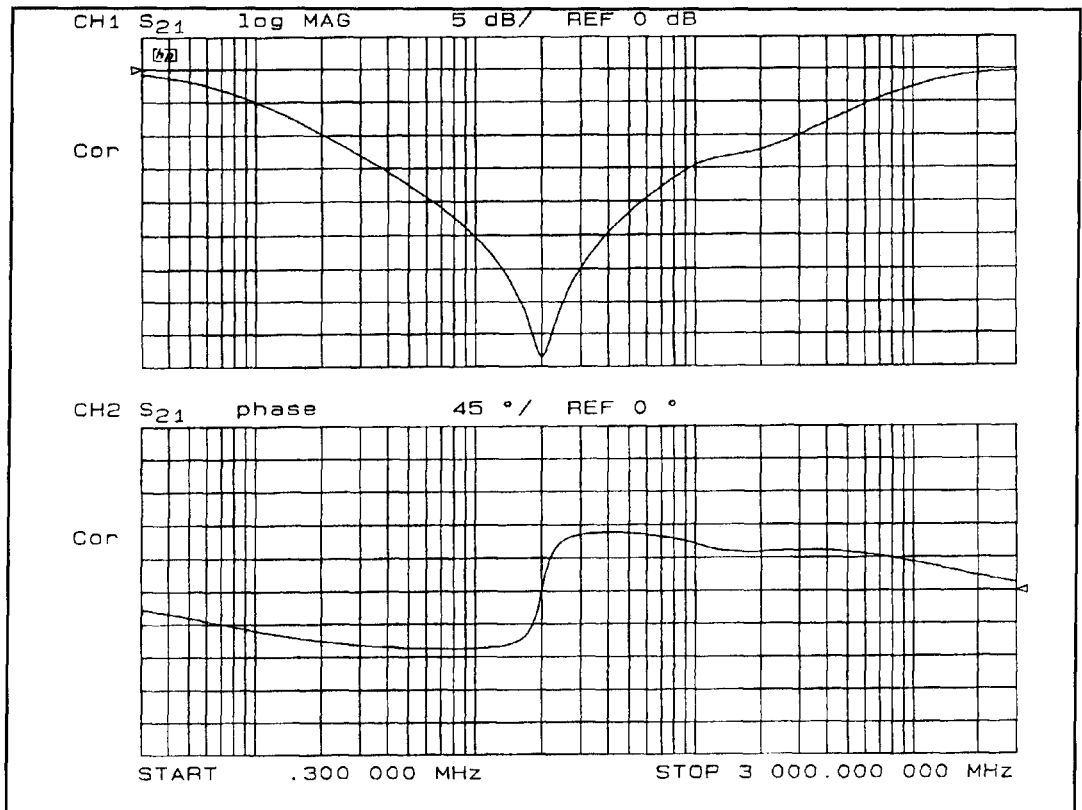


Figure 14. Response of a CM101 capacitor with series 15-ohm resistor in parallel with 0.01- $\mu$ F CK05 capacitor (minimum lead lengths).

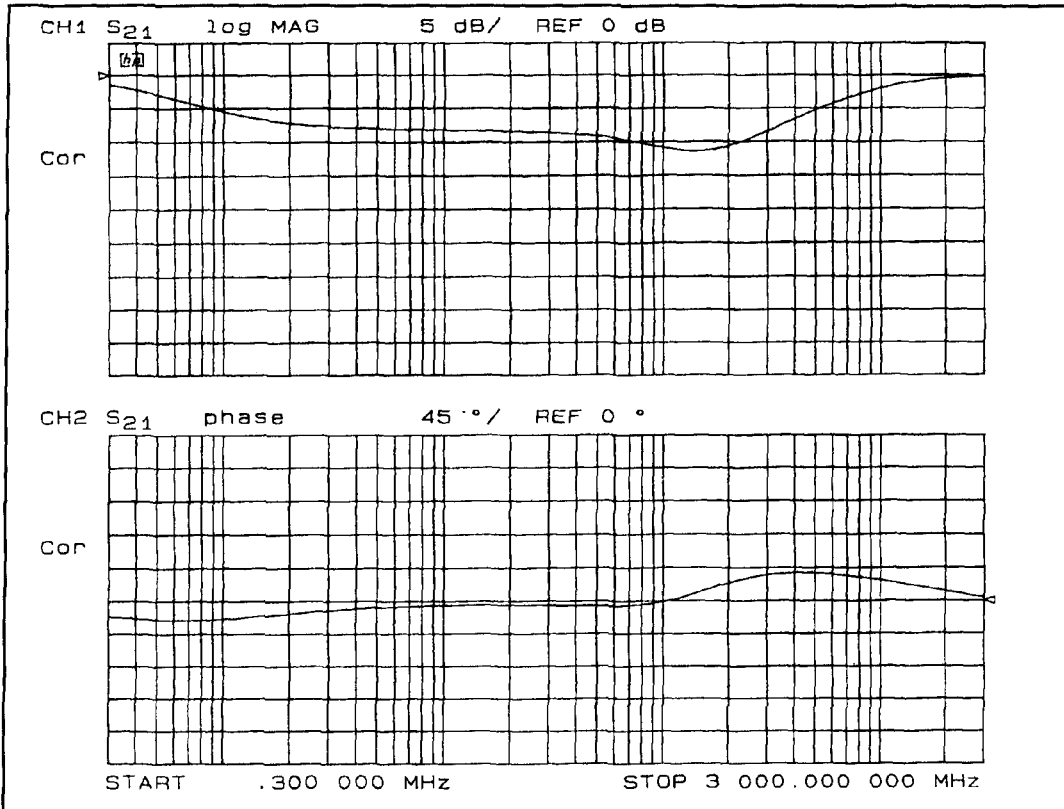


Figure 15. Response of a 100-pF CM101 capacitor in series with a 15-ohm RCR07 resistor in parallel with a 0.01- $\mu$ F CK05 capacitor in series with a 15-ohm RCR07 resistor (minimum lead lengths).

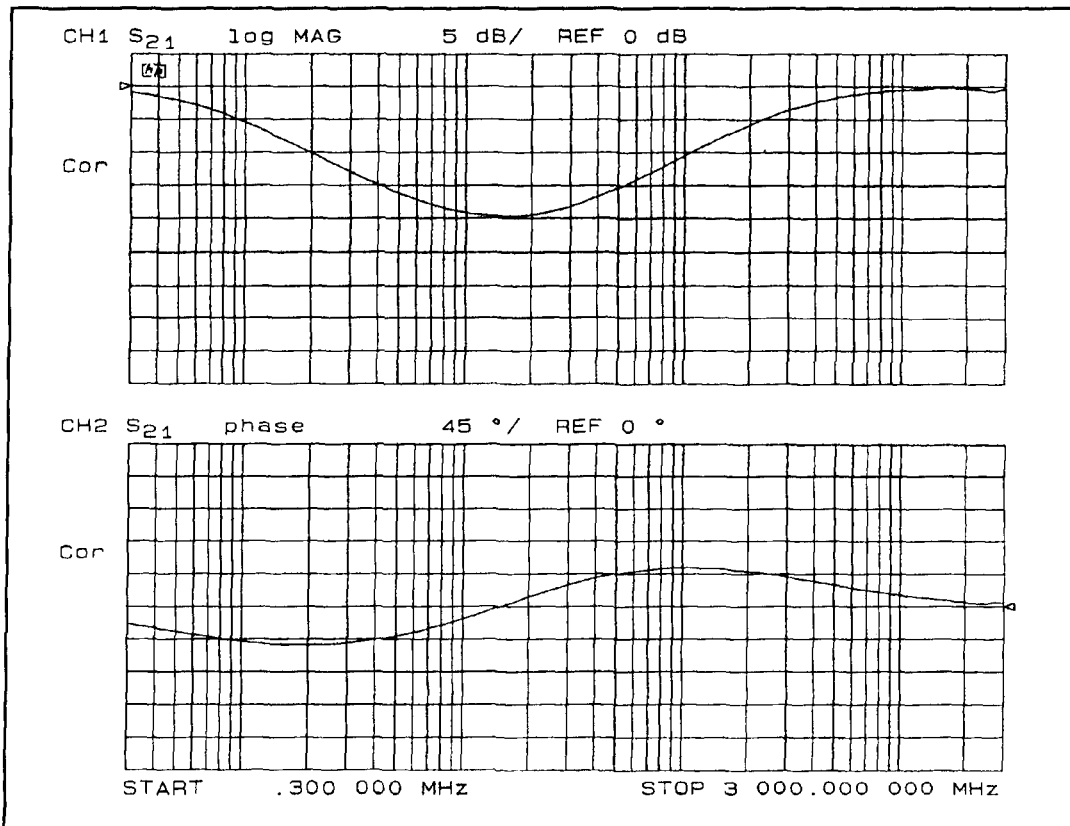


Figure 16. Response of a 0.01- $\mu$ F CK05 capacitor in series with a 2.7-ohm RCR07 resistor (minimum lead lengths).

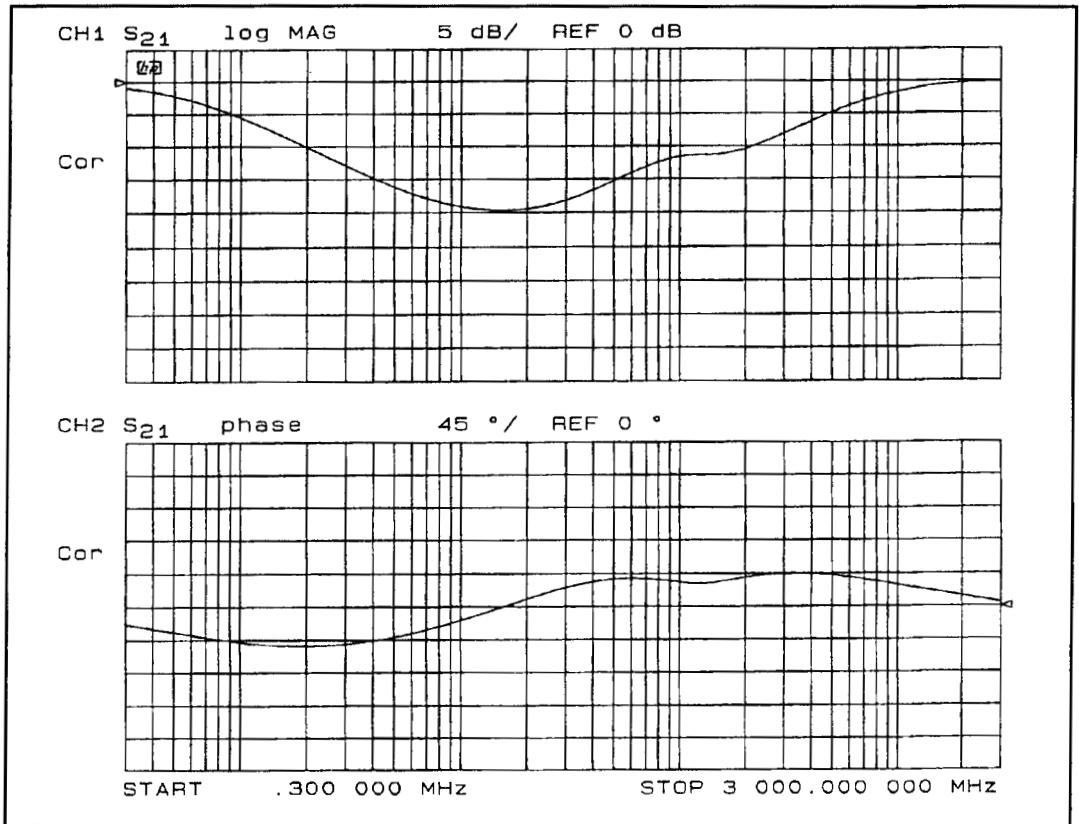


Figure 17. Response of a 100-pF CM101 capacitor in series with 15-ohms in parallel with 0.01- $\mu$ F CK05 in series with 2.7 ohms (minimum lead lengths).

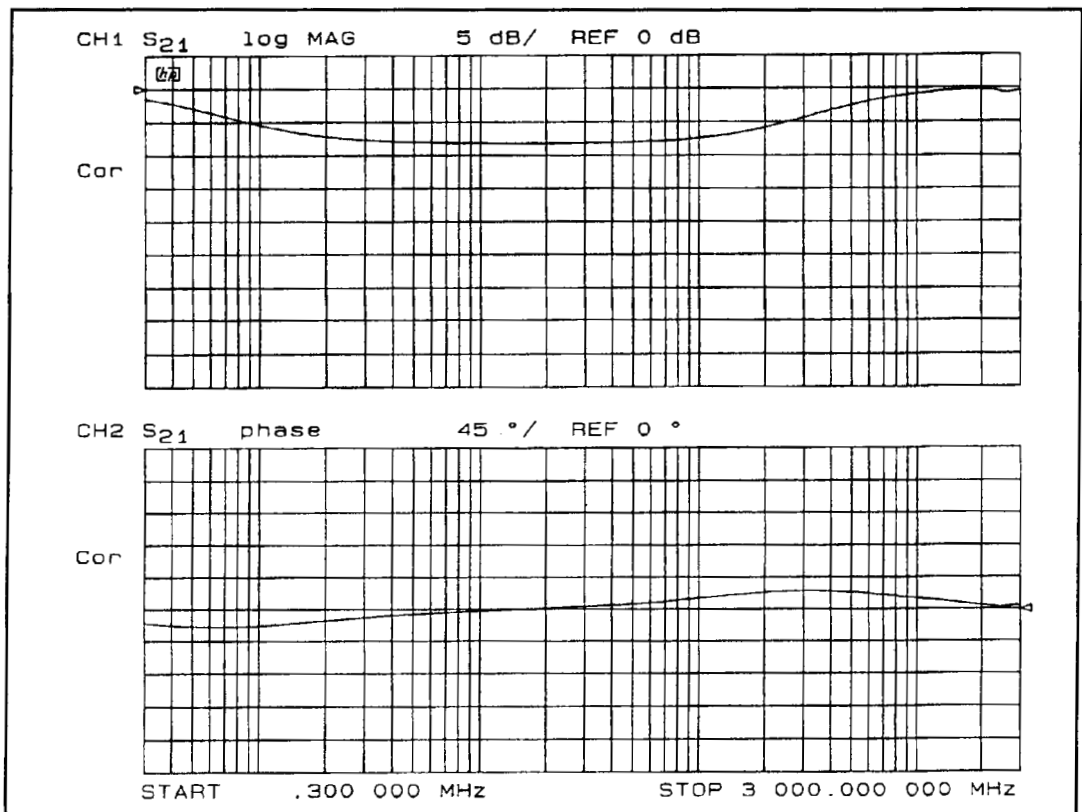


Figure 18. Response of a 0.01- $\mu$ F CK05 capacitor in series with a 15-ohm RCR07 1/4-watt resistor (minimum lead lengths).

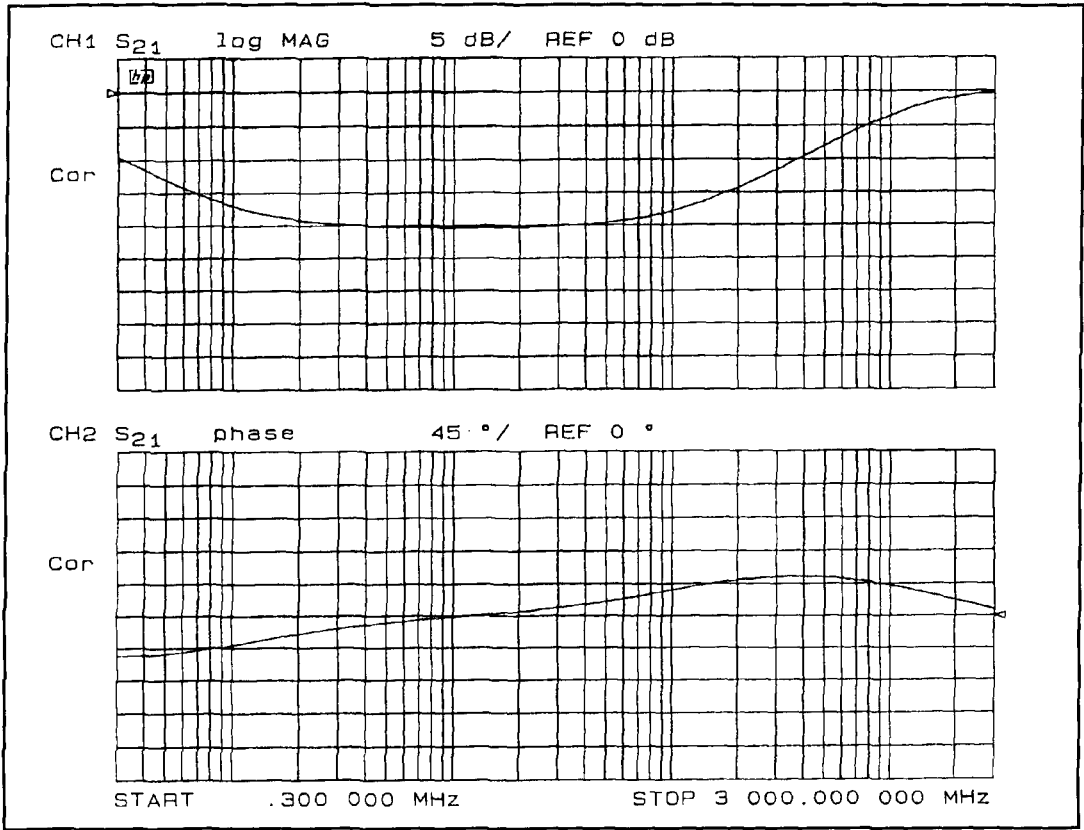


Figure 19. Response of six networks in parallel, each consisting of a 0.01- $\mu$ F CK05 capacitor in series with a 15-ohm RCR07 1/4-watt resistor (minimum lead lengths).

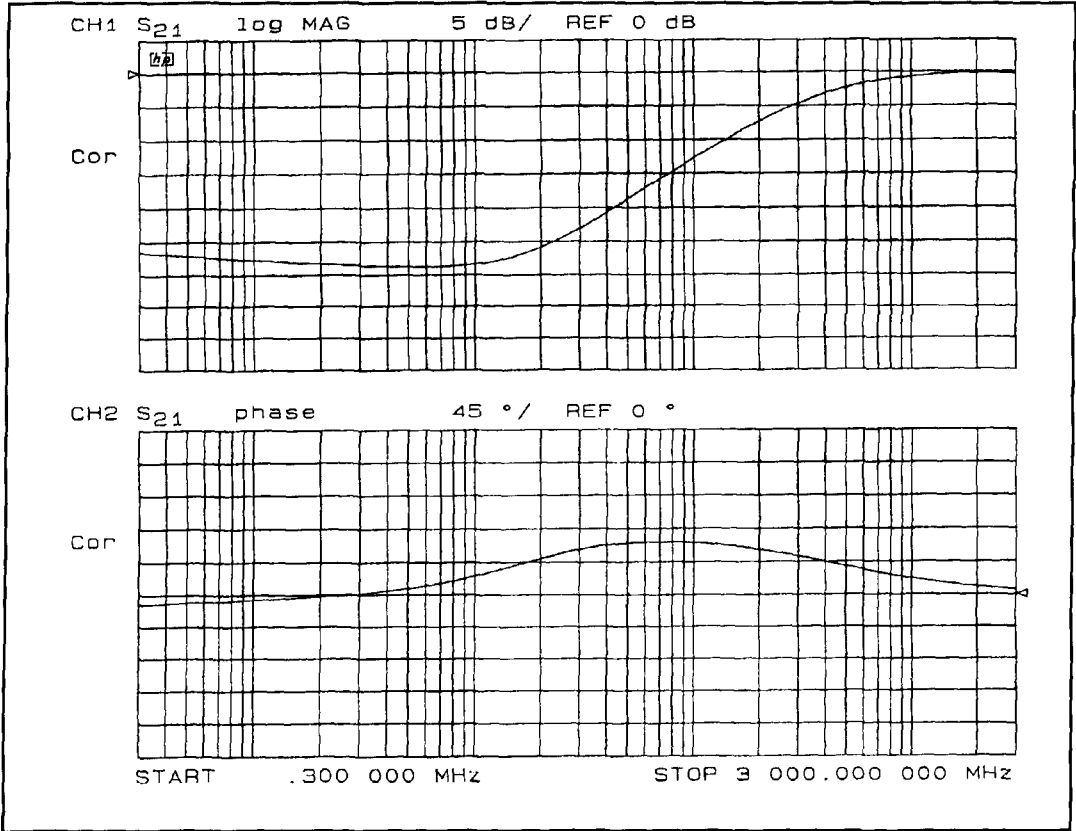


Figure 20. Response of a 10- $\mu$ F/35-volt 196-D capacitor (minimum lead lengths).

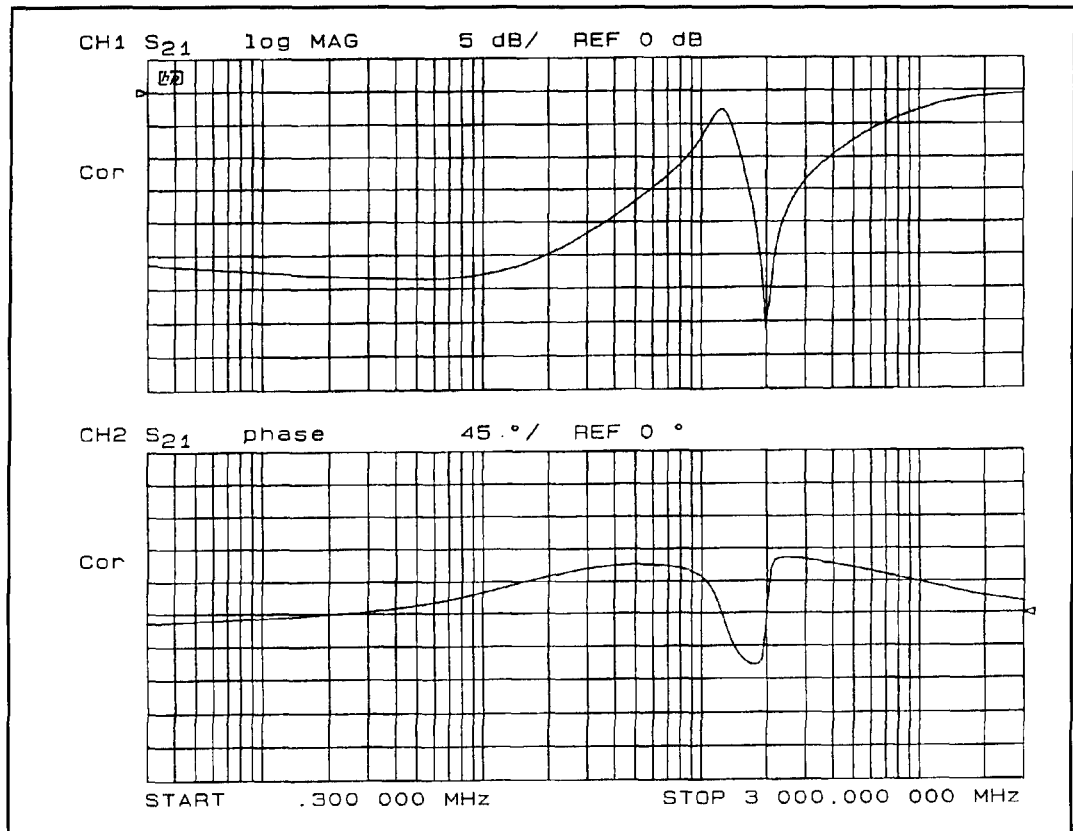


Figure 21. Response of a 10- $\mu$ F/35-volt 196-D capacitor in parallel with a 100-pF CM101 capacitor (minimum lead lengths).

paralleled by the 0.01- $\mu$ F CK05 capacitor. With only the 100-pF capacitor damped, the parallel resonance is eliminated, as expected, because it was the 100-pF capacitor that was forming a parallel resonance with the inductance of the larger capacitor. **Figure 15** shows the response of the 100-pF CM101 capacitor with a series 15-ohm resistor paralleled with the 0.01- $\mu$ F CK05 capacitor, and also with a 15-ohm series resistor. Compare this response with that of **Figure 9**. Note that the response is now quite well behaved, and all of the resonance originally seen with simple paralleling of the two capacitors is eliminated.

As noted above, a resistance of only about 1.3 ohm is required to damp the 0.01- $\mu$ F capacitor. The smallest resistor available at the time of this work was a 2.7-ohm RCR07 1/4 watt. **Figure 16** shows the response of the 0.01- $\mu$ F capacitor damped with 2.7 ohms. As expected, it is quite well damped. **Figure 17** shows the response of the 0.01- $\mu$ F capacitor damped with 2.7 ohms in parallel with the 100-pF capacitor damped with 15 ohms. Again, all resonances seen when the undamped capacitors were paralleled are completely eliminated when the capacitors are resistively damped.

A problem with damping bypass capacitors is that the amount of bypassing—that is, the

minimum impedance provided by the bypass capacitor—is limited by the series resistance added to damp the capacitor. For example, **Figure 18** illustrates the response of the 0.01- $\mu$ F capacitor over damped with 15 ohms. The over-damped 0.01- $\mu$ F capacitor provides only about a 15.5-ohm bypass impedance to ground at the midpoint in the response (about 2 MHz). However, because the damped capacitors may be successfully paralleled, the impedance may be reduced to any desired value (within reason) by simply paralleling several of the damped capacitors.

For example, consider that a nominal 0.05- $\mu$ F bypass capacitor with a 20-dB attenuation ratio is desired. The required equivalent series resistance must be about 2.5 ohms. A total of six of the 0.01- $\mu$ F capacitors, each with 15-ohm damping, would provide 0.06- $\mu$ F and 2.5 ohms series resistance. The response of such a composite bypass capacitor is shown in **Figure 19**. The important point to note in **Figure 19** is that with each of the capacitors well damped, paralleling of several capacitors is very effective in providing improved bypassing. The response of the parallel combination is very well behaved, exhibiting no series or parallel resonances.

The 0.01- $\mu$ F capacitor used above is rather small for most bypassing needs. A value of per-

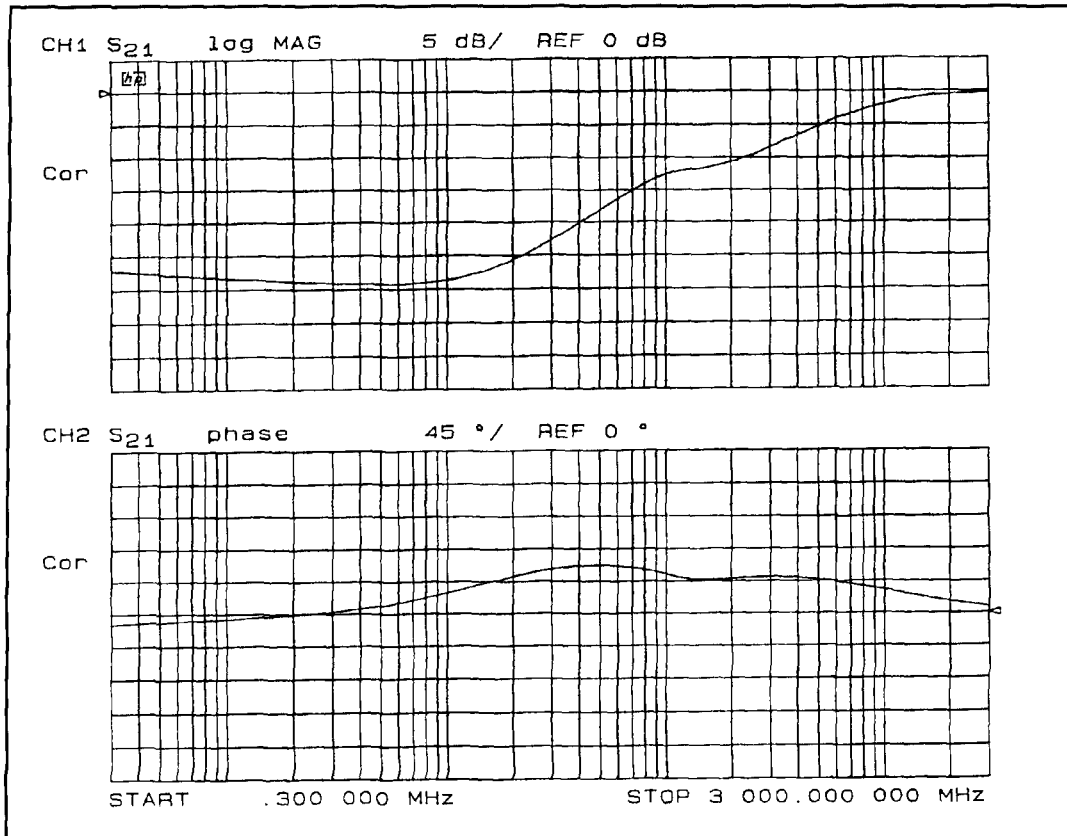


Figure 22. Response of a 100-pF CM101 capacitor in series with 15 ohms in parallel with a 10- $\mu$ F/35-volt 196-D capacitor (minimum lead lengths).

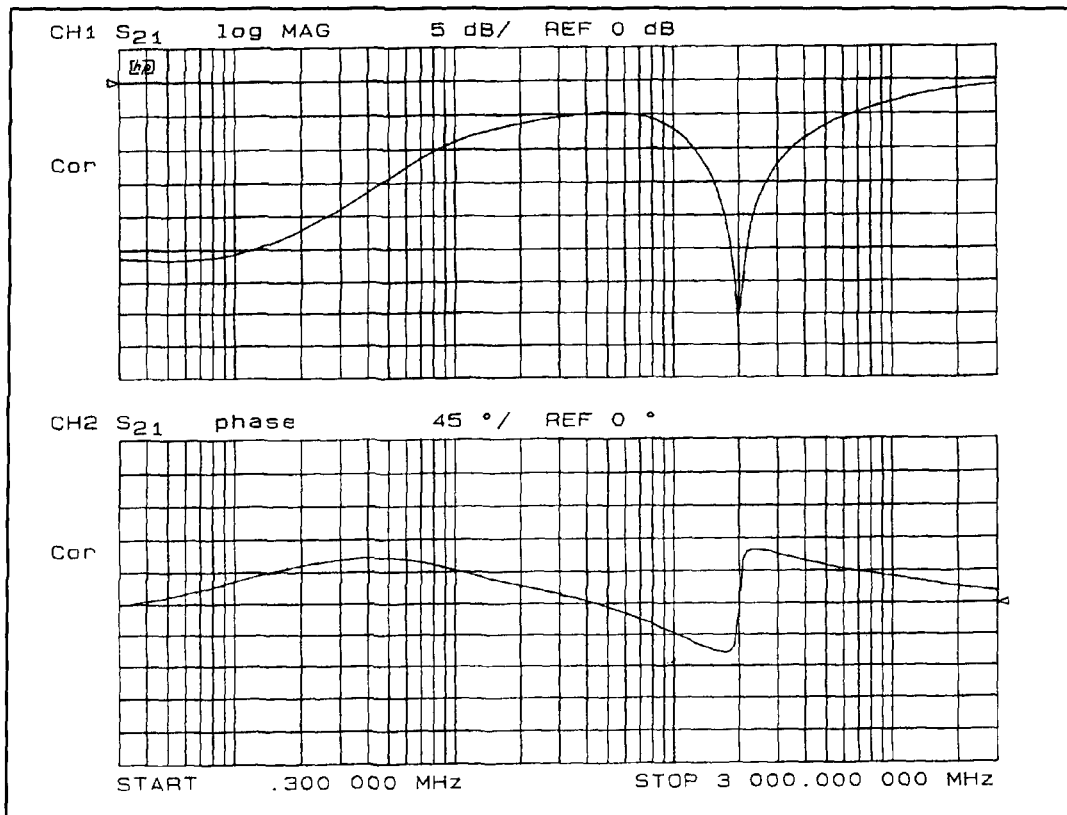


Figure 23. Response of a 10- $\mu$ F/35-volt 196-D capacitor with ferrite bead in parallel with a 100-pF CM101 capacitor (minimum lead lengths).

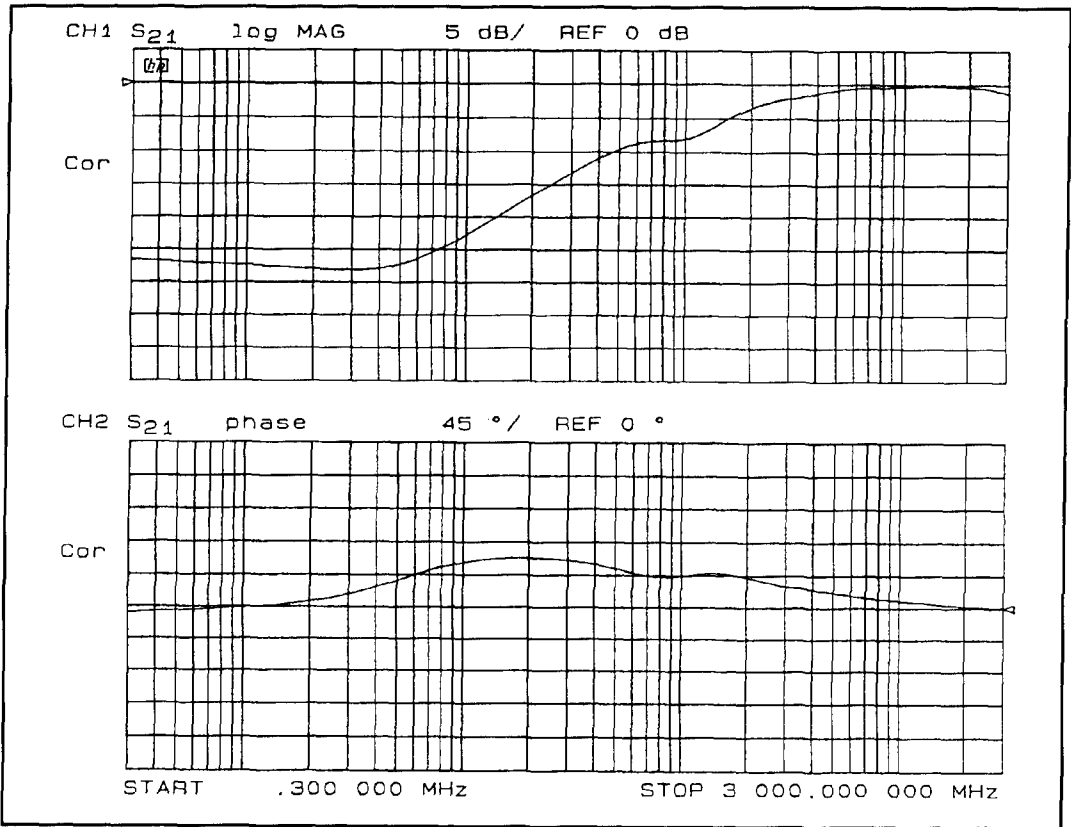


Figure 24. Response of a 10- $\mu$ F/25-volt type 30-D aluminum electrolytic capacitor (minimum lead lengths).

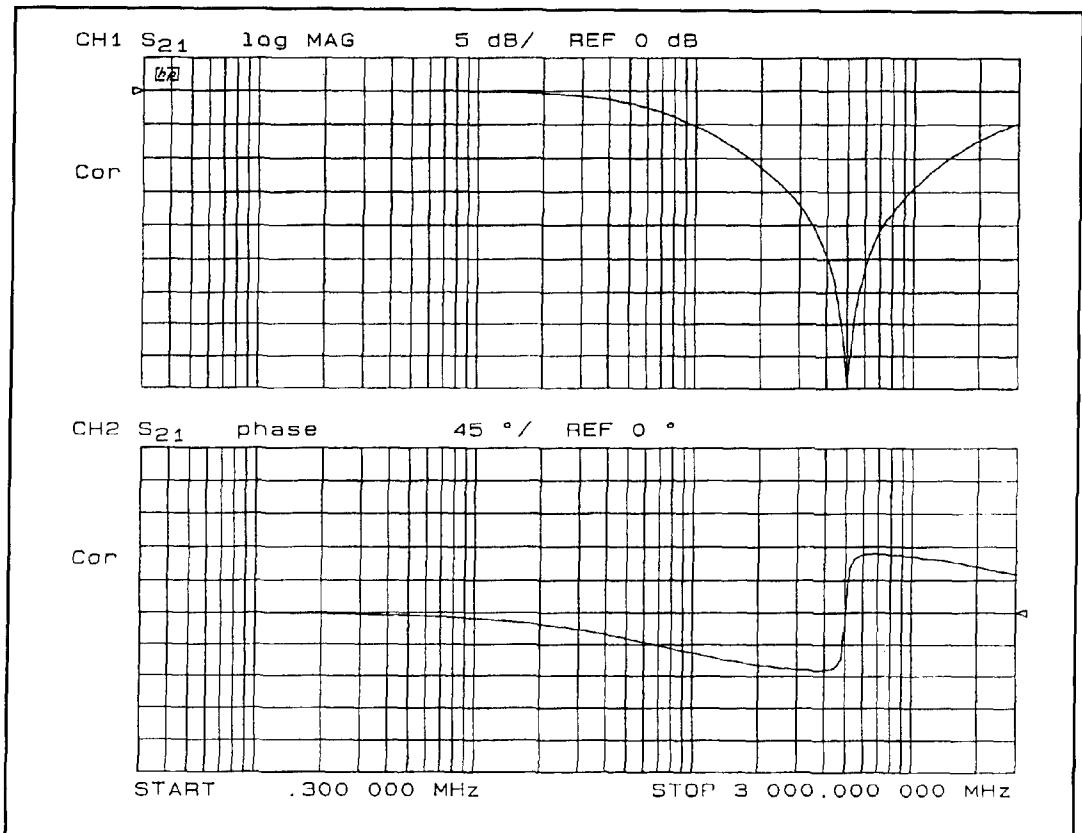


Figure 25. Response of a 100-pF ceramic chip capacitor (minimum lead lengths).



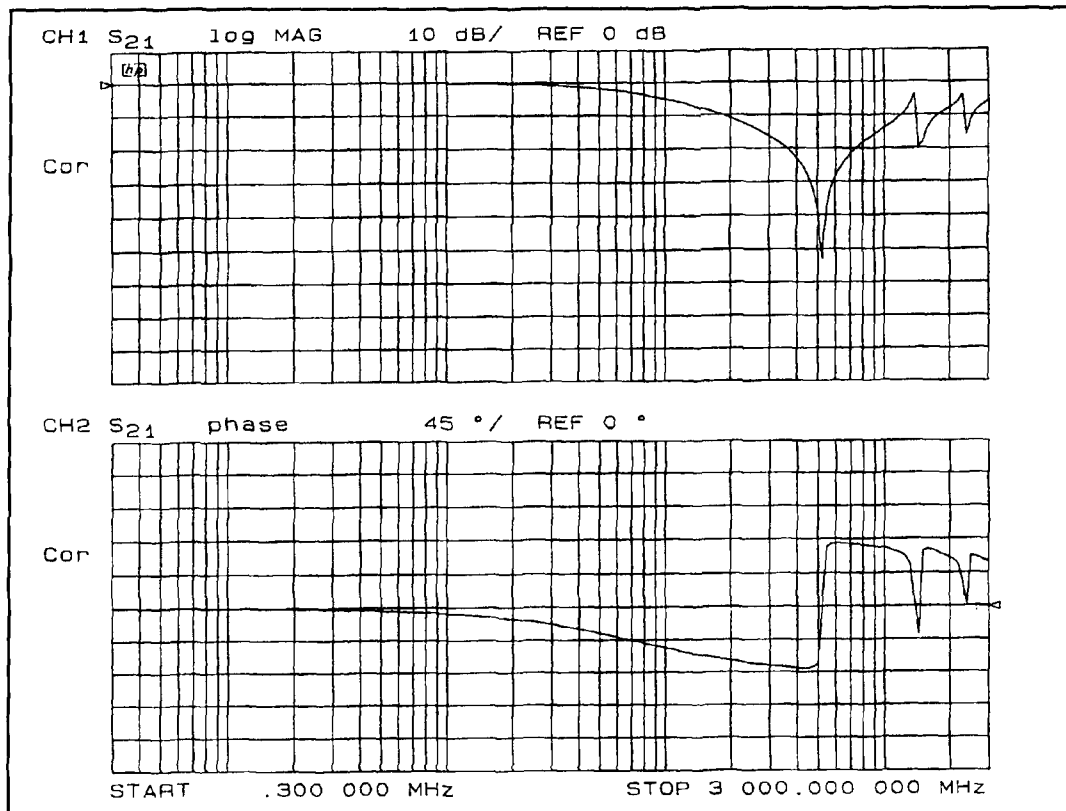


Figure 26. Response of a 100-pF porcelain chip capacitor (minimum lead lengths).

haps 10- $\mu$ F is more appropriate for general amplifier applications. **Figure 20** shows the response of a 10- $\mu$ F/35-volt 196-D dipped tantalum capacitor. The response of this unit is well behaved with the high-frequency performance limited by the inductance of the configuration. The response of **Figure 20** shows a highly damped response even without any additional resistance. The minimum in the magnitude response at the point where the phase is zero, i.e., the impedance is real, is about -28 dB at 2.8 MHz—so the equivalent series resistance is about 1 ohm. If the series-resonant frequency is taken as approximately 2.8 MHz, the resistance required for critical damping is only about 0.008 ohms. The internal losses of the 10  $\mu$ F 196 D then tend to make it self damping. However, because the internal losses cannot necessarily be accurately described as simple lumped elements of constant value over frequency, temperature, time, etc., one must still exercise care when paralleling these components.

**Figure 21** shows the response with the 10- $\mu$ F capacitor paralleled with the 100-pF capacitor, both with no damping. The parallel resonance is again seen. With only the 100-pF capacitor damped (with 15 ohms), the result is the response of the parallel combination shown in **Figure 22**. Because there are only two capacitors in this case, the single damping resistor is

sufficient to provide adequate damping.

Often a ferrite bead is used on the larger of two paralleled capacitors to control parasitic resonances. **Figure 23** shows the response of the 10- $\mu$ F capacitor in parallel with the 100-pF capacitor with a bead on the ungrounded lead of the 10- $\mu$ F capacitor, but without any other damping. Although the resonances are eliminated with the bead, very little bypassing is provided in the mid band ( $\approx -10$  dB). Nevertheless, the beaded configuration generally provides adequate bypassing for RF applications; however, the response of the parallel combination with the 100-pF capacitor damped with 15 ohms, as shown in **Figure 22**, provides much better bypassing over almost the entire useful range.

Tantalum capacitors, even in relatively large values, provide quite good RF performance, as shown in **Figure 20**. The larger physical size of other capacitors, like aluminum electrolytics, results in higher inductance in those units. **Figure 24** shows the response of a 10- $\mu$ F/25-volt type 30-D aluminum electrolytic. Although it too is relatively well behaved, it is limited to much lower useful frequencies than the tantalum. Also, at about 100 MHz, there is an anomalous response. This is not a particular concern, but does indicate that there are other parasitic elements in this unit that could cause unexpect-

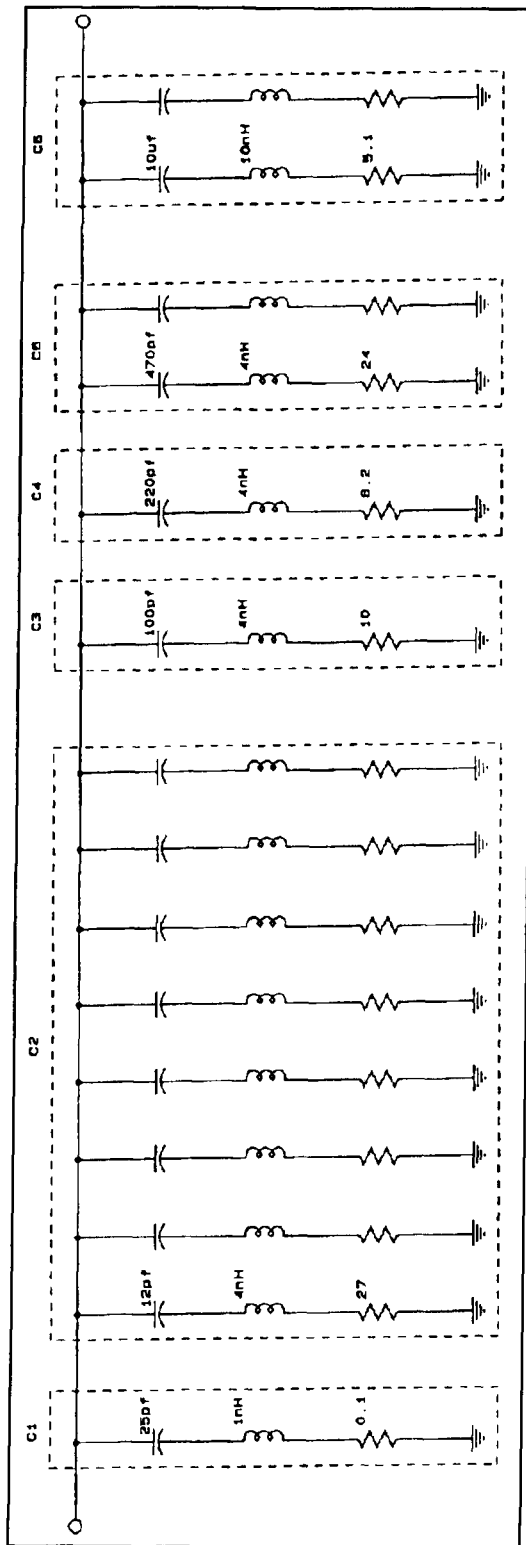


Figure 27. Composite bypass-capacitor schematic.

ed results if this capacitor were paralleled with other capacitors—particularly if no damping were employed.

### Performance of chip capacitors

It would be useful to very briefly look at the

performance of chip capacitors. **Figure 25** shows the response of an inexpensive 100-pF ceramic chip capacitor. The 100-pF value was selected to allow easy comparison with the performance of the CM101 silver mica unit analyzed above. This particular part is a general-purpose device and is not particularly specified for high-frequency service. Nevertheless, it is a well behaved unit. The series resonance is at about 500 MHz indicating an equivalent series inductance of about 1 nH, and the 45-dB resonant depth indicates an equivalent series resistance of about 0.14 ohm. This is certainly acceptable performance. **Figure 26** shows the response of a 100-pF high-performance porcelain RF chip capacitor (note the change in magnitude scale). The resonance is slightly higher than the general-purpose chip, indicating an equivalent series inductance of about 0.9 nH. This is due to the slightly larger cross section of the porcelain unit. The length of the two was nominally the same. The resonance depth is about 53 dB, indicating an equivalent series resistance of about 0.056 ohm. So, the high-performance porcelain chip capacitor exhibits almost the same equivalent series inductance as the much less costly general-purpose ceramic unit. However, the equivalent series resistance of the porcelain unit is almost one-third that of the ceramic unit. This comparison shows that the quality of the capacitor has little bearing on equivalent series inductance, but greatly affects equivalent series resistance.

The particular type of porcelain unit used for this analysis was selected to also show another anomalous response. Note the two additional resonances in **Figure 26** at about 1.5 GHz and 2.5 GHz. Because the actual construction of this capacitor is not known, the specific cause of these resonance responses cannot be stated with any real certainty. However, it is very likely that these are the result of high-Q piezoelectric resonant modes in the basic porcelain material. This particular unit would be a poor choice for use above about 1 GHz. The lower cost ceramic unit would actually be a much better choice for simple bypassing. However, the porcelain unit would be the better choice for high-Q resonators due to the much lower equivalent series resistance.

### Summary

To summarize this review of the bypass capacitor, a composite bypass capacitor was designed to provide a nominal bypass impedance of about 2.5 ohms or less (-20 dB in the test fixture) from about 10 kHz to 1 GHz. The schematic of the composite unit is shown in **Figure 27**. That configuration was constructed

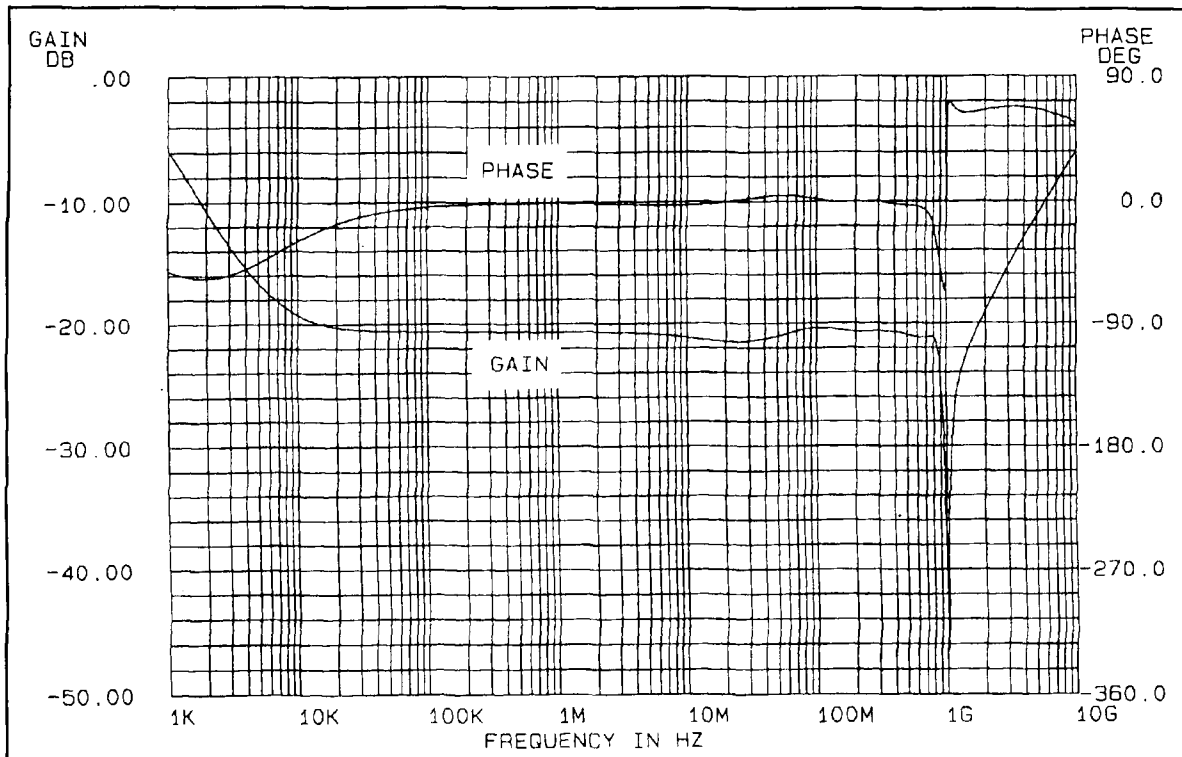


Figure 28. Composite bypass-capacitor response.

empirically using a CAE program. All capacitors except those of  $C_6$  are ceramic chip units. The two capacitors of  $C_6$  are 196-D tantalum units. The parasitic values used in the schematic are taken from the data presented above.

Capacitor  $C_1$  is undamped to allow it to provide maximum bypassing at its series-resonant frequency. That resonant frequency was selected to be at the upper frequency limit of the desired bypassing, 1 GHz. As stated, that capacitor is a chip capacitor with an equivalent series inductance of 1 nH and a series resistance of about 0.1 ohm, as was found for the typical chip capacitor investigated above.

Capacitor  $C_2$  is composed of eight chip capacitor-resistor networks in parallel. For this capacitance, a nominal 100-pF capacitance resonating at about 725 MHz and having about a 3-ohm resistance is required. However, the minimum inductance of a chip capacitor with a series resistor (1/4-watt unit) was found above to be about 4 nH to 1 nH for the capacitor and 3 nH for the resistor. Therefore, eight capacitors of 1/8th the total value are paralleled, each with a series resistor equal to about a factor of eight greater than the desired resistance. The total capacitance of  $C_2$  is then 96 pF, and the inductance about 0.5 nH. Note that the resonant frequency of 12 pF and 4 nH is the same as that of  $8 \times 12$  pF and  $4$  nH/8. At the series-resonant frequency, the eight 27-ohm resistors are paralleled, providing an effective series resistance of

3.3 ohms. The 27-ohm series resistor is slightly larger than that required to provide critical damping, so the composite should be slightly over damped.

Capacitor  $C_3$  provides a damped new graph-series resonance at about 250 MHz. Here a single capacitor and resistor may be used because the required inductance and resistance are within practical values available. The same is true of  $C_4$ , which provides a damped resonance at about 170 MHz.

For capacitor  $C_5$ , a nominal 1000-pF capacitor with about 2-nH series inductance and 12-ohm series resistance is required. Again, because the minimum practical inductance is 4 nH for the capacitor-resistor series combination, two parallel networks are required. However, here a series resistance of only about 4 ohms is needed to ensure adequate damping. The 24-ohm resistance provides a very high degree of over damping. If this configuration were actually used, a resistance of 5.1 ohms might be a good choice because it would provide improved bypassing. However, the purpose here is to demonstrate that a specific bypassing performance can be provided, so the 24-ohm value is used to provide the desired 2.5-ohm bypass resistance characteristic in the overall composite unit.

Finally, a nominal 20- $\mu$ F capacitor providing a damped resonance at about 500 kHz is required in  $C_6$ . Because the minimum practical

series inductance is about 10 nH, two capacitors must be used. Also, the 5.1-ohm resistance was selected to provide the desired 2.5-ohm overall resistance characteristic. If a specific resistance were not required, it is likely that no additional resistance would be necessary, because the inherent series resistance of the tantalum capacitor would very likely be sufficient to adequately damp the combination.

The response of the composite bypass capacitor is shown in the simulation of **Figure 28**. Note that the general response is very well behaved with no severe resonances except the series resonance of the undamped chip capacitor  $C_1$  at about 1 GHz. Particularly, there are no parallel-resonant modes that would compromise bypassing at the parallel-resonant frequencies. The overall response of the composite unit is a nominal -20 dB loss with 0-degree phase that corresponds to the 2.5-ohm resistive characteristic originally specified for this composite bypass capacitor.

In conclusion, all capacitors include parasitic

elements that must be considered in circuit design. When capacitors are used for bypassing in RF applications, the resonant circuits formed by the capacitors and the equivalent series inductance of the configuration result in both series and parallel resonances. Further, for typical RF capacitors, the equivalent series inductance is primarily a function of the physical configuration of the circuit and almost totally independent of the capacitor quality. When two or more capacitors are used in parallel, for broader frequency coverage in a bypassing application for example, parallel-resonant circuits will be formed resulting in virtually no bypassing at the parallel-resonant frequencies. That can result in severe instability in the device being bypassed. However, the performance of a composite bypassing network comprising several separate capacitors can be very significantly improved by ensuring that each capacitor is sufficiently damped. Such damping eliminates the severe parallel-resonant modes and provides a well-behaved bypassing characteristic. ■

## PRODUCT INFORMATION

### QUALCOMM Introduces DAC Series.

QUALCOMM Incorporated VLSI Products Group introduced the new Q2500 Series of Digital-to-Analog Converters (DACs). The QUALCOMM Q2500 DACs are for use with high speed direct digital synthesis (DDS) and other high speed applications. The Q2500 DAC series provides same-source building blocks for the communications system designer.



The Q2510 and Q2520 devices offer 10- or 12-bit digital inputs, respectively.

The 10-bit, high-speed Q2510 DAC is TTL-interface compatible and allows update speeds to 100 Msps. The Q2510 is available in 28-pin PDIP and SOIC packages, operating over the  $-25^{\circ}\text{C}$  to  $+85^{\circ}\text{C}$  temperature range.

The 12-bit, high-speed Q2520 is also TTL compatible and will update to 80 Msps. The QUALCOMM Q2520 is available in 28-pin PDIP and PLCC packaging with an operating temperature range of  $-25^{\circ}\text{C}$  to  $+85^{\circ}\text{C}$ .

Prices for the Q2510 start at \$38.10 (quantity 1000). The Q2520 is available for \$34.50 (quan-

tity 1000). For more information call 619-587-1121; FAX 619-452-9096; or write QUALCOMM, 10555 Sorrento Valley Road, San Diego, CA 92121-1617.

### Microwave Tuning Elements added to Sprague-Goodman line.

Sprague-Goodman Electronics, Inc. has introduced a new line of Microwave Tuning Elements.

Six variations are available 1) Metallic tuning elements, 2) Dielectric tuning elements, 3) Metallic and dielectric tuning rotors, 4) Dielectric resonator tuners, 5) LC tuning elements, and 6) Resistive tuning elements.

For more information, contact Sprague-Goodman Electronics, Inc., 134 Fulton Avenue, Garden City Park, NY 11040-5395; Phone: 516-746-1385, FAX: 516-746-1396.



# QUARTERLY DEVICES

## *The Motorola MC13175/6 UHF FM/AM transmitter*

---

FCC Title 47, Part 15 opened the door for development of many short-range communication and remote-control devices operating in the UHF region. Along with Part 15 comes a host of new low-cost RF products that show promise as building blocks for more serious projects. For example, this edition of “Quarterly Devices” focuses on Motorola’s MC13175/6 UHF transmitter chips. Although designed for low-power unlicensed consumer products, they may also serve as an exciter/modulator or local oscillator (LO) for much higher-power systems.

### Overview

The MC13135/6 series is versatile—capable of internally generating analog AM and FM signals, as well as Amplitude Shift Keying (ASK) for data. This “multi-mode” capability makes these chips useful in a wide range of applications, including voice, video, data, and spread spectrum transmitter projects.

Perhaps the most interesting feature is the chip’s built-in PLL system which establishes operating frequency. Instead of multiplying up in traditional fashion, the MC13175/6 uses a free-running CCO (current-controlled oscillator) that routes a sample of its output signal through a divider network. The divided-down signal is then compared to a low-frequency on-board crystal-controlled reference oscillator in order to generate the required error signal for PLL operation. This scheme provides crystal-controlled frequency stability at the UHF output frequency without the need for discrete multiplier stages.

The primary difference between the MC13175 and MC13176 lies in the internal divider network; the MC13175 divides its output signal down by 8, while the MC13176

divides down by 32. This means that the MC13175 operates at 8 times its crystal reference frequency, and the MC13176 operates at 32 times its referent. To generate output on 432 MHz with the MC13175, you would use a 54-MHz third-overtone crystal ( $54 \times 8 = 432$ ). The MC13176 does the same job with a less-expensive 13.5-MHz fundamental-mode crystal ( $13.5 \times 32 = 432$ ). Thus, for higher-frequency applications, the MC13176 is the more logical choice of the two.

Although designed for 260-470 MHz Part 15 applications, the MC13176 will generate fundamental output well into the 900-MHz range, making it useful as a simple and cost-effective LO for receivers and transverters operating in this region. It is also possible to design simple LOs for 1296 MHz by running the device at 600-MHz and doubling the output. On the other end of the spectrum, the low-frequency hold-in limit of the device’s CCO is around 200 MHz.

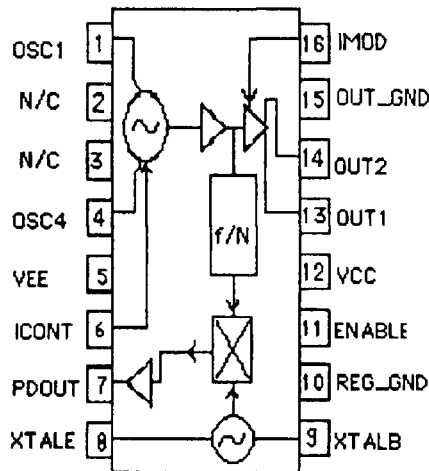
The MC13175/6 is very easy to power, operating on any supply voltage between 1.8 to 5.0 volts DC with a maximum current drain of 35 mA. There’s also a power-down mode for standby periods. RF power output is adjustable to +10 dBm—providing a respectable drive level for higher power RF strips. The chip uses differential output, which is convenient for direct connection to loop antennas in consumer-type products. A balun-type matching network is recommended when the device is used to drive subsequent stages.

Nearly all the required circuitry needed to support operation is on-board the chip, keeping the external parts count low and holding pc-board landscape to a minimum. In most applications, simple additions like a reference crystal, a couple of tuned circuits, a loop filter, and possibly a microphone preamplifier are all you need. The MC13175/6 comes in SO-16 surface-mount dual-inline plastic packaging, which is



**D SUFFIX**  
**PLASTIC PACKAGE**  
**CASE 751B**  
**SO-16**

**PIN CONNECTIONS AND  
 FUNCTIONAL BLOCK DIAGRAM**



superior to standard P-type dip packaging at the operating frequencies involved.

**A look inside**

Refer to **Figure 1** for an internal functional block diagram and electrical characteristics. In a typical application, the CCO oscillator tank coil is connected between pins 1 and 4 (pins 2 and 3 are unused). Parallel capacitance for the oscillator tank is supplied internally by CCO circuitry and by parasitic capacitances on the pc board. Vee (supply ground) is connected at pin 5. Current inputs for controlling oscillator frequency are supplied to pin 6. These are normally supplied from two sources: from the chip's phase detector which has its output at pin 7, and from an external speech-amp circuit when FM modulation is used. In most applications, a simple loop filter is all that is required between pins 6 and 7 to stabilize and filter the CCO loop response. The reference oscillator crystal—and its associated feedback divider—are connected between pins 8 and 9. Vcc (supplied to pin 12) is decoupled for RF at pin 10. Pin 11 is the enable/power down input. Differential RF out-

**Figure 1. Internal functional block diagram and electrical characteristics of the MC13175/6.**

**MAXIMUM RATINGS** ( $T_A = 25^\circ\text{C}$  unless otherwise noted).

Ratings	Symbol	Value	Unit
Power Supply Voltage	VCC	7(max)	Vdc
Operating Supply Voltage Range	VCC	1.8 to 5.0	Vdc
Junction Temperature	$T_J$	+150	$^\circ\text{C}$
Operating Ambient Temperature	$T_A$	-40 to +85	$^\circ\text{C}$
Storage Temperature	$T_{stg}$	-65 to +150	$^\circ\text{C}$

**ELECTRICAL CHARACTERISTICS** (Figure 2;  $V_{EE} = -3.0\text{ V}$ ,  $T_A = 25^\circ\text{C}$ , unless otherwise noted).

Characteristic	Symbol	Pin	Min	Typ	Max	Unit
Supply Current (power down: $I_{11}$ & $I_{16} = 0$ )	$I_{EE1}$		-1	—	—	$\mu\text{A}$
Enable Supply Current ( $I_{16} = 0$ )	$I_{EE2}$	—	—	-10	—	mA
Total Supply Current (Transmit Mode) ( $I_{mod} = 2\text{ mA}$ ; $f_o = 320\text{ MHz}$ )	$I_{EE3}$	—	—	-35	—	mA
Output Power ( $f_o = 320\text{ MHz}$ ; $V_{REF} = 500\text{ mV}_{p-p}$ ; $f_o = N \times f_{REF}$ ) $I_{mod} = 2\text{ mA}$ (See Figure 7 & 8) $I_{mod} = 0\text{ mA}$	$P_{out}$	13,14	—	4.7 -40	—	dBm
Hold-in Range ( $\pm\Delta f_{REF} \times N$ ) MC13175 (See Figures 7) MC13176 (See Figures 8)	$\pm\Delta f_H$	13,14	—	11 11	—	MHz
Phase Detector Output Error Current	$I_e$	7	—	30	—	$\mu\text{A}$
Enable Settling Time (See Figure 22b)	$T_{ENABLE}$	11,6	—	20	—	$\mu\text{sec}$
Amplitude Modulation Bandwidth (See Figure 24)	BWAM	16	—	25	—	MHz
Spurious Outputs (on)	$P_{son}$	13,14	—	-50	—	dBc
Spurious Outputs (off)	$P_{soff}$	13,14	—	-50	—	dBc
Maximum Divider Input Frequency	$f_{div}$	—	—	950	—	MHz
Maximum Output Frequency	$f_{out}$	13,14	—	950	—	MHz



put appears at pins 13 and 14, and output ground (an additional Vee) is connected to pin 15. Finally, transmitter power output (or the AM modulation waveform) is controlled via pin 16.

Motorola has prepared a well-detailed prototype information booklet for the MC13175/6. This provides a far more inclusive pin-by-pin description of the device and its internal circuitry, plus all the information and formulas you need to implement various modulation modes and special functions. The prototype information document is "must" reading for anyone contemplating original designs.

## A typical circuit

**Figure 2** shows a typical application of the MC13176, functioning as an FM exciter on 320-MHz. An inexpensive 10-MHz computer-clock reference crystal is used to control the transmitter's frequency of operation ( $10 \text{ MHz} \times 32 = 320 \text{ MHz}$ ). A  $6.8 \mu\text{H}$  inductor in series with the crystal creates a VXO circuit that extends the reference oscillator's frequency-

netting range. FM audio (voice or data) is routed to the frequency-control input at pin 6, where it is combined with the PLL correction signal from pin 7 (note the simple RC loop filter). The CCO tank coil is slug-tuned to bring the oscillator into PLL locking range. Transmitter output level is set via the 1.1 k resistor on pin 16, and the device is set in the enable mode by supplying current through the 27 k resistor at pin 11. A simple coaxial balun is used to unbalance transmitter output for feeding into a 1/4-wave rod antenna.

The MC13175/6 application note outlines sample designs for AM, FM, and ASK applications, as well as suggested circuits for inputting audio and data signals. It also details loop filtering requirements and provides an external loop amplifier circuit for wide-range PLL lockup.

## For more information...

If you are interested in finding out more about the MC13175/6 UHF FM/AM transmitter chip contact your local Motorola sales office, or call the Motorola Design Line at 1-800-521-6274. ■

## PRODUCT INFORMATION

**Digital Radio Systems, Inc.'s DPK-9600 TNC**  
DRSI introduces the DPK-9600 High-Speed Packet Radio TNC. The DPK-9600 features a single-chip Direct FSK modem which can be jumpered for 4800, 9600 or 19,200 baud data rate over-the-air and is compatible with K9NG and G3RUH modems.



The DPK-9600 provides 100% code compatibility for any TNC-2 clone firmware and is supplied with version 1.1.8a firmware in the U.S. and The Firmware 2.4c with DAMA for export. It will run all network PROMs with no modifications. The DPK-9600 is 5" wide x 6" deep x 1.25" high and weighs 14 ounces. It uses 50 milliamps at 9-20 volts DC. Front panel LED's can be disabled, reducing power consumption to approximately 35 ma. The DPK-9600's retail list price is \$249.95.

For more information, contact Digital Radio Systems at 2065 Range Road, Clearwater, FL 34625, or call 813-461-0204, or FAX 813-447-4369.

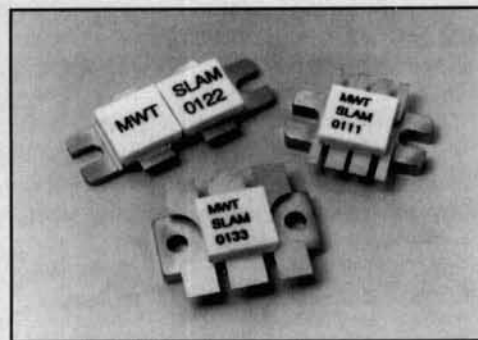
## Microwave Technology, Inc. expands its "SLAM" series

Microwave Technology, Inc. has expanded its "SLAM" series of ultra-linear, Class A power modules operating in the HF band from 2 to 32 MHz.

These modules are self-biased, internally matched and temperature compensated, and require no external ballast resistors or other temperature compensating circuitry.

The SLAM-0122 is a push-pull device that operates at 50 Watts with 14 dB typical gain. It requires a single 40 volt power supply. The SLAM-0133 is a single-ended device with 10 watt power output with 13.5 dB gain and requires a 32 volt supply.

For more information, contact MwT at 4268 Solar Way, Fremont, CA 94538. Telephone: 510-651-6700.





# SPREAD SPECTRUM COMMUNICATIONS

## *A historical and technical overview*

---

As we all know, the RF spectrum is a finite and exceedingly valuable resource. The practical limits of the existing spectrum, together with the exponential need for new communications, has created an increasing demand for what little free or underutilized space remains. In general, the policy for traditional commercial services has been to place limits on RF power and bandwidth to minimize interference and maximize the number of channels that can be assigned to a given band. For example, although the FCC continues to license a few “clear channel” AM broadcast stations, most frequencies are assigned to multiple stations. The nature of the service, local and regional geography, as well as signal propagation qualities at the operating frequency define the minimum physical spacing between stations sharing the same frequency. The interstation distance ranges from a few hundreds of meters for low-power cellular radio nodes to hundreds or thousands of kilometers for high-powered commercial broadcast stations.

Although congestion may not be obvious to the casual listener on the AM or FM broadcast bands, the current explosion in cellular telephone systems and wireless computer networks is a driving force behind the quest for more communications channels.<sup>1</sup> These and other new technologies that rely on relatively low-power, high-density environments operating at UHF and above suggest that new ways of managing signal bandwidth are critical to maximizing spectrum utilization.

In general, the type of modulation and information filtering used determine the RF signal bandwidth requirements. For example, where a properly filtered voice modulated AM signal requires a bandwidth of about twice that of the modulating frequency, SSB and other techniques can be used to reduce the RF bandwidth requirements significantly. On the amateur

bands, the use of SSB instead of AM has resulted in tremendous spectrum savings, as well as increased efficiency of communications.

However, despite the savings of spectrum by SSB, the bands remain crowded. The introduction of the newer spread spectrum techniques can further increase efficiency of spectrum use.

Paradoxically, techniques such as spread spectrum are designed to increase spectrum utilization by generating an extremely wide signal relative to the information to be transmitted. In conventional RF communications, the amplitude, frequency, or phase of an RF carrier is varied in accordance with the information to be transmitted. With the exception of those that use FM with a high modulation index, the bandwidth requirements of these systems are simply a function of the information bandwidth.

In contrast to conventional narrowband communications schemes, spread spectrum systems take voice or other relatively narrow-bandwidth information and distribute it over a band that may be several MHz wide. This distribution or “spreading” is accomplished by modulating the information to be sent with a wideband signal. Unlike conventional signals, spread spectrum signals occupy a bandwidth at least an order of magnitude greater than the information bandwidth.<sup>2</sup> The other distinguishing feature of spread spectrum communications is that a function other than the information to be sent (i.e., a wideband spreading signal) determines the RF bandwidth.<sup>3</sup>

Spread spectrum techniques can be used to create additional information channels that can coexist in a band filled with conventional narrowband communications. In addition, spread spectrum systems have characteristics uniquely suited for applications requiring privacy, signal covertness, interference rejection, ranging measurements, selective addressing, and multiple access.

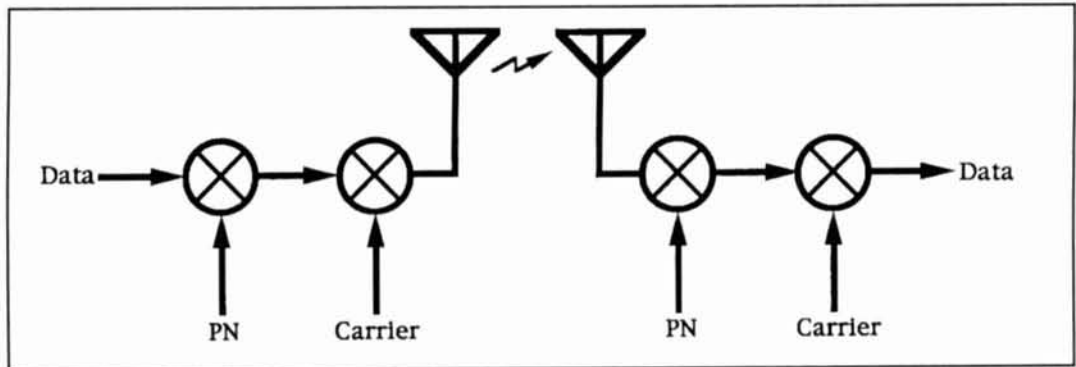


Figure 1. Overview of a direct sequence communications system. Data: digital voice or other information; Carrier: RF source; PN: pseudo-noise source.

In the amateur bands, present law permits the use of spread spectrum only above 420 MHz. Because the assigned bands are fairly narrow, a spreading system could be developed to use the entire band.

### The development of spread spectrum

Despite the recent flurry of interest in spread spectrum communications for mobile, computer, and personal communications, the technology is actually well over four decades old. The first commercial spread spectrum communications system in this country, developed around 1949, linked New Jersey and California.<sup>4</sup> Within a few years, the Army Signal Corps commissioned the development of a long-range, jam-proof HF radio teletype link. The Navy, Air Force, and National Security Agency soon followed the Army's lead by developing spread spectrum communications systems of

their own. Given the strategic nature of the technology, spread spectrum work was classified, and, until the past few years, research and development efforts in the field were withheld from the public.

The features of spread spectrum communications that stirred original military interest included resistance to jamming, low detectability, and low interference to other systems operating on the same band. Unlike the conventional techniques used to ensure message security, such as cryptography, spread spectrum communications can not only conceal the message, but the sender's identity and location. This ability to create a high level of uncertainty at the "unfriendly" receiver concerning the signal frequency, phase, and transmission time can be attributed in part to the signal spectrum, which can be made to appear as ordinary broad-band noise.

Despite the intense military interest in spread spectrum technology, the early systems were for the most part impractical and expensive prototypes. Like the first computers, the initial

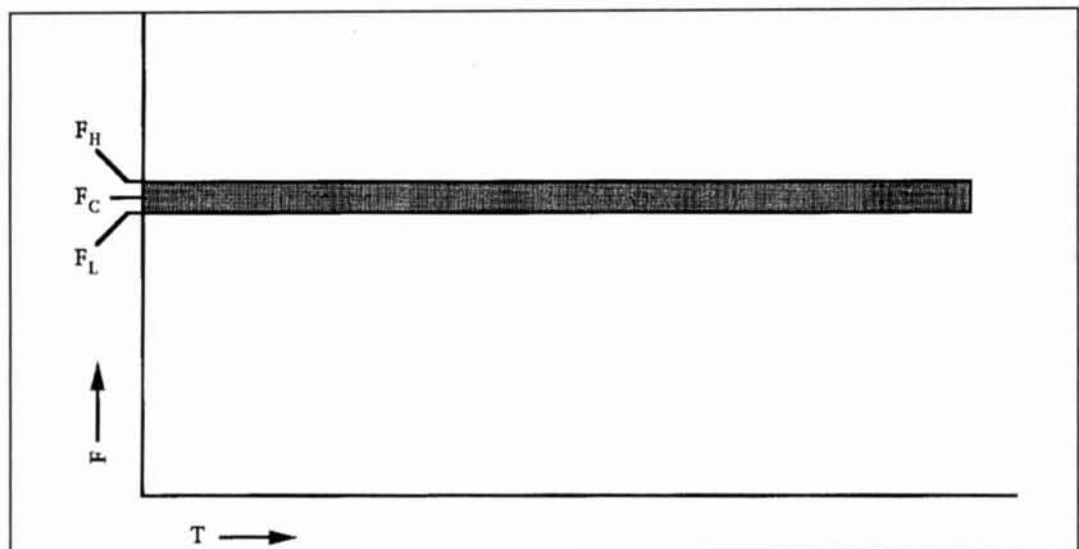


Figure 2. Spectral occupancy versus time characteristics of a conventional amplitude modulated (AM) signal. Bandwidth ( $F_H - F_L$ ) is a function of the modulation frequency,  $F_C$ : signal center frequency;  $F_H$ : signal high frequency limit;  $F_L$ : signal low frequency limit.

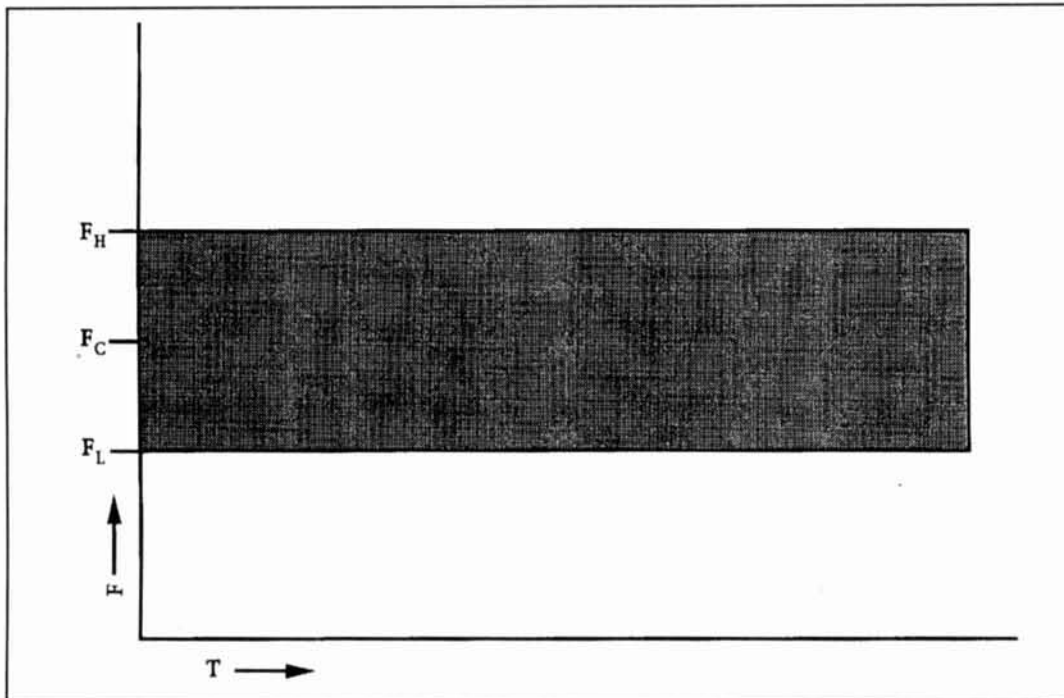


Figure 3. Spectral occupancy versus time characteristics of a direct sequence (DS) signal. Bandwidth ( $F_H - F_L$ ) is a function of the spreading signal, typically a pseudo-noise source. Because of spreading, the signal bandwidth is much larger than the information bandwidth. Compare with Figure 2.

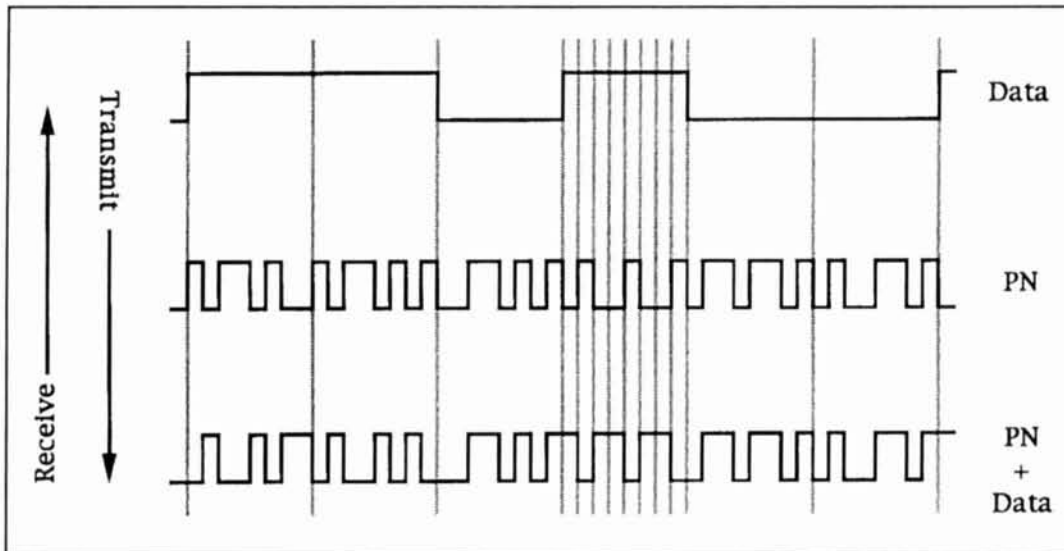


Figure 4. Generating a direct sequence signal involves the combination of a pseudo-noise source (PN) with the data to be transmitted (data). In this example, combination involves the exclusive-OR sum of the data and the PN source. On receive, a PN sequence identical to that used in transmit is XORed with the spreading signal to extract the data. In this example, the PN source is clocked at eight times the data rate; i.e., there are eight chips per data bit.

spread spectrum systems relied on rooms full of power-hungry and heat-producing vacuum tubes. Practical spread spectrum became a reality only with the advent of the transistor and integrated circuits. The lower cost afforded by inexpensive digital circuitry, together with the lifting of security restrictions, has made spread

spectrum a viable alternative for many commercial applications.

### Spread spectrum applications

Modern military and commercial applications of spread spectrum range from satellite

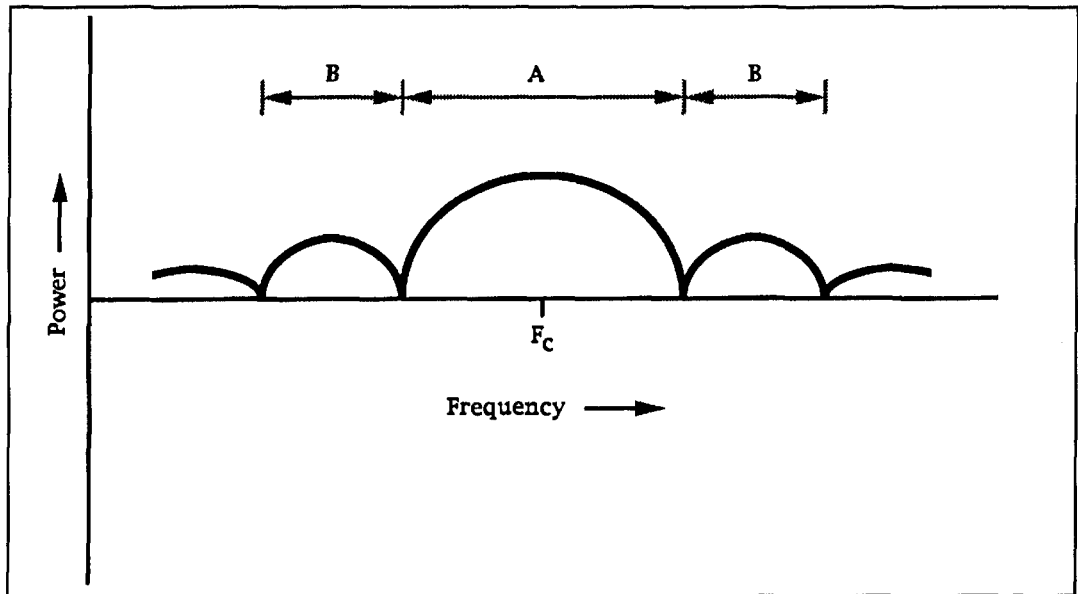


Figure 5. The power spectrum for a binary phase shift keying (BPSK) DS signal in which the carrier has been suppressed. The envelope of the distribution is primarily a function of the chip rate. The main lobe (A) is  $1/2$  of the chip rate wide, centered around the carrier frequency,  $F_C$ . Approximately 90 percent of the signal power lies in the main lobe; the side lobes (B) contain only about 10 percent of the total signal power.

and terrestrial communications to navigation and radar. Spread spectrum technology continues to play a central role in military communications because spread spectrum signals can be made to be resistant to jamming and interference, are difficult to detect and intercept, can provide message privacy, and can support multiple, selectively addressed users.

Spread spectrum ranging and radar systems use the spreading signal to determine signal transmit and receive times precisely. For example, the Global Positioning System (GPS) uses spread spectrum technology to allow personnel with military GPS receivers to determine their location within about 1 meter (commercial receivers are handicapped to a resolution of about 100 meters). The GPS system is based on 21 satellites, each transmitting on the same frequency, but using a different spreading signal.<sup>5</sup>

One of the most appealing characteristics of spread spectrum communications to spectrum-hungry commercial services is that it allows multiple signals to occupy the same RF bandwidth simultaneously with minimal mutual interference. Adding spread spectrum signals to a communications band already full of conventional communications signals has the effect of raising the overall background noise level. Each additional spread spectrum signal adds to the noise floor, causing a gradual decrease in the communications efficiency for all users of the band. In comparison, adding conventional communications users in such a situation would cause severe interference to other conventional stations. Another reason that spread spectrum allows multiple stations to coexist in the same

band is that spread spectrum receivers are relatively immune to both intentional and accidental interference. All signals not spread by the expected spreading signal, whether narrowband or spread, are suppressed.

Perhaps the greatest commercial interest in the use of spread spectrum technology is in the areas of wireless computing and mobile cellular communications. Spread spectrum techniques are being used to provide wireless local area network (LAN) installations in offices and factories. In addition to minimizing interference from other electronic devices, spread spectrum provides a number of benefits over competing technologies. For example, most computer network protocols allow only one computer to transmit over the network at one time. In contrast, spread spectrum techniques can be used to support network access without the delay or collisions, even under heavily loaded conditions.

The advantages of using spread spectrum communications for cellular telephone systems include relative insensitivity to fading, freedom from interference, and spectral efficiency. In one spread spectrum cellular radio system design, the spectrum is divided into only two channels—one for the stationary base and one for mobile transmissions. Each station is distinguished from others transmitting on the same channel by a unique spreading signal. In this code division multiple access (CDMA) scheme, each signal is created with a unique spreading signal, and only receivers using the same spreading signal can clearly decode and receive the information.<sup>6</sup> Interference is minimized by assigning encoding signals that differ greatly

from one station to the next. In addition, the wideband nature of the spread spectrum signal provides some protection against fading, relative to fading that might occur if a fixed-frequency, narrowband signal were used.

## Types of spread spectrum communications

Spread spectrum communications systems vary considerably in design and function, depending on the application and operational constraints. The most common spread spectrum designs include direct sequence (DS), frequency hopping (FH), time hopping (TH), pulsed FM, and a variety of hybrid techniques. These systems are described in more detail in the paragraphs that follow.

**Direct sequence.** Figure 1 provides a simpli-

fied view of the most common spread spectrum system design, direct sequence (DS)—sometimes referred to as pseudo-noise (PN) spread spectrum. On the transmitter side, the data and spreading signal or PN code (e.g., 01000101101010111...) are mixed, and the resulting wideband signal is used to modulate an RF carrier for transmission. The modulation may be AM, FM, or more commonly, some form of phase shift keying. Although not shown here, it is also common to modulate the carrier source with the data and then modulate the resulting signal with a spreading signal before transmission. In either case, the mixing is usually performed digitally.

Since the information bandwidth is relatively narrow compared to that of the spreading signal, the bandwidth of the transmitted signal (assuming no RF filtering) is essentially determined by the bandwidth of the code function.

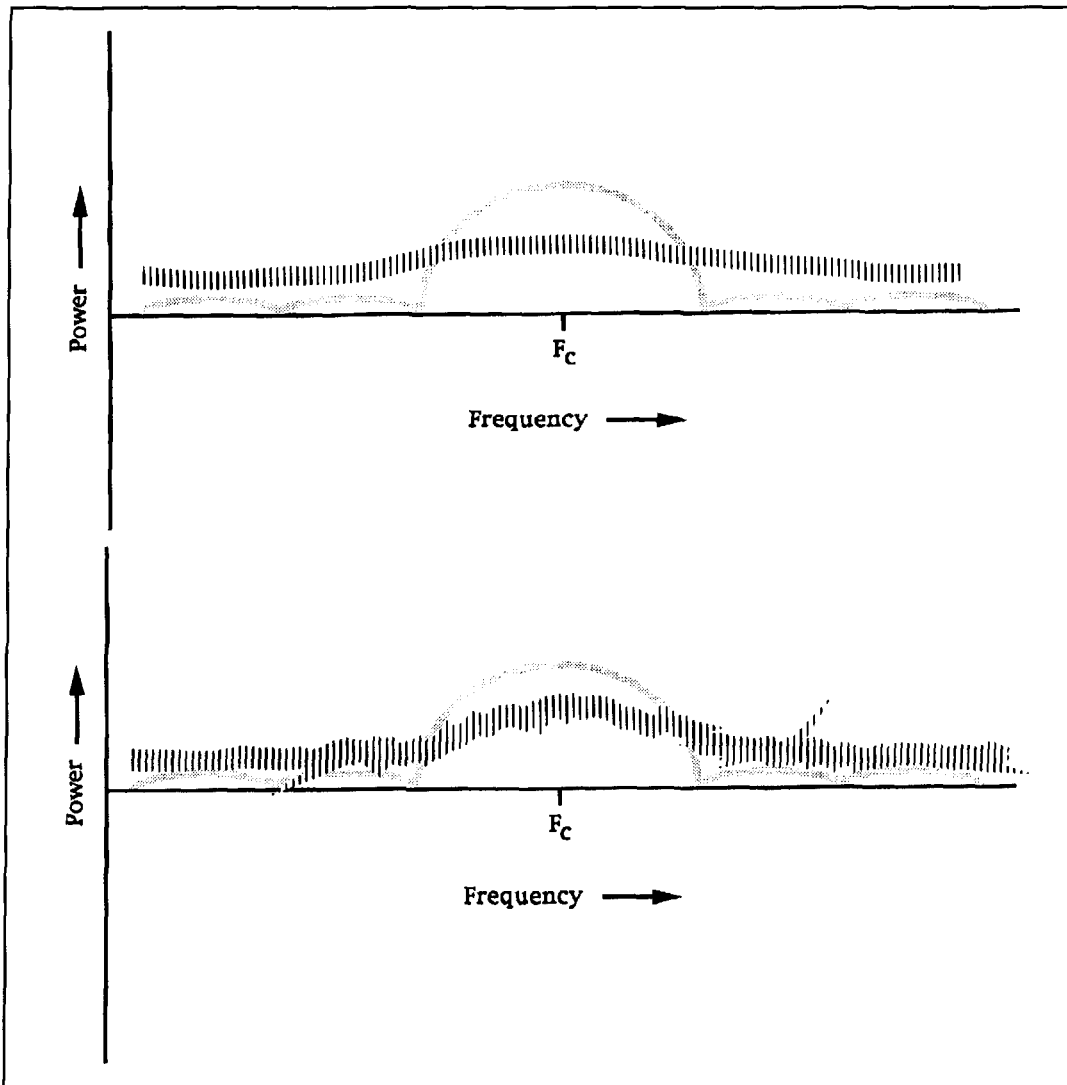


Figure 6. A typical received power spectrum for a DS signal (foreground), superimposed on the power spectrum of a DS transmitter in which the side lobes have been clipped (A). To a conventional receiver, the wideband signal appears as noise (B).  $F_c$ : center frequency.

Because of the spreading signal, the transmitted signal's frequency spectrum appears as a band of noise centered around the carrier frequency (compare **Figures 2 and 3**). Obviously, the transmitter hardware must be capable of wide-band operation.

On the receiver side, the incoming DS signal is *despread* by generating a local replica of the PN code in the receiver and then synchronizing this local PN signal to the one that is superimposed on the incoming waveform. Multiplication or remodulation of the incoming signal by the local PN signal collapses the spread signal into a data-modulated carrier by removing the effects of the spreading sequence (see **Figure 4**).

Correlation, a measure of signal similarity, is used to identify a signal that has been spread with a particular pseudo-noise sequence. Correlation is typically performed with a mixer followed by a low-pass filter. The filter output provides an average of the mixer output; a high output (1) indicates that the mixer signals differ little from each other, while a low output (0) indicates no matches. In addition to averaging correlation information, the low-pass filter also has the function of reducing noise while passing the narrowband information. After despreading, the narrowband signal can be filtered and handled by a conventional demodulator. It is the despreading process that is responsible for spread spectrum's immunity from noise and interference.

As shown in **Figure 4**, the smallest duration within the DS signal is the *chip*. Since the chip duration is much shorter than the data bit duration, the spectral components cover a much wider frequency spectrum than the data bits, so

the data is spread over a wide spectrum at a very low spectral density. Typical chip frequencies are between 1 and 10 MHz, with carriers in the low GHz region.

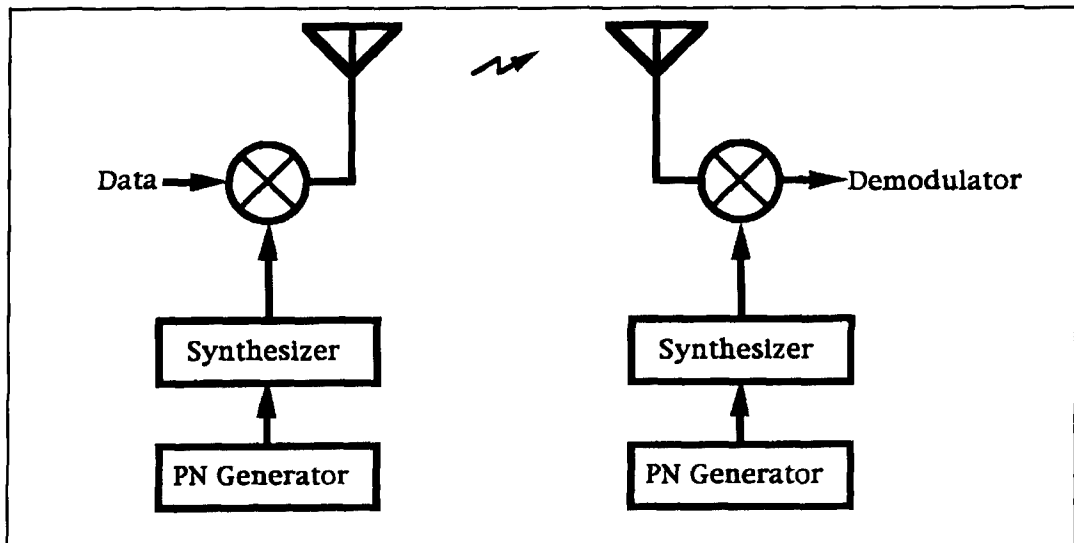
**Figure 5** shows the power spectrum for a binary phase shift keying (BPSK) DS signal in which the carrier has been suppressed. The envelope of the distribution of a DS spread spectrum signal can be approximated by:<sup>5</sup>

$$E(f) = t_c/2 \times \{\sin(\pi f t_c)/\pi f t_c\}^2$$

where  $t_c$  is the chip duration and  $f$  is frequency. The chip rate defines the spread of the signal, in addition to the size of the main lobe and the location of the nulls. The main lobe is (chip rate)/2 wide, centered around the carrier frequency, and nulls occur at multiples of 1/(chip rate).

As shown in **Figure 5**, 90 percent of the signal power lies in the main lobe. The side lobes, which contain 10 percent of the total signal power, are commonly clipped to reduce interference. Stripping the side lobes has the effect of restricting the rise and fall times of the modulating code, because the sidelobes contain much of the harmonic power of modulation. **Figure 6A** shows a typical received power spectrum for a DS signal, superimposed on the power spectrum of a DS transmitter in which the side lobes have been clipped. **Figure 6B** shows the signal as seen by a conventional receiver.

**Frequency hopping.** Frequency hopping (FH) spread spectrum is perhaps the easiest to understand conceptually (see **Figure 7**). In FH, the carrier remains at a given frequency for a duration called the dwell time and then hops to a new frequency somewhere in the spreading



**Figure 7.** Overview of a frequency hopped (FH) communications system. Transmitter (left) synthesizer frequency is controlled by a pseudo-noise signal. On the receiver end, the FH signal is captured by moving the receiver synthesizer frequency in step with a pseudo-noise signal that matches the spreading signal used at the transmitter. After despreading, the resulting narrowband signal can be processed by a conventional demodulator.

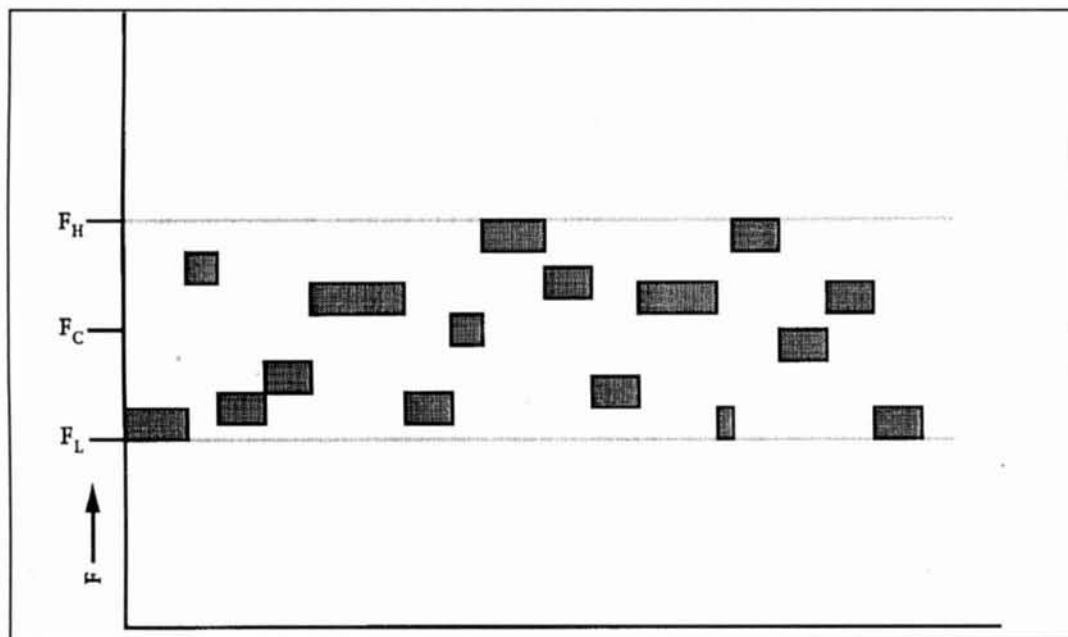


Figure 8. Spectral occupancy versus time characteristics of a frequency hopped (FH) signal. In some FH schemes, each data bit is pseudo-randomly assigned a subchannel whose instantaneous bandwidth is relatively narrow compared to the long-term, average bandwidth ( $F_H - F_L$ ) of the system. Total bandwidth is determined by the range of subchannels available and the data rate.

bandwidth (see **Figure 8**). Frequency selection is based on a pseudo-random selection from a list of available channels. In other words, the spreading sequence represents the channels to which the frequency hopper will travel. Ideally, each channel should be occupied with equal probability, and the probability of hopping from any channel to any other channel should be equal. Dwell times are generally on the order of a few milliseconds or less in order to avoid interference with conventional communications systems. In addition, the list of potential hopping channels can be modified to reflect frequencies known to be in use.

In order to receive an FH signal, the FH spread spectrum receiver must move from channel to channel according to the PN sequence used at the transmitter. In both the transmitter and receiver, the maximum rate at which frequency hopping can occur is dependent on the technology used to determine frequency. For example, one common approach is to use frequency synthesizers based on phase locked loops (PLLs). However, PLL settling time, the time required for the PLL to settle on a given frequency, limits the hopping rate. To overcome this limitation, some designs make use of two or more PLLs that are cycled into use so the PLL used has had ample time to settle while others are in use. When hopping rates greater than a few hundred hertz are required, direct digital synthesis (DDS) is a viable alternative. DDS, which makes use of values from sine wave lookup tables used to drive D-to-A

converters, is capable of providing hopping rates in excess of several kHz.

The bandwidth requirements for an FH system depend on the relative durations of the data and hopping rates. Slow FH designs use several data bits per hop, and hop rates are generally less than 100 hops per second. Fast FH, in contrast, uses several hops per data bit, and the hop rate is generally greater than 100 hops per second. The bandwidth of slow FH spread spectrum signals is essentially equal to that of the data signal, while the bandwidth of fast FH is equal to the reciprocal of the hopping duration.

The optimum hopping rate for a FH system is determined by the type of information to be sent, the information transfer rate, the amount of redundancy used, and the distance to the nearest potential source of interference.<sup>3</sup> For example, assume that 1000 frequencies are available. Unless some form of redundancy were used, an interfering signal on one of the frequencies chosen would cause one error in 1000 or an unacceptable error rate of  $1 \times 10^{-3}$ . Similarly, two interfering signals within the band of hopped frequencies would result in an error rate of  $2 \times 10^{-3}$ . The expected error rate can therefore be expressed as:

$$\text{error rate} = I/N$$

where  $I$  is the number of interferers with power sufficient to jam a channel used by the FH system and  $N$  is the total number of channels available.

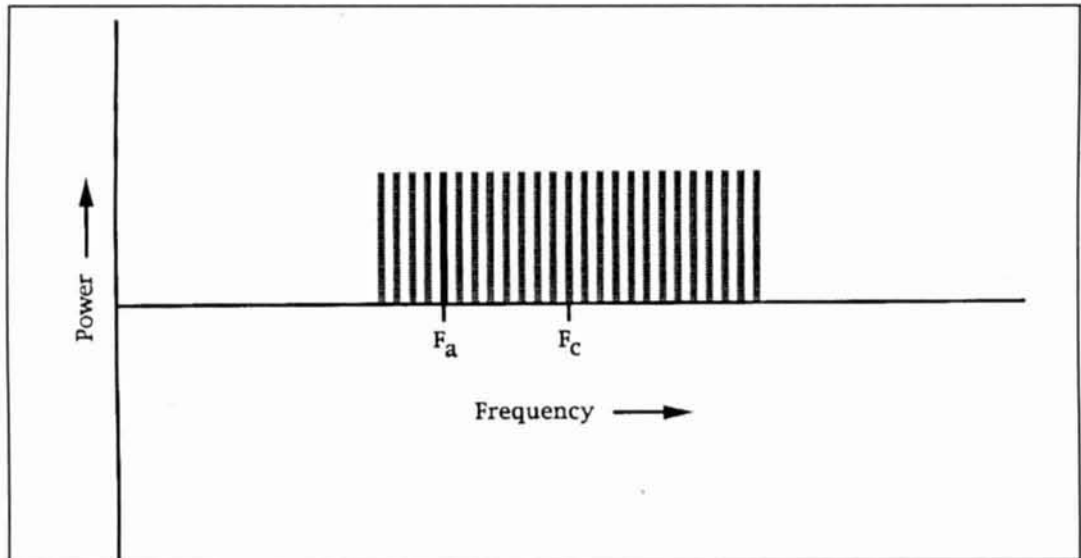


Figure 9. The long-term power spectrum for a FH spread spectrum signal. The instantaneous spectrum is much narrower, because it consists of a carrier that moves pseudo-randomly among the available channels.  $F_A$ : currently active channel;  $F_C$ : average signal center frequency.

Error rates can be reduced by using some form of redundant transmission. For example, three or more frequencies can be used for each bit of information; i.e., multiple chips per data bit. If the receiver's bit decision is made on the basis of two out of three chips being correct (e.g., at three frequencies per bit, and a received pattern of "101," the bit would be termed a "1"), a single channel interferer would cause no more than  $p^2(3 + p)$  errors,<sup>3</sup> where  $p$  = the error probability for a single trial. For a 1000 channel FH system ( $p = 1/1000$ ), the expected error rate would be:

$$\begin{aligned} \text{error rate} &= p^2(3 + p) \\ &= (1/1000)^2 \times (3 + 1/1000) \\ &= 3 \times 10^{-6} \end{aligned}$$

The error rate has been reduced by a factor of 3000 by increasing the hop rate by a factor of three. The limitations of lowering the bit error rate by sending more chips per bit lie in the capabilities of the frequency synthesizer and the RF bandwidth available. Because the bandwidth required increases in direct proportion to the hopping rate, a tradeoff must generally be made between sending more chips per bit and reducing the number of frequencies available. Alternatively, the bandwidth requirement may be reduced by adjusting the transmit frequency spacing so significant overlap occurs between channels. The overlap reduces the bandwidth requirement for the transmitted spread spectrum signal.

The average or long-term power spectrum for

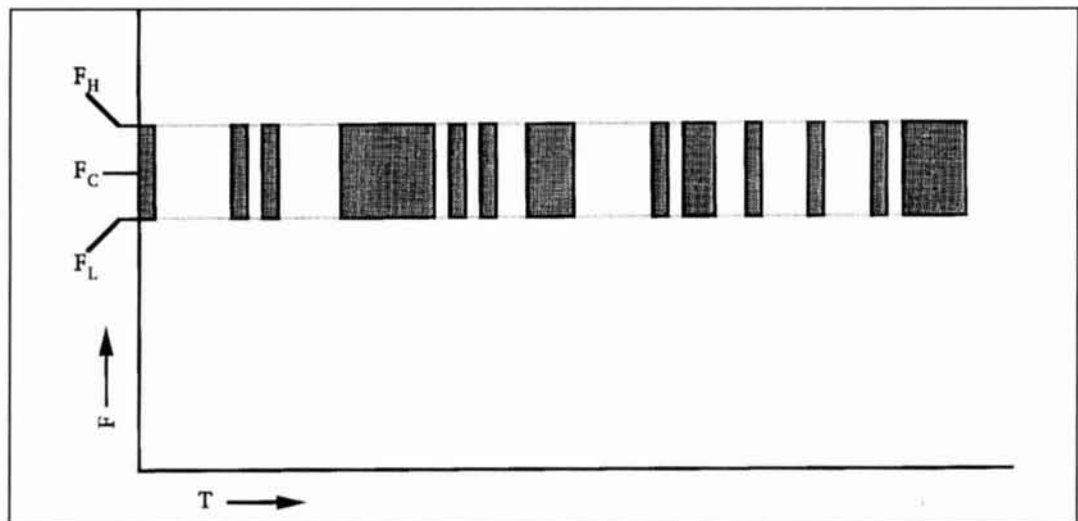


Figure 10. Spectral occupancy versus time characteristics of a time-hopped (TH) signal. Bandwidth ( $F_H - F_L$ ) is primarily a function of the pulse repetition rate, which is varied in a pseudo-random manner.



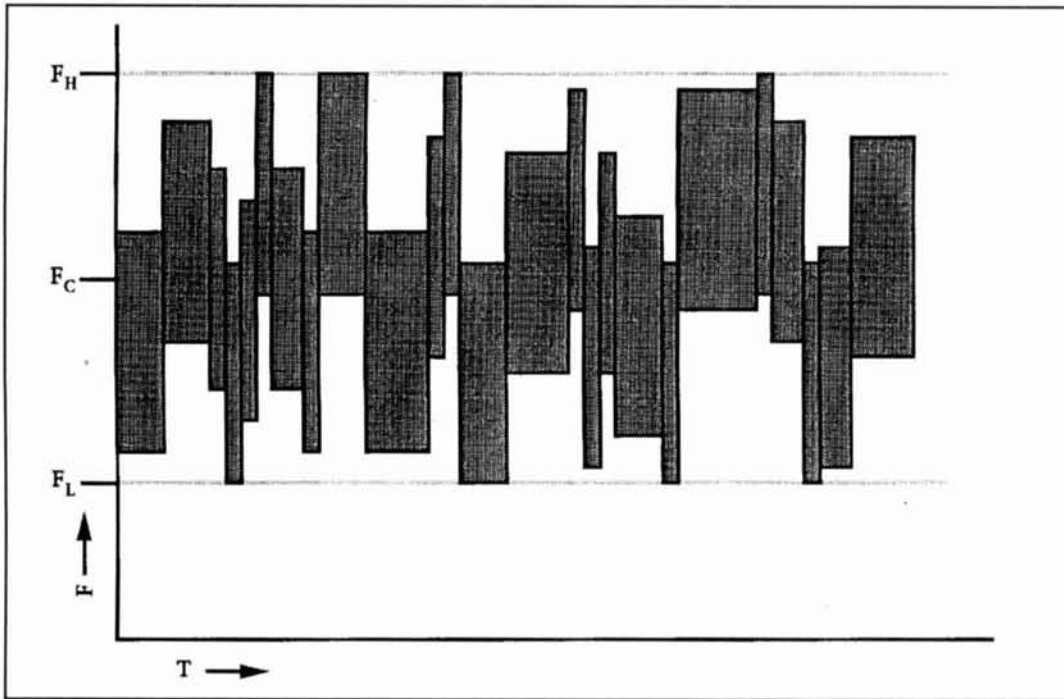


Figure 11. Spectral occupancy versus time characteristics of a Frequency Hopped/Direct Sequence (FH/DS) signal. This hybrid spread spectrum technique is basically a DS system in which the center frequency is periodically hopped. Note that although in this example the DS channels overlap somewhat, in many cases there is no overlap. Compare with Figures 8 and 10.

an FH spread spectrum signal is shown in **Figure 9**. The instantaneous spectrum is much narrower because it consists of a carrier that moves pseudo-randomly among the available channels. One of the features of FH is that channels within the spread spectrum bandwidth known to be allocated for other purposes can be avoided by removing the channel from the

lookup table of possible FH channels. The input and output frequencies of conventional repeaters can be easily avoided in this manner, reducing false repeater keying and potential repeater interference.

**Time hopping.** In time hopping (TH) spread spectrum systems, the carrier is keyed on and off by a PN generator (see **Figure 10**). As in

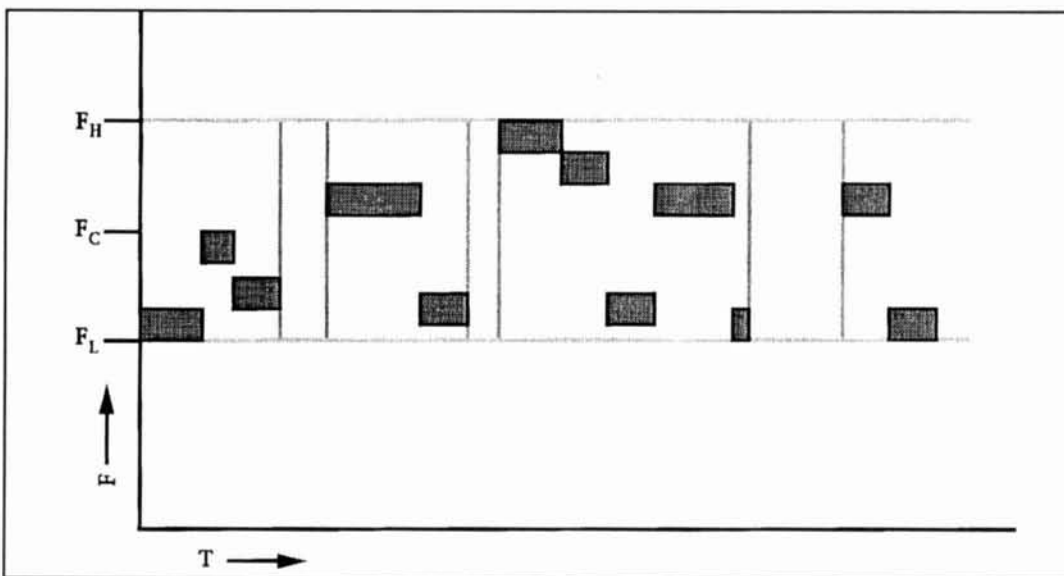


Figure 12. Spectral occupancy versus time characteristics of a Frequency Hopped/Time Hopped (FH/TH) signal. As in simple FH, each data bit is pseudo-randomly assigned a subchannel. However, as in TH, there are randomly assigned periods when no signal is transmitted. The long-term, average bandwidth ( $F_H - F_L$ ) of the signal is a function of the range of subchannels available, the repetition rate, and the data rate. Compare with Figures 8 and 10.

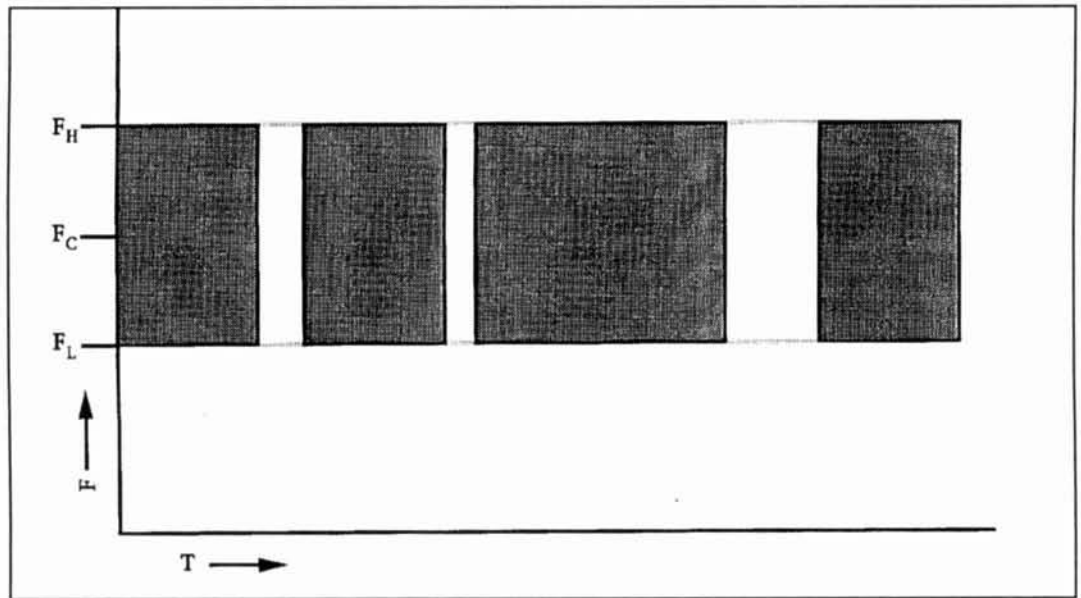


Figure 13. Spectral occupancy versus time characteristics of a Time Hopped/Direct Sequence (TH/DS) signal. This hybrid form of spread spectrum is basically a DS system that is switched on and off at pseudo-random intervals.

conventional CW transmissions, the bandwidth of a TH signal is a function of the keying rate. TH systems can vary considerably in duty cycle. For example, one extreme might be a burst transmission lasting a fraction of a second that is transmitted only once a day. Such a signal would be very difficult to detect, intercept, and exploit.

**Hybrid techniques.** Hybrid spread spectrum techniques, created by combining two or more forms of spread spectrum into a single system, can often outperform systems based on single techniques or provide the same capabilities with less expense or complexity.

**Frequency Hopping/Direct Sequence.** FH/DS spread spectrum is in essence a DS system whose center frequency hops periodically (see Figure 11). This hybrid technique is used for multiplexing, for multiple and discrete address, and to extend spectrum spreading. FH/DS is often the most attractive means of achieving extended spectrum spreading, because PN code generator limitations place practical limits on DS spreading signal bandwidths and FH frequency hopping rates. For example, to achieve a 1 GHz wide spectrum, a DS system would require a 500 Mbps PN generator, and an FH system would require a 1 GHz bandwidth frequency synthesizer. A cheaper and more realistic alternative to realizing the same 1 GHz bandwidth would be to use a 100 Mbps PN generator and a frequency synthesizer capable of generating 20 frequencies spaced 50 MHz apart.

**Time Hopping/Frequency Hopping.** FH characteristics are often added to TH systems in antijamming situations (see Figure 12). While TH spread spectrum signals can be blocked

easily by a carrier at a single frequency, a combined TH/FH system is impervious to blocking transmissions from a single frequency source.

**Time Hopping/Direct Sequence.** TH/DS spread spectrum is basically a DS system that is switched on and off (see Figure 13). Switching occurs at pseudo-random intervals, as determined by the PN generator used for spreading or despreading. The time division afforded by the time hopping permits more channel users, and therefore more access.

## Spread spectrum technology

As described above, spread spectrum communications systems can vary greatly in design and function. Even so, there are several key concepts relevant to all spread spectrum communications systems, including the role of the pseudo-noise code sequence, synchronization, despreading, and the concept of process gain.

## The pseudo-noise (PN) code sequence

Key to the operation of all spread spectrum systems is the generation of a wideband pseudo-noise (PN) code sequence for spreading and despreading. The pseudo-random "noise" sequence is designed to appear to be random (noise), but in practice only approaches a random distribution.<sup>7,8,9,10</sup>

The code sequence is referred to as pseudo-random because it is deterministic; given any part of the sequence, it is possible to generate

the remaining sequence. With a truly random generator, it would be impossible to determine future sequences. For example, consider the repeating sequence {...7 2 4 1 5 7 2 4 1 5 7 2 4 1 5...}. With only five elements in the sequence, the deterministic nature of the sequence is obvious; i.e., given the value 4, we know that 1, 5, 7, and 2 follow before the sequence repeats. A practical PN code generator may create sequences with hundreds or thousands of elements, thereby obscuring the deterministic nature of the sequence. For example, the civilian spreading code used in the GPS system, which allows position to be determined to within about 100 meters, is 1023 bits in length, repeated every 1 ms. By comparison, the secure military code, which supports an accuracy of 1 meter, is a 267 day long code sequence running at 10.23 Mbps.<sup>5</sup>

PN sequences are often generated with linear-feedback shift registers, each of which consist of several one-bit memory registers that shift their contents with each clock pulse. The output of each shift register is deterministic; initializing the shift register with a given sequence will start the entire cycle from that point. The maximum length of the repeating sequence is a function of the number of shift register stages (N), as defined by  $2^N - 1$ .<sup>11</sup> For example, a shift register with 12 stages can provide a sequence with up to  $2^{12} - 1$  or 4095 elements before the sequence repeats.

The type of spread spectrum communications used determines the nature and use of the PN generator. For example, the PN sequence used in DS systems is most commonly a binary pseudo-random sequence (e.g., 01001101010111...) with a variable repetition rate that is usually some submultiple of the clock frequency. In other types of spread spectrum systems, the sequence may take the form of a numerical sequence such as {...2, 45, 1, 255, 87, 109, 14...}, where each value is keyed to a particular channel, time delay, or other signal parameter.

## Despreading

In the initial work on spread spectrum communications, the purpose of the PN modulation was to create a signal that, to an “unfriendly” receiver, appeared to have great uncertainty and complexity, while simultaneously appearing perfectly deterministic and predictable to a “friendly” receiver. The goal was to dilute the signal energy to the point where it fell below the noise floor of a conventional receiver, and was undetectable by this type of receiver. As illustrated in **Figure 4**, this duality is possible because of the PN modulation; only a receiver with a copy of the spreading PN sequence can *despread* a spread signal

into a data-modulated carrier.

Despreading is the process of correlating the received spread spectrum signal with a local reference PN signal. When the received signal and local PN generator are properly synchronized (a nontrivial task), the information signal collapses to its original bandwidth before spreading. The resulting narrowband information signal can be handled by conventional filtering and demodulation.

To receivers without access to the original spreading code, whether conventional or spread spectrum, spread spectrum signals appear as noise. Received signals that bear little resemblance to the receiver PN code sequence are spread by the PN spreading signal to noise. Similarly, the greater the degree of similarity between the spreading PN code and the sequence used to spread a potentially interfering signal, the less able the spread spectrum receiver is to reject the interference. Ideally, the PN spreading sequence should have a low correlation with shifted versions of itself to avoid interference. Reception of shifted sequences can be a problem in cases of multipath reception or when multiple spread spectrum units operate with the same PN sequence, but with different starting points within the sequence. There is a considerable amount of literature on optimum PN sequences for spread spectrum.<sup>11</sup>

## Synchronization

A prerequisite to despreading, and the major operational and design problem in spread spectrum work, is the synchronization of the received signal and the local pseudo-noise sequences. The simplest approach to synchronization is to examine all possible phase relationships between the received PN sequence and the locally generated PN sequence until a match is found. This can be achieved by allowing the two signals to slip past each other by offsetting their clock frequencies, while despreading the two signals. Once a correlation is found, a tracking loop can be used to maintain synchronization.

Synchronization, which usually occurs after the despreading operation,<sup>12</sup> is often viewed as a two-part process, rough and fine synchronization.<sup>13</sup> Rough synchronization or *PN acquisition*, the process of bringing the received and local PN signals into rough alignment, is generally based on a method that provides some measure of correlation—for example, the received and local PN signals are multiplied. Fine synchronization or *PN tracking* attempts to maintain bit or channel dwell timing through some type of tracking loop.

Two methods used to achieve rough synchronization are *epoch* and *phase* synchronization.

In the process of epoch synchronization, the transmitter periodically sends a special synchronization sequence that can be easily detected by the receiver—typically by a digital matched filter. This filter consists of a shift register that clocks the received signal one bit at a time and compares it against the unique sequence. Phase synchronization attempts to identify which of the  $2^N-1$  (the maximum length of the repeating sequence) possible phases the PN sequence could be in. At least  $N$  bits must be received without error for phase to be determined;<sup>13</sup> more than  $N$  bits must be received with noisy signals or in conditions with a poor signal-to-noise ratio. Obviously, in cases of both large  $N$  and poor signal conditions, PN acquisition time can be significant. Synchronization acquisition time can often be reduced by using timing references available at both the transmitter and receiver site, such as local AM broadcast carriers, to start the PN generators.<sup>2,14</sup>

Fine synchronization, which focuses on the bit or channel level, generally uses a feedback loop to minimize the timing error between the transmitter and receiver PN sequence by adjusting the receiver spreading code clock. Dithering is one commonly used fine synchronization technique. With this technique, the spreading code clock is continually rocked back and forth, and the point at which best synchronization is achieved is used to synchronize the receiver clock. A feature of dithering is that it can track a spreading code whose timing is changing due to changes in propagation or relative motion between the transmitter and receiver.

Depending on the nature of the application, it may be possible to minimize the problem of synchronization by transmitting a preamble before a spread spectrum signal. A preamble signal, which can alert spread spectrum receivers to initialize their synchronization procedure, can take the form of a tone appearing on a prearranged frequency or frequencies. For example, the falling edge of a tone can signal the beginning of PN code generation. Alternatively, a digital sequence preamble can be used to identify the transmitter carrier center frequency, synchronize the receiver local clock, and enable rough synchronization. For instance, a sequence of all zeros can be used to identify the carrier center frequency, and a series of alternating ones and zeros can be used to establish timing information. Similarly, the spreading code can be identified through the use of a synch trigger within the preamble (that is, epoch synchronization).

## Process gain

An important concept in spread spectrum communications is that of process gain. Process

gain is the difference between output and input signal-to-noise ratios; the greater the process gain, the better. For example, a system with an input S/N ratio of 20 dB and an output S/N ratio of 26 dB has a process gain of 6 dB. The process gain afforded by FH and DS spread spectrum systems operating in conditions of negligible interchannel interference can be approximated by the following relationship:<sup>3</sup>

$$\text{process gain} = \text{BW/DR}$$

where BW is the RF bandwidth of the transmitted spread spectrum signal and DR is the data rate in the information channel. The bandwidth (BW) value for a DS system is approximately equal to twice the system code clock rate. The bandwidth (BW) value for an FH system is approximately equal to the number of frequency channels available multiplied by the channel bandwidth. The process gain for an FH system can be computed from the equation above or by simply summing the number of frequency choices; e.g., the process gain associated with an FH system with 1000 available channels would be 1000 or 30 dB ( $10 \log 1000/1 = 30$  dB). Although the above relationship holds for FH and DS spread spectrum, the process gain provided by TH spread spectrum is the reciprocal of the transmit duty cycle.<sup>3</sup>

Although the optimum bandwidth for spread spectrum depends on the desired system capabilities, maximum process gain and signal hiding calls for wide bandwidths. In addition, from the equation above, we can see that, for a given process gain, the RF bandwidth requirements increase with increasing data rates. However, it is often impractical to increase the RF bandwidth to support higher data rates. For example, a system with a 10 Mbps code clock rate and 1 Kbps information rate would provide a process gain of:

$$\begin{aligned} \text{process gain} &= \text{BW/DR} \\ &= (2 \times \text{system code clock} \\ &\quad \text{rate})/(\text{data rate}) \\ &= (2 \times 10^7)/(1 \times 10^3) \\ &= 2 \times 10^4 \\ &= 43 \text{ dB} \end{aligned}$$

One way to increase the process gain of this system would be to increase the system code clock rate. For example, doubling the clock rate to 20 Mbps provides an increase in process gain of only 3 dB. This strategy is expensive and fraught with problems, given the cost and bandwidth limitations of ICs operating above a few hundred MHz. Also, because the PN code generators must operate without error for hours or days, as code rates increase, operating errors must decrease in inverse proportion. This means that a 100-MHz clock must operate error

free for  $3.6 \times 10^{11}$  bits to allow a system to operate for one hour. However, high speed digital circuitry is more sensitive to noise, consumes more power, and is more susceptible to error than low speed circuitry. A 100 Mbps code generator with an error rate of only  $1 \times 10^{-8}$  would, on average, make one mistake every second—an unacceptable condition for spread spectrum communications.

Given the practical limitations of increasing process gain by increasing RF signal bandwidth, the next best way to increase process gain is to decrease the information bandwidth. Process gain varies inversely with the data rate;<sup>2</sup> reducing the data rate by a factor of two provides an additional process gain of only 3 dB. Despite this modest gain, there are no real technical limitations as to how much the data rate can be decreased. For instance, following the previous example with a clock rate of 10 Mbps, but with a data rate of 10 bps instead of 1 Kbps, the process gain is:

$$\begin{aligned} \text{process gain} &= \text{BW/DR} \\ &= (2 \times 10^7)/(10) \\ &= 2 \times 10^6 \\ &= 63 \text{ dB} \end{aligned}$$

which represents an additional 20 dB of process gain. A data rate of only 10 bps might seem agonizingly slow; however, it may be the only possible solution for some harsh communications environments.

## Going further

Ready for more? Construction enthusiasts should see the articles by Kesteloot<sup>15,16</sup> for a good introduction to practical spread spectrum hardware projects. Dixon<sup>3</sup> provides an excellent and very readable introduction to spread spectrum techniques. For a good overview of the techniques used in synchronization, see the text by Simon.<sup>12</sup> Dillard's text<sup>17</sup> offers insight

into how spread spectrum is used in secure communications.

As was mentioned earlier, radio amateurs can experiment with DS and FH spread spectrum communications above 420 MHz. However, since the FCC rules regarding spread spectrum communications logging, station identification, and other operating procedures are very different from those applied to standard communications, it would be advisable to consult part 97 of the FCC rules and regulations before you set out to explore this technology. ■

## REFERENCES

1. L. Wirbel, "Spread the Spectrum," *Electronics Engineering Times*, March 8(736): 56-8, 1993.
2. M. Egtvedt, "Spread-Spectrum Modulation," *Electronic Engineering Handbook*, Fink and Christiansen, editors, 1989. McGraw-Hill Book Company, New York.
3. R. Dixon, *Spread Spectrum Systems*, 1984, John Wiley & Sons, Inc., New York.
4. L. deRosa and M. Rogoff, *Application of Statistical Methods to Secrecy Communication Systems*, 1950.
5. R. Lucas, "Spread-Spectrum for Multiple Access," *Electronics and Wireless World*, 97(1659): 23-5, 1991.
6. P. Wayner, "Stretching the Ether," *Byte*, 18(2): 159-65, 1993.
7. C. Regena, "The Random Function," *Compute!*, 9(Nov): 85, 1987.
8. C.A. Whitney, "Generating and Testing Pseudo-Random Numbers," *Byte*, (Oct): 128-464, 1984.
9. S.K. Park and K.W. Miller, "Random Number Generators: Good Ones Are Hard to Find," *Commun ACM*, 31(10): 1192-201, 1988.
10. B. Wichman and D. Hill, "Building a Random-Number Generator: A PASCAL Routine for Very-Long-Cycle Random-Number Sequences," *Byte*, 12(Mar): 127-9, 1987.
11. R. Dixon, *Spread Spectrum Systems*, 1976, John Wiley & Sons, Inc., New York.
12. M. Simon, J. Omura, R. Scholtz, and B. Levitt, "Pseudonoise Code Acquisition in Direct-Sequence Receivers," *Spread Spectrum Communications*, Pickholtz, editor, 1985, Computer Science Press, Rockville, Maryland.
13. R. Schetgen, "Special Modulation Techniques," *The ARRL Handbook for Radio Amateurs*, Schetgen, editor, 1993, The Amateur Radio Relay League, Newington, Connecticut.
14. A. Kesteloot, "Extracting Stable Clock Signals From AM Broadcast Carriers for Amateur Spread-Spectrum Applications," *QEX*, (Oct): 5-9, 1987.
15. A. Kesteloot, "Spread-Spectrum: A Fascinating Mode—and Legal for Hams to Use!," *73 Amateur Radio Today*, (June): 12-13, 1989.
16. A. Kesteloot, "A Practical Direct-Sequence Spread-Spectrum UHF Link," *QST*, 73(5): 14-21, 1989.
17. R. Dillard, *Detectability of Spread Spectrum Signals*, 1989, Artech House, Inc., Norwood, Massachusetts.

## PRODUCT INFORMATION

### New Switchcraft audio and video components catalog

Switchcraft has released a new, 16-page, 4-color, AVP-3 audio and video components catalog featuring the company's audio/video products including jacks, plugs, connectors, patch panels and cords. Product photographs, dimensions, specifications, ordering and mating information is included.

To request the AVP-3 catalog, contact Switchcraft's Marketing Communications Department at 312-792-2700 extension 243.



**Peter O. Taylor**  
AAVSO Solar Division  
4523 Thurston Lane, #5  
Madison, Wisconsin 53711-4738

# THE SOLAR SPECTRUM

*The realm of the Sun*

**M**any of us are familiar with the magnetopause, which marks the boundary between the continuous flow of particles away from the Sun—the solar wind—and the Earth's magnetic field. However there is a less well known, but somewhat analogous, border that encompasses the realm of the Sun: it is known as the *heliopause*.

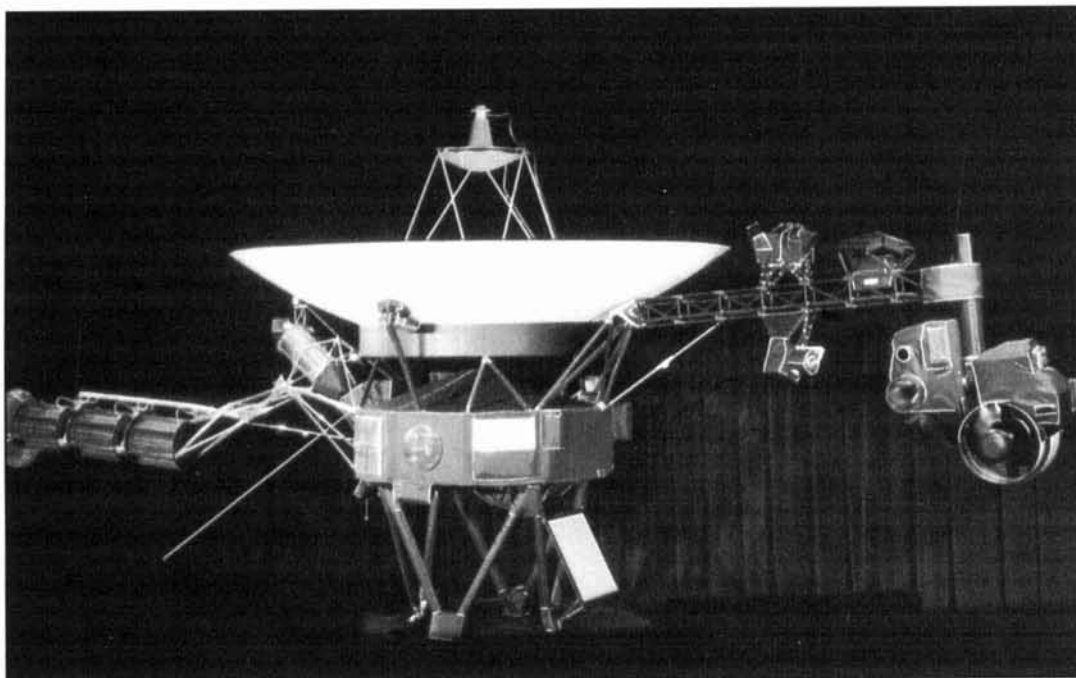
The heliopause represents the frontier between the farthest reaches of the solar wind and the surrounding interstellar wind. It encloses the *heliosphere*, where, within a huge bubble of magnetized gas, the Sun dominates all that comes in contact with its electromagnetic radiation and gravitational force. Until recently, little was known regarding this isolated region of space where the Sun's influence merges with that of the stars.

Now, new information about the heliopause has been obtained by sensitive instruments aboard the extraordinarily productive Voyager spacecraft (**Photo A**). The latest data from the

Voyagers suggests that the heliopause, once thought to be much closer to the center of the solar system, actually lies some 100 astronomical units (AU) distant from the Sun . . . about 15,000 million kilometers away. (One AU represents the average Earth-Sun distance, or about 149.6 million kilometers.) Moreover, these new observations indicate that the conflict between solar and stellar winds generates what may be the most powerful radio emission in the entire solar system; a burst the Voyagers found to exceed 10 trillion watts of energy!

Both Voyager spacecraft, now located far beyond the orbits of Neptune and Pluto, have contributed to the new findings. Voyager 1 completed successful flybys of both Jupiter and Saturn and is now rising above the ecliptic plane—the great circle marked by the intersection of Earth's orbit with the celestial sphere—floating toward the stars at a distance of some 52 astronomical units.

Voyager 2 also visited Jupiter and Saturn



**Photo A.** The Voyager 1 spacecraft. Photograph courtesy of NASA.

before completing the survey of the giant outer planets through spectacular encounters with Uranus and Neptune. The craft is currently at a distance from the Sun of over 40 AU. Both vehicles were launched in 1977, and are managed by the Jet Propulsion Laboratory in Pasadena, California.

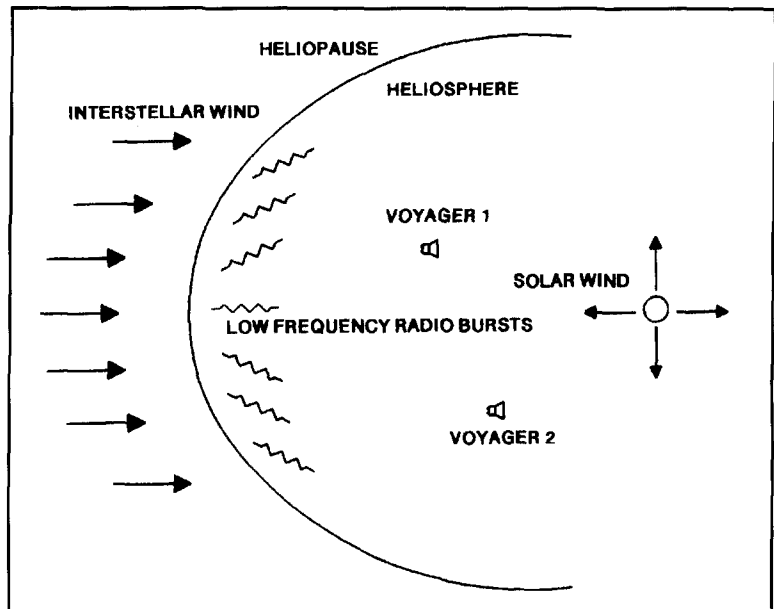
Beginning July 1992, the radio antennas of both spacecraft—instrumentation referred to as the Plasma Wave Sub-System—recorded mysterious low-frequency radio bursts that are believed to emanate from the extreme edge of the solar system; the heliopause. The signals, which reached a peak last December, are much stronger than a similar set of bursts recorded by the Voyagers in the early to mid-1980s. They were not detected at the Earth because the emission occurred at a very low frequency (2 to 3 kHz), which could not penetrate the Earth's magnetic field. In fact, we would not even know the bursts had taken place if it were not for the Voyagers and their unique locations in space.

According to University of Iowa professor Dr. Donald Gurnett, the principal investigator of the Plasma Wave Sub-system,<sup>1</sup> a preliminary interpretation of these events indicates that the radio signals are created as a cloud of electrically charged gas, called "plasma," which expands from the Sun and interacts with the cold interstellar gas beyond the heliopause (see **Figure 1**).

Apparently, the intense solar flare activity that occurred in May and June 1991, spawned an enormous cluster of energetic particles which, carried by the solar wind, moved rapidly outward over the solar system. We described this remarkably powerful series of eruptions in our special report to *Communications Quarterly*, which appeared in the Summer 1991 issue. Dr. Gurnett and his colleagues believe that when the cloud of particles reached the farthest reaches of the solar system and slammed into the interstellar wind, a violent reaction occurred which produced the immense burst of energy recorded by the Voyagers.

A hint at the coming interaction between solar and stellar winds actually came only a few months after the solar eruptions ended. Once again the phenomenon was recorded by the ever-vigilant Voyagers. Three months after the flares occurred, both crafts recorded a sudden decline in galactic cosmic rays—the nuclei of heavy atoms born in cataclysmic events like exploding stars. Astronomers believe that these ultra-high energy particles were deflected away by a blast of solar wind hurtling through interplanetary space at a speed between 600 and 800 kilometers a second—nearly double the average wind speed.<sup>2</sup>

When the Voyager measurements are compared with the solar eruptions that began this series of events, it is possible to draw some conclusions concerning the nature and location of the heliopause. Accordingly, that boundary is now thought to lie somewhere between 90



**Figure 1.** According to an interpretation of a recent discovery by NASA's Voyager spacecraft, intense radio bursts occur at the heliopause as a result of the clash between clouds of energetic particles carried by the solar wind, and the interstellar wind. (Prepared from a similar NASA representation of these events.)

and 150 or more astronomical units from the Sun, much farther than anticipated. Furthermore, the emissions have slowly changed to a higher frequency, suggesting a substantial heliopause thickness. (Higher frequency radio emission is believed to occur as the burst of solar wind penetrates deeper into the denser regions of the heliopause.)

Further information about the heliosphere has been unearthed through a separate study, conducted by a joint U.S.-French team analyzing information obtained with the Hubble Space Telescope's High Resolution Spectrograph.<sup>3</sup> The data gathered by HST provides additional evidence that gases in the vicinity of the solar system congregate in clouds emersed in a generally gas-free portion of the solar neighborhood, and sheds new light on the beginnings and consequent development of this region.

The group of scientists examined reflected sunlight near Earth's sister planet, Mars. Among other findings, they learned it is likely that incoming atoms of interstellar hydrogen are stripped of their electrons near the heliopause by protons from the Sun, a process that slows the speeding hydrogen particles by some 20 percent.

Apparently helium atoms are not similarly affected, possibly because the solar wind protons can't steal their electrons as easily as from the simpler hydrogen atoms. This study appears to confirm portions of the Voyager information, since both research efforts indicate a distance of about 100 AU for the heliopause. If these conclusions prove to be correct, the Voyager

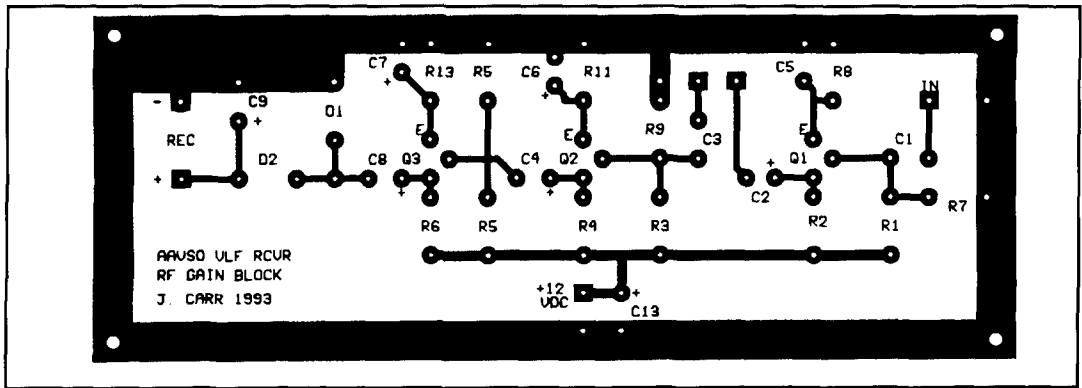


Figure 2. Joseph J. Carr devised this pboard pattern with *EASY-PC* software. The board is designed to be used with a version of Arthur Stokes' popular VLF receiver.

spacecraft should pass beyond the Sun's energy influence around the year 2010.

These discoveries are important events for astronomers and other scientists who have long theorized about the structure of the heliosphere and the exact whereabouts of the heliopause. Equally exciting, the new findings hint at future revelations as the Voyagers continue their long journey to the outer regions of the solar system, and beyond.

### More from Joseph J. Carr

As we reported in our last column, Joseph J. Carr, K4IPV, continues to develop the VLF receivers designed by Arthur Stokes, N8BN. With the aid of a small computer and *EASY-PC* software,<sup>4</sup> Mr. Carr has devised patterns for several printed circuit boards, which he intends to place in the public domain.

The pc board pattern shown in **Figure 2** is intended for use with a version of the Stokes receiver that uses one, rather than the normal two, tuning inductors. (A schematic and parts list for this version is available from the author.) The tuning circuit for the design is meant for off-board mounting, so it is not included on the pattern. For those who do not make pc boards themselves, Mr. Carr suggests contacting Fred Reimers, N9ATW (18N640 Field Court, Dundee, Illinois 60118). If demand warrants, Mr. Reimers may add one or more of these to his list of commercially available boards.

### Recent activity and short-term outlook<sup>5</sup>

Solar activity between the first of June and the end of August was generally in the low range. The most active period occurred during the final week of June and the first few days of July, when a large, complex sunspot group in the Sun's Southern Hemisphere (NOAA/USAF Region 7530) dominated the solar disk.

The group represented the second appear-

ance of a long-lived spot complex that was also an active flare producer during its previous transit of the visible hemisphere (Region 7518). All told, this group is credited with approximately 150 separate flare eruptions, including at least 13 class M flares—several designated as major events. However, due to the extreme easterly location of this group when it was most active, the terrestrial impact from these flares was minimal.

In spite of the occurrences in Region 7530, the geomagnetic field was most active during several days in mid-August. Reports of aurorae from the northern United States and Canada also increased markedly around this time. It is interesting to note that this period of minor to severe disturbances (depending on latitude) was not related to flare eruptions, but rather to "gusts" in the solar wind associated with a strategically located coronal hole. Such activity is to be expected during the remainder of 1993 and throughout 1994, since it is particularly likely to occur during the declining portion of the solar cycle.

The outlook for the next several months calls for more of the same, although geomagnetic activity may well increase around the Fall equinox. The rate of solar activity appears to have moderated recently, slowing the rapid decline that has been the rule for much of the post-maximum phase of this cycle. If this trend continues during the immediate future, the end of solar cycle 22 and the onset of cycle 23—now forecast to occur sometime during 1996—may be delayed. However, it is still very early, and any such predictions necessarily carry a low statistical weight. ■

#### REFERENCES

1. Public Information Office, Jet Propulsion Laboratory, Pasadena, California 91109, May 26, 1993.
2. J.A. Lockwood and W.R. Webber, *Journal of Geophysical Research, Space*—May 1, 1993.
3. Clarke, et al., *Science*, May 21, 1993.
4. Mr. Carr reviews this software in an upcoming issue of *73 Amateur Radio Today*.
5. A portion of this information was obtained from the SELDADS data base.



# TECH NOTES

## Simple projects

*Simple construction projects are always popular, and this month's "Tech Note" lead article, "A Simple APT Weather Satellite Interface," garnered from the pages of Break-In magazine, is one of those pieces we have carefully picked to satisfy the demand for short, useful, and unique projects.*

*Peter Bertini, K1ZJH  
Senior Technical Editor*

## Simple APT Weather Satellites Interface

*Robin Ramsey, ZL3TCM  
Reprinted from Break-In, March 1992*

**I have always** had a fascination with satellites, not just amateur but commercial and military types as well. In the past, I've seen adverts in magazines for hardware and software to decode APT (automatic picture transmission) satellite signals. The main hurdle here is the high cost of the dedicated equipment required to get even mediocre pictures. The interface described here was developed by Graham ZL3OY, David, ZL3TPD, and me as a cheap conversion interface so we could use existing fax receive setups.

### Satellites

The two main types of APT satellite are the Russian Meteor series and the USA's NOAA series—orbiting about 1100 km and 850 km, respectively. At the end of 1991, the main frequencies used were:

137.5 and 137.62 MHz for NOAA  
137.3 and 137.85 MHz for Meteor

The American satellites sound like a ticking clock, while the Russian ones sound like a honking goose—quite unmistakable once heard.

The picture is sent as a 50 kHz deviation FM transmission of a 2400 Hz tone. The picture is produced by a spinning mirror focusing the image from below line-by-line onto either a visible light or infrared detector. (or both in the case of NOAA). This video signal amplitude modulates a 2400 Hz signal, giving us a video output when synch pulses are added. As the satellite moves forward, a new area of ground underneath it is covered and transmitted.

The APT information is actually made up on the NOAAs from high resolution pictures captured and transmitted on 1600 MHz. This high resolution format is digital and not recoverable using standard format APT gear. The satellites

transmit continuously; i.e., there are no start or stop tones sent. If it were possible to continuously receive a suitable signal, you would get a very long strip of the earth approximately 2400 km wide. From Christchurch it is possible to see from Antarctica right up to the islands north of New Zealand on most overhead passes. If a beam is used, the center of Australia can be seen with none of New Zealand in the picture. An improvement in the picture quality (noise free) is also noticed on horizon pictures to the east and west when a beam is used.

Initially, I talked both to local amateurs already receiving these pictures and many like myself looking for cheaper alternatives. We concluded that there must be a way around the cost barrier. Especially since most of these amateurs already had equipment they were keen to try out without the need to use dedicated equipment. All that was needed was a new approach to the reception technique.

Further contact with local amateurs already receiving APT showed me how easy it was to get carried away on upgrades to equipment in the quest for better pictures. The occasional look would not justify the purchase of the commercial or kitset types of systems.

My first step was to have a listen to these birds go over and determine what was required for reception. A couple of very good programs are around for predicting when the birds are visible: Quiktrak 4 and Instant Track 1, both available from AMSAT at reasonable cost. Don't forget that money sent to AMSAT all goes towards amateur satellites of the future. Tuning into one of the satellites gave quite strong signals even on a discone, which was encouraging and meant that nothing too fancy would be required.

I also heard other types of APT from satellites such as Feng-Yung/Chinese, Okean/Russian, and Cosmos/Russian military. I have not attempted to receive these yet and there is little information around on these satellites.

### Aerials

The most popular types for this mode are the crossed dipole and J pole antennas. The odd keen local has a beam for this frequency and at least one of the Christchurch weather watchers often gets pictures from the middle of Australia, where I can only see the East Coast. I wanted to keep things simple and cheap, so I went for a J pole. It worked fine except the pre-amp kept blowing up after rain. I also tried a crossed dipole, which worked well from around

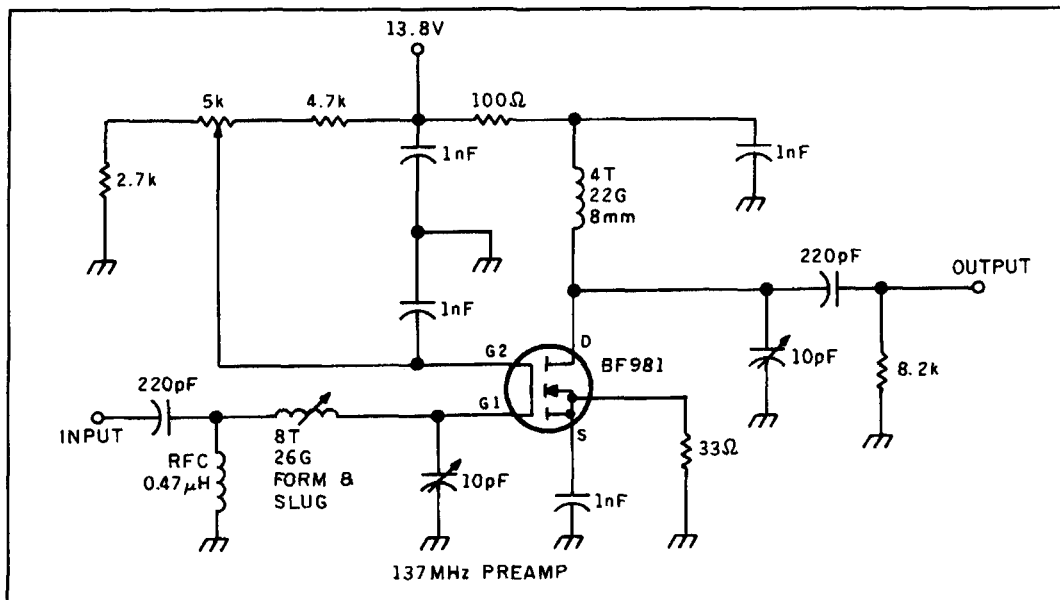


Figure 1. Final form of preamplifier.

six degrees above the horizon to overhead where the J pole gets a little weak. Below about 40 degrees, the J pole outdoes the crossed dipoles, so I will eventually use both, when I cure the preamp going AWOL in the rain.

The crossed dipole dimensions are as follows:

Dipoles	1035 mm
Reflectors	1090 mm
Spacing	810 mm

For the balun, I recommend using only RG-8 or better coax as it does make a difference.

Remember, we want a minimum loss reception system. Every effort should be made to eliminate loss. After some frustration, I can assure you that spending a bit at the start will save you redoing it all again at more expense later.

I had the good fortune to try the Instant Track program, which shows path loss of a particular satellite as well as sky noise in its direction. At my location, the APT birds have between 130 and 148 dB path loss. Although stronger than most amateur craft, the signals are still weak and care is needed to get noise free pictures.

## Preamps

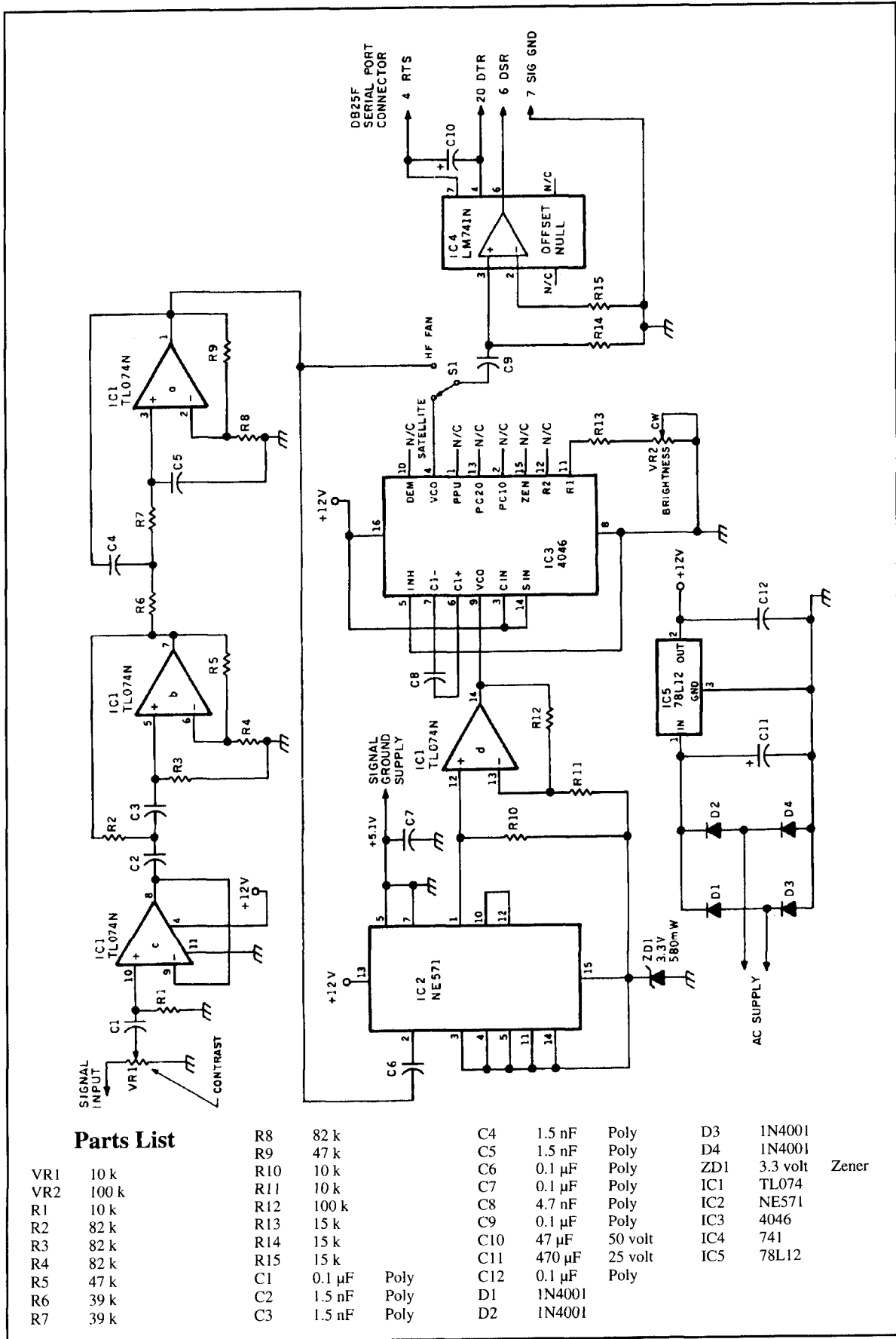
I have tried many different preamps while receiving APT. I live only one mile away from 100 kW FM and TV broadcast transmitters which get right into my system, especially when the required bandwidth is 50 kHz. Also, when using my favored BF981 preamp in conjunction with a J pole, the FET would blow up every time it rained, which was about five times in one week. I tried "back-to-back" diodes across the input, but the noise figure became shocking, so these were removed. My final choice is shown in **Figure 1**. It gave supe-

rior results to all the others—especially when constructed as shown. The construction was mostly from Murray, ZL3TIB, and proved very reliable and reproducible. The actual circuit is a composite from five or more circuits I've seen and tried, and the data obtained from the transistor data books. Doubtless it could be improved on, but it works well in my harsh radio environment.

## Receivers

I tried my R7000 and got some reasonable pictures in the wide FM mode, which is about 150 kHz bandwidth. However, any slight fade produced noise interference on the picture. In the narrow FM mode, the bandwidth was far too small to be of use. I could have replaced the wide FM filters with 50 kHz ones, but these filters are as scarce as hen's teeth. The only place I could locate these was in Australia, and I have still not heard from the supplier after four months.

I tried my Pro 2006 scanner. Once again the wide FM was too wide, but the narrow FM mode was a little wider than the R7000, so I could get some images when decoding signals from the Meteor series of satellites. I was not satisfied and wanted better results from both series of satellites. Studying the workshop manual on this scanner produced some options: (1) install 50 kHz wide FM filters which I can't get, (2) remove narrow FM filter and replace it with a 0.01 capacitor. Option 3 works well, and I can get noise free pictures horizon-to-horizon from both Meteor and NOAA spacecraft. I put in a switch to select either the capacitor or the narrow FM filter when I want the scanner for use as it was intended. I'm not sure what the IF



**Parts List**

VR1	10 k	R8	82 k	C4	1.5 nF	Poly	D3	1N4001
VR2	100 k	R9	47 k	C5	1.5 nF	Poly	D4	1N4001
R1	10 k	R10	10 k	C6	0.1 μF	Poly	ZD1	3.3 volt Zener
R2	82 k	R11	10 k	C7	0.1 μF	Poly	IC1	TL074
R3	82 k	R12	100 k	C8	4.7 nF	Poly	IC2	NE571
R4	82 k	R13	15 k	C9	0.1 μF	Poly	IC3	4046
R5	47 k	R14	15 k	C10	47 μF	50 volt	IC4	741
R6	39 k	R15	15 k	C11	470 μF	25 volt	IC5	78L12
R7	39 k	C1	0.1 μF	C12	0.1 μF	Poly		
		C2	1.5 nF	D1	1N4001			
		C3	1.5 nF	D2	1N4001			

Figure 2. Interface schematic.

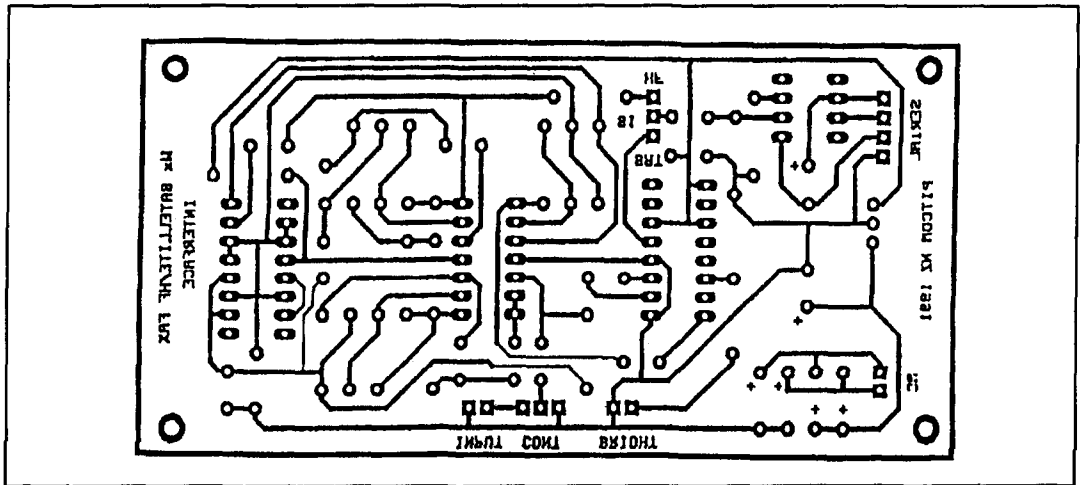


Figure 3. PC board, trace side..

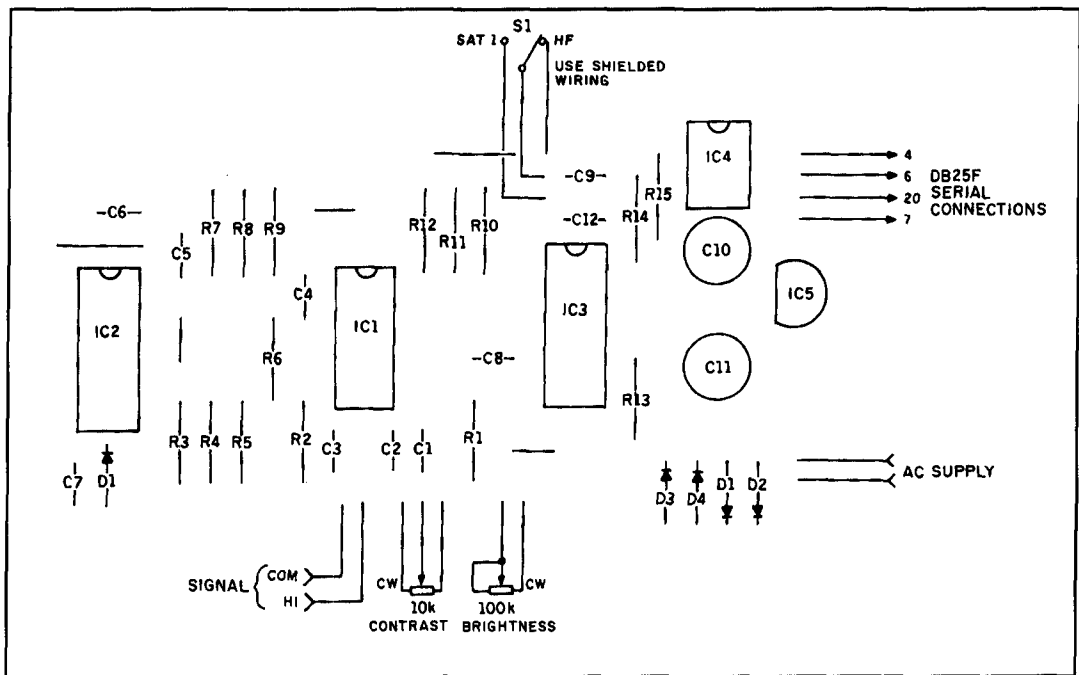


Figure 4. PC board, component side.

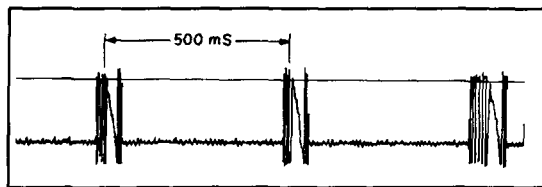
bandwidth is with the capacitor, but it does the trick and I'm sure this option would work for a number of other scanners pressed into service for APT reception.

### The interface and decoding pictures

There are many amateurs decoding weather fax pictures using PK232s, Kanterms, or a number of homemade terminals. We have had great success with many pictures decoded from Australia and New Zealand regularly, some good pictures in a simulated grey scale during the Gulf crisis being transmitted from Guam. These simulated greyscales are already dithered to form a black and white image ready for transmission on a standard black or white only

setup. Also we have had some success with real greyscale pictures from Hawaii, which are satellite photos retransmitted on 23.331 MHz.

There are several programs for PCs which can decode greyscale: AEA-Fax, HF-Fax, and one describe in *QST* in September 1991 by K3BC and available free from him. From experience, I recommend programs that have a tunescope which allows for easy tuning into a fax picture—especially when our interface is used. I have tried HF-FAX and it works well. It was purchased in the USA several years ago and may differ from the one advertised more recently in *Break In*. The standard interface for this and other software is just a 741 op amp connected directly to an RS232 port, as can be seen in our interface. All we have added is



**Figure 5. Sample tune scope display with White Level = 2202 Hz, Black Level = 1400 Hz.**

additional circuitry to perform decoding of APT data. Our interface should work with any greyscale program.

The interface converts the amplitude modulated 2400 Hz video signal from the satellite into a 2200 Hz/1400 Hz standard fax signal. The interface shown as **Figure 2** is pretty simple and quite cheap at around \$20–25, which compares well with alternatives for APT reception. David, ZL3TPD, had the basic design worked out and a prototype under construction within one day. The results were nothing short of spectacular. After more discussion and one mod to the circuit plus a comment, “wouldn’t it be nice to have a PC board for this circuit.” David did it again. The board layout was made, and I produced some boards which David made up (see **Figures 3** and **4**). It worked the first time and no other mods have been required to date. A fantastic achievement by David, all credit going to him for getting this interface going.

On air adjustment is straightforward, especially if a tune scope is available. Tune in an APT signal, then adjust the two controls to suit. Brightness controls the center frequency of “fax” signal, while the contrast sets the deviation from the center frequency. A little playing with these controls soon gives you the feel. See **Figure 5** for examples of what to expect to see using a frequency versus time display as on a tune scope.

The only defect encountered to date is the inability to automatically start the picture receiving process. This is because the satellite continuously transmits the picture information with no start or stop tones, as seen on normal weather fax transmissions. I intend to use one of the older Exar chips used in RTTY to give the required start tone sequence of 2200 Hz and 1400 Hz alternating at 400 Hz when a satellite is heard. However, this may take some time because of many other small projects I have on the go. If you beat me to it, please let me know via ZL3TCM @ ZL3AC.

## Displays

I started with a PC using a Hercules graphics card in conjunction with a CGA emulator, and got fairly useless results from HF fax signals from Hawaii. This setup gave only four levels

of grey and was not very impressive when viewed on an orange monochrome screen. On an EGA display (all I could afford), the results were more impressive, as long as I custom-colored the picture using a palette of black, blue, grey, etc. Any other fixed palette, such as monochrome or false color, gave sharp, distinctive gradations of level, but were adequate if you captured the entire picture before you tried to make out any land mass. I tried printing out some pictures on a nine-pin dot-matrix printer with mixed results, then a laser printer with very good results. This gave 300 dots per inch resolution, was every bit as good as the EGA display, and was in genuine greyscale.

However, it didn’t take long before I wanted to upgrade. I saved up and sold the EGA to get a VGA display and the pictures are fantastic now when received in monochrome. The pictures are near photographic quality, but the print quality hasn’t improved, so now the display surpasses the hard copy—even with the laser printer.

I’m currently experimenting with photographing the actual monitor screen. First efforts involved using standard color film and local processing at one of the “one hour” stores, which was a bit of a waste of time, because the pictures are best viewed in monochrome. The only advantage to using false color would be to distinguish between close levels. Better results would be obtained from a film like Ilford FP4 black and white, and custom processed, but I no longer have a darkroom, so I’m not sure what my next step will be. You can use any camera capable of timed shutter control of around one second, set up on a tripod in a dark room with only the monitor on.

## Summary

This project has occupied a few of us for some months now, and some excellent pictures have been obtained. It’s fun comparing what you have received with what is shown on the TV weather broadcast. This hardware interface would have to be one of the cheapest ways of getting good quality pictures.

Lastly, I would like to thank Graham, ZL3OY, David, ZL3TPD, and Murray, ZL3TIB, for their assistance and information, which has allowed this project to be a success.

## Pliers-Type RF Current Probe

*Reprinted with permission from Pat Hawker’s “Technical Topics” column, Radio Communication, October 1992.*

**In an earlier “Technical Topics,”** Dave Plumridge, G3KMG, mentioned that he had seen “many years ago in TT” a device involving a spring-type of clothes peg with two half toroids

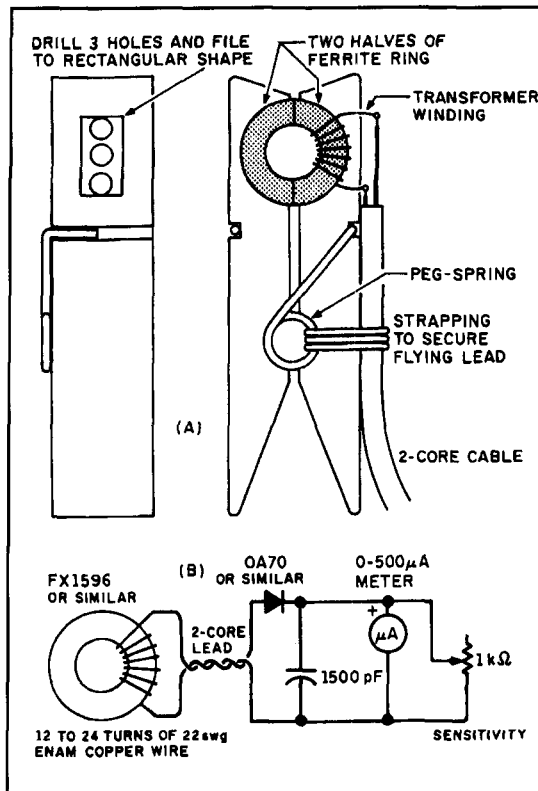


Figure 1. (A) Basic construction details of G3ZHY's peg sensor as described in 1972. (B) The complete ferrite ring RF meter.

which could be clipped round a coax cable to form a current transformer with diode rectifier and meter, that could be used to measure the current flowing on the outer braid of a coax feeder cable (or the RF current flowing in an antenna wire-element or an open-wire feeder).

This brought in some correspondence. Colin Greenaway, G3UGG, Eric Sandys, G12FHN, and Malcolm Horton, G4DMH, explained why I had not been able to trace the item in TT. The reason was simple; it was not in TT, but described in an article by M.R. Irving, G3ZHY, called "The Peg Antenna Meter," *Radio Communication*, May 1972, pages 297 and 301. G3ZHY pointed out that he had used a small toroid that would fit over a wire conductor or small-diameter coaxial cable, but the two halves would not close on thicker cables, and his device had not been intended to work above about 3.5 MHz. He had noted that "the only items needed are those components shown in **Figure 1**, some fixing cement such as Plastic Padding, together with hand-drill, hacksaw, and a small file. A little patience is required when cutting the ferrite ring, and two attempts were necessary before I obtained a satisfactory clean break. Those seeking perfection should try to obtain the use of a jeweler's diamond saw."

Dick Rollema, PA0SE, also noticed G3KMG's remarks and has sent along full

details of a rather more ambitious pliers-type RF current probe (**Figure 2**) that he described in his "Reflecties door PA0SE" column in *Electron* (VERON) in July 1990. He writes: "Such devices are widely used by the 50 Hz power people. In German it is called a stromzange, in Dutch a stroomtang (i.e., current pliers).

"When the toroid is clamped over a conductor the latter acts as a one-turn primary of a current transformer, the winding on the toroid core forming the secondary. Thus, in **Figure 2**, the secondary current flows through R, resulting in an RF voltage across R. This is rectified by diode D and, after the bypass capacitor C, fed to a high-resistance voltmeter; a digital type is easy to read. (Note that the voltmeter must be far enough from the source of RF to ensure that there is no direct pickup. G3VA.) To make the reading independent of frequency, the reactance of the secondary winding should be at least  $5R = 5 \times 56 \text{ ohms} = 280 \text{ ohms}$ .

"I used a 'violet' Philips toroid of 38 mm outer diameter, 4C6 ferrite,  $\mu = 125$ . The measured self-inductance of the winding is  $15.6 \mu\text{H}$  (before splitting the ring it was  $17.7 \mu\text{H}$ , showing the effect of the two air gaps, even though the two halves fit together perfectly). The  $15.7 \mu\text{H}$  inductance results in a reactance of 345 ohms at 3.5 MHz, so is more than enough. On 1.8 MHz, the reactance is only 177 ohms, introducing an error of almost 5 percent, but nevertheless is still a useful sensor.

"The difficult part is splitting the ring. First make two sharp notches opposite each other, using a file (a thin diamond disk would be ideal if you are on good terms with your dentist). The toroid is then clamped with its lower half in a vice. A piece of wood is put against the upper half and struck a blow with a hammer. With luck, a clean fracture . . . without it, try again. My second try was successful. After the secondary winding has been wound on, the two halves are fitted to the 'pliers' with some epoxy.

"I made my pliers out of paxolin, but other materials such as perspex, suggest themselves. The toroid is kept closed by a spring and opens when the pliers are squeezed. Keep the two halves in the closed position while the epoxy cures. Be very careful that no epoxy gets into the gaps between the two parts of the ring (voice of experience)!

"To calibrate the sensor, I used the arrangement shown in **Figure 3**. The sensor is clamped over the upturned U-loop. The transmitter should have variable CW output. R is a 50 ohm dummy load. If the meter, V, shows a voltage of  $U_{dc}$ , then the current through the loop is  $I_{rms} = (0.707(U_{dc} + 1.2 \text{ volts}))/50 \text{ ohms}$ . The 1.2 volts accounts for the voltage drop across the two silicon point-contact diodes. By varying the transmitter output (do not exceed the maximum per-

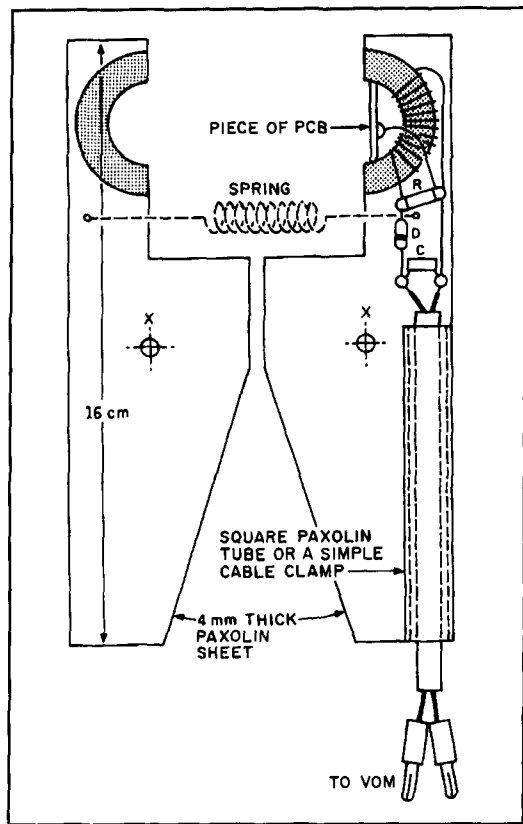


Figure 2. Details of PA0SE's pliers-type RF current probe. R is 56 ohms, 0.25 watts. D is a germanium point-contact diode. C is a 10 nF ceramic. The Philips toroid is OD 37 mm, ferrite 4C6  $\mu$  125, violet. The winding is 10 turns of 27 SWG enamelled wire. Inductance is about 17.5  $\mu$ H.

missible voltage across the diodes) the DC output voltage of the sensor can be plotted against the primary RF current. I found a perfectly linear response when the RF current was 50 mA or more. Below that value, the diode in the sensor introduces some nonlinearity. The conversion factor in my case worked out at 5 volts DC per 1 A RF current. The response proved frequency independent up to at least 30 MHz.

"I have found this RF current probe a most useful device. I use a long-wire antenna about 37 meters long on 1.8, 3.5, and 7 MHz. The wire is fed directly at one end via an L-network. The 'cold' side of the network is connected to a radiator of the central heating system. The current sensor shows that indeed most of the current flows via the radiator pipe to earth. But considerable current also flows via the mains cord of the transmitter with even some via the earthed triangular metal mast that supports one end of the antenna. It is interesting to note that the current does not divide evenly over the structural elements of the mast!

"I have also used the current probe to check the current in the loops of my 'Optiquad' antenna. It was gratifying to note that on 28,

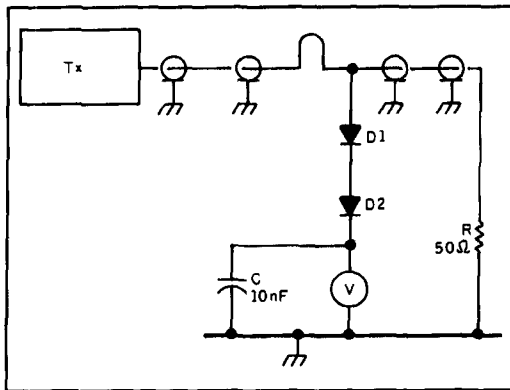


Figure 3. Arrangement used for calibrating the probe. D silicon point-contact diodes, R 50 ohms, C 10 nF ceramic capacitor.

21, and 14 MHz, where I made my measurements, current maxima occurred at the expected positions and also that the current in the left and right halves of the loop were equal. This showed that the baluns were performing correctly, although I already knew that because the sensor detected no current on the outside of the coaxial feeder cables connected to the baluns. (The baluns used by PA0SE were described in *Eurotek* by G4LQI, *Radio Communication*, August 1992, page 51.)

"To sum it up, I consider that an RF current probe of this type is a must for serious antenna experiments."

Wire-loop current probes and their uses are described by Les Moxon, G6XN, in *HF Antennas for All Locations* (page 237) and a ferrite probe with a fixed gap in the toroid was included in *TT*, November 1984, stemming from Dick Kelsall, G4FM. This was basically similar to the PA0SE probe, but without the advantage of being able to open and close the gap so that it could be used with large diameter cables, etc. G4FM also experienced the problem of cutting ferrite cores without fracturing them. His solution was to cast the ring in polyester resin (from a glass fiber repair kit) inside a suitable piece of plastic tubing resting on a sheet of glass laminate. When the resin was set, he was able to file a slot through the resin and ferrite core without difficulty.

## Decoupling Capacitors—Why Use Two When One Will Do?

Some common circuit techniques tend to be accepted as the thing to do without anyone questioning them. Once the technique becomes common practice, it acquires an air of authority and a momentum of its own. One such technique could be the use of a parallel combination of large and small-value capacitors to extend the effectiveness of RF bypassing over

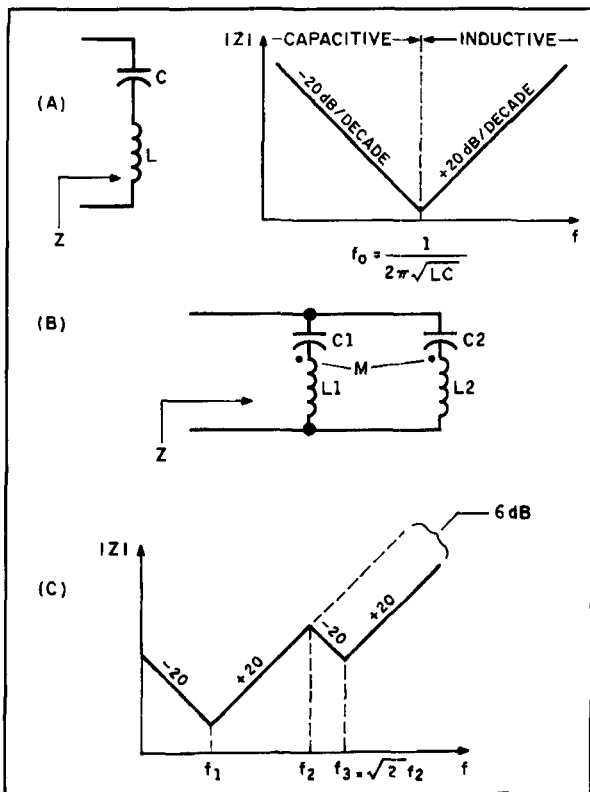


Figure 1. (A) The effect of lead inductance on capacitor impedance with minimum impedance at the self-resonant frequency. (B) Equivalent circuit for parallel decoupling capacitors and the Bode plot of impedance.

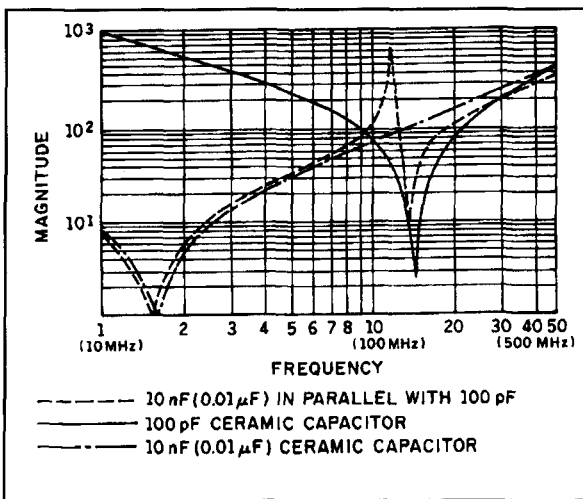


Figure 2. Measured impedance of a 0.01  $\mu\text{F}$  and a 100 pF ceramic capacitor individually and in parallel from 10 to 500 MHz. (Source: Paul, *IEEE Trans EMC*.)

an extended frequency range.

The technique probably stemmed from the long established system of connecting an RF-type capacitor (ceramic or mica, etc.) across a high-value electrolytic capacitor when it is necessary to provide RF bypassing/decoupling. This

was (and is) an effective system because electrolytic capacitors tend to be very ineffective for RF bypassing, although providing low impedance at 50 to 100 Hz and audio frequencies. It must therefore have seemed logical to adopt this approach when providing RF bypassing in low-impedance solid-state circuits such as power amplifiers when an RF decoupling capacitor may need to be on the order of 0.1  $\mu\text{F}$  or so and capable of passing significant RF current.

But should capacitors be augmented by a second, lower-value capacitor? This is common practice and a typical arrangement is shown, for example, in the 15-watt linear amplifier in the highly respected ARRL book *Solid State Design for the Radio Amateur* (Chapter 4).

Dr. Brian Austin, G0GSF, draws attention to a paper "Effectiveness of Multiple Decoupling Capacitors" by Clayton R. Paul (University of Kentucky) in *IEEE Transactions on Electromagnetic Compatibility*, May 1992, pages 130 to 133. This calls into question the effectiveness of using a parallel combination of large and small-value capacitors to increase the effectiveness at the higher frequencies and to overcome the effect of lead inductance. With detailed mathematical analysis, computed and experimental results, the author shows that this scheme is *not* significantly effective (see **Figures 1 and 2**). The improvement at high frequencies is at most 6 dB over the use of a single large value capacitor.

The paper concludes that the use of a small-value capacitor in parallel with a larger-value capacitor only minimally reduces the high-frequency impedance of either capacitor. A rather large reduction in impedance occurs only over a very small frequency range. Component and installation variations can cause this narrow range to vary in frequency so that the seeming benefit cannot be relied upon: "Therefore, this scheme does not significantly improve the high-frequency impedance of capacitors above their self-resonant frequencies. Between the self-resonant frequencies of the two capacitors, the impedance of the parallel combination exhibits a resonance where the impedance is actually larger than that of either capacitor. Above the self-resonant frequencies of both capacitors, the impedance of the parallel combination is reduced by at most 6 dB. This high-frequency reduction of some 6 dB may not be worth the expense of the additional capacitor or its installation, and could be obtained by using only the larger value capacitor while simply cutting its lead lengths in half!"

As G0GSF puts it: "This paper presents a good case against doing what we have all taken for granted." ■



# TRANSMITTING SHORT LOOP ANTENNAS FOR THE HF BANDS: PART 2

*A short loop antenna for 20 to 10 meters*

In part 1, I emphasized the need to keep ohmic losses in loop antennas at a minimum.<sup>1</sup> Keeping this in mind, I built a short loop with some innovative elements not used in conventional loops built before mine.

On a visit to the International Hamfest in Friedrichshafen, Germany, I found an excellent vacuum capacitor of Soviet Army origin. The capacitance is 100 pF and the insulation is 20 kilovolts. For my loop conductor, I decided to try a piece of discarded Heliac coax cable. This cable is excellent for this particular use. The external diameter is about 1-3/4 inches. Because it's made of copper plate with a special knurling, the outer conductor is easy to bend by hand when forming the circular loop. This helical knurling increases the conductive surface about 20 percent. The inner conductor isn't used.

## Main characteristics of the loop

To obtain magnetic radiation from the loop, I chose a conductor length of 10 feet, 4 inches—slightly more than 1/4 wavelength on 28 MHz. The corresponding diameter of the loop is 3 feet, 3 inches. These measurements aren't critical; variations of a few inches have no effect on the



CALCULATION OF MAIN PARAMETERS OF A SHORT LOOP ANTENNA

(copper conductor)

Loop conductor length (max. 1/3 lambda)		feet: ?	10.3
Loop conductor diameter		inches: ?	1.75
Frequency		megahertz: ?	14
Power		watt: ?	100
For circular shape loop, type	1		
For octagonal shape loop, type	2		
For square shape loop, type	3		
? 1			
loop area		sq. feet:	8.381111
Radiation Resistance		milliOhms:	9.120782
Loss Resistance		milliOhms:	2.193424
Radiation Efficiency (basis ohmic loss)		%:	80.61353
Efficiency expressed in dB		dB:	-9359204
Loop Inductance		microHenry:	1.995049
Inductive Reactance		Ohms:	175.4941
Quality Factor		Q:	775.5475
Bandwidth		kHz	18.05177
RF Voltage across capacitor		kiloVolts:	3.689227
Tuning Capacitance		picoFarads:	64.77817
Distributed Capacitance		picoFarads:	8.446
Ok			

CALCULATION OF MAIN PARAMETERS OF A SHORT LOOP ANTENNA

(copper conductor)

Loop conductor length (max. 1/3 lambda)		feet: ?	10.3
Loop conductor diameter		inches: ?	1.75
Frequency		megahertz: ?	21
Power		watt: ?	100
For circular shape loop, type	1		
For octagonal shape loop, type	2		
For square shape loop, type	3		
? 1			
loop area		sq. feet:	8.381111
Radiation Resistance		milliOhms:	46.17396
Loss Resistance		milliOhms:	2.686385
Radiation Efficiency (basis ohmic loss)		%:	94.50191
Efficiency expressed in dB		dB:	-2455942
Loop Inductance		microHenry:	1.995049
Inductive Reactance		Ohms:	263.2411
Quality Factor		Q:	269.3811
Bandwidth		kHz	77.95647
RF Voltage across capacitor		kiloVolts:	2.662934
Tuning Capacitance		picoFarads:	28.7903
Distributed Capacitance		picoFarads:	8.446
Ok			

CALCULATION OF MAIN PARAMETERS OF A SHORT LOOP ANTENNA

(copper conductor)

Loop conductor length (max. 1/3 lambda)		feet: ?	10.3
Loop conductor diameter		inches: ?	1.75
Frequency		megahertz: ?	28
Power		watt: ?	100
For circular shape loop, type	1		
For octagonal shape loop, type	2		
For square shape loop, type	3		
? 1			
loop area		sq. feet:	8.381111

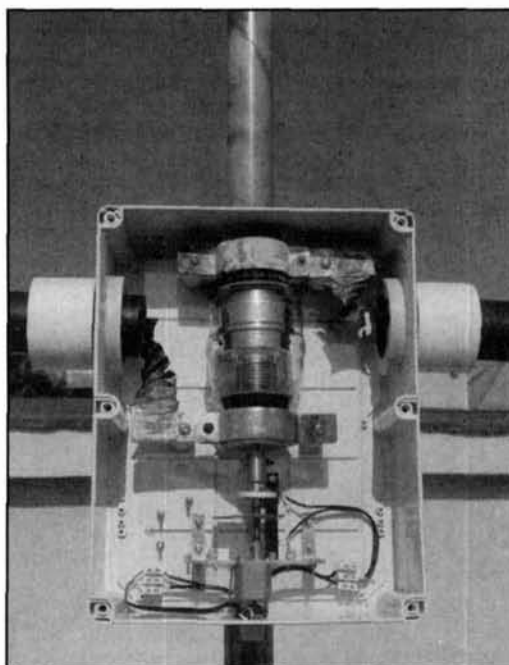
Radiation Resistance	milliOhms:	145.9325
Loss Resistance	milliOhms:	3.10197
Radiation Efficiency (basis ohmic loss)	%:	97.91862
Efficiency expressed in dB	dB:	-9.134741E-02
Loop Inductance	microHenry:	1.995049
Inductive Reactance	Ohms:	350.9881
Quality Factor	Q:	117.754
Bandwidth	kHz	237.7839
RF Voltage across capacitor	kiloVolts:	2.032984
Tuning Capacitance	picoFarads:	16.19454
Distributed Capacitance	picoFarads:	8.446
Ok		

**Table 1.**

final result. However, I do suggest you use a smaller diameter loop if you use a conventional capacitor. The residual capacitance of a conventional capacitor is higher than that of a vacuum capacitor and, consequently, you'll be unable to obtain resonance on 28 MHz. A loop with a conventional split stator/butterfly capacitor needs to be about 2 feet, 8 inches in diameter to resonate properly at 28 MHz. **Table 1** gives the main parameters of the antenna calculated using GWBasic for operation on 10, 15, and 20 meters.

## The tuning unit

My tuning unit is based on the vacuum capacitor I used (see **Photo A** and **Figure 1**). Obviously, modifications will be necessary for other types of capacitors. The unit must be mounted inside a watertight box. A plastic box for outdoor electrical use provides a simple and inexpensive solution. The size of the box depends on the capacitor used and the dimension of the tuning motor coupled to the capacitor's tuning shaft via a ceramic coupler. These boxes aren't strong enough to withstand the dynamic stresses originating at the ends of the loop conductor connected to the vacuum capacitor. Strengthen the bottom and sides of the box with Plexiglas™ boards about 3 mm thick. Build the tuning unit directly on the bottom reinforcing board, place it inside the box, and affix it to the box with nuts and bolts (see **Figure 2**). This attachment scheme also helps keep the collar clamps on the back of the box in place. Now connect the collar clamps to a strong plastic pipe supporting the entire antenna (**Figure 3**). Mark a center line on the Plexiglass board; you must align the motor and capacitor perfectly along this line. Connect the rotor and stator contacts of the vacuum capacitor to the ends of the loop using copper-plate collar clamps built in such a way that they also support the capacitor with either Teflon™ or ceramic standoffs. This step lets you achieve two important objectives at once. First, all the



**Photo A.** A closeup of the tuning unit.

dynamic stresses from the loop ends are placed on the standoffs—not the vacuum capacitor body. Second, it's possible to braze the loop ends to the collar clamps before placing the capacitor in position. As a result, you won't have to risk damaging the capacitor with the brazing heat.

Next, cut the loop ends lengthwise about 5 inches and eliminate half the conductor. Clear the remaining conductor half from the vinyl jacket; this will act as a terminal strip and will be brazed to the copper clamps. The flexible strip will also absorb dynamic stresses. At the same time, a single joint is all that's needed to connect the loop end—reducing ohmic losses. The loop ends enter the box through two holes made on each side of the box. These large holes will weaken the plastic and you'll need to place two thick Plexiglas boards along the sides con-

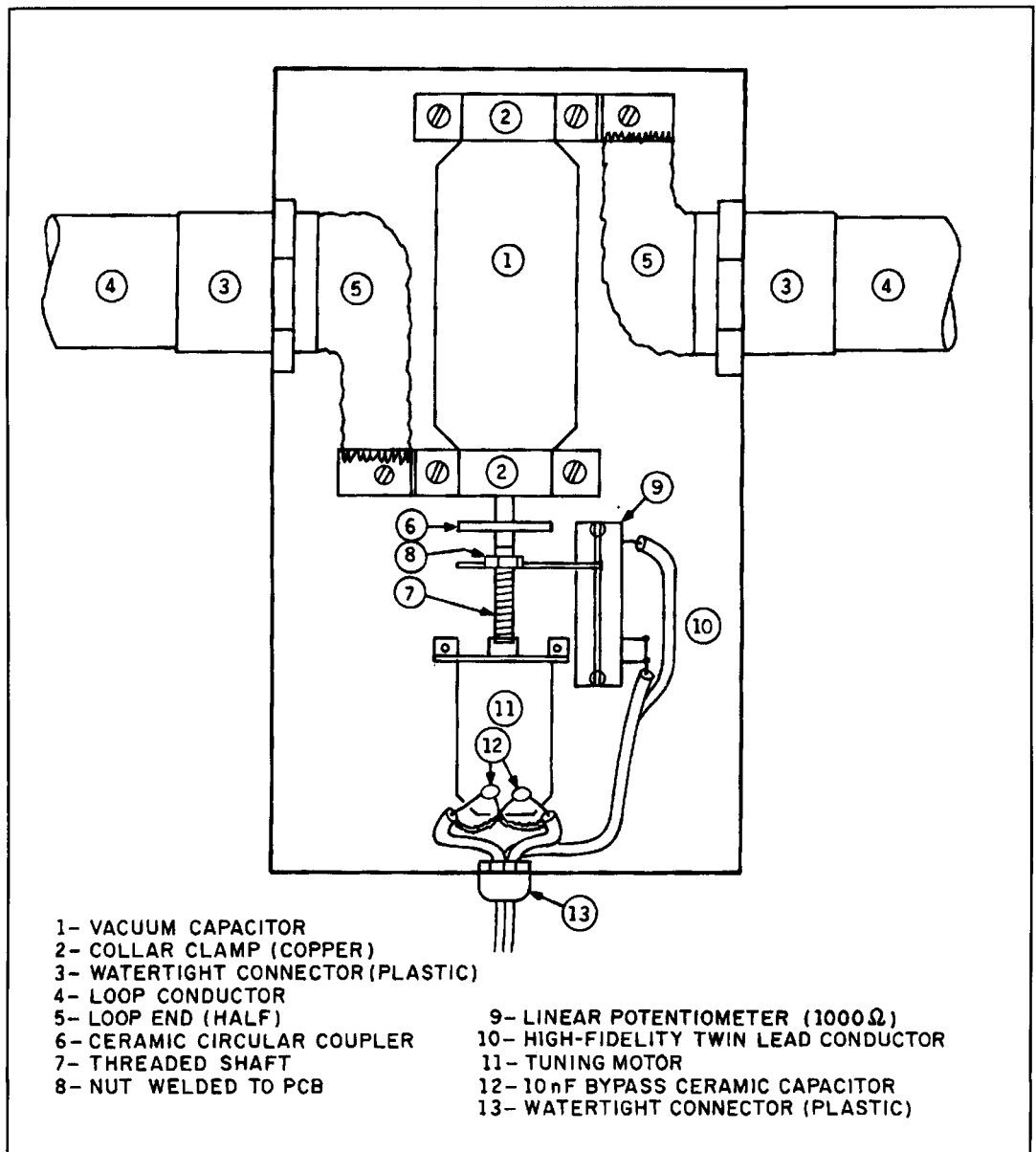


Figure 1. The tuning unit.

taining the holes for reinforcement. To keep the entrance points watertight, use standard connectors of adequate diameter to accept the conductor. These standard connectors are plastic and are kept in place on the side of the box with a large plastic nut. This nut also holds the Plexiglas reinforcing plate inside the box. For my antenna, I used specially made Teflon connectors instead of standard connectors. However, Teflon connectors aren't absolutely necessary; standard connectors work very well.

### Tuning motor and tuning indicator assembly

A reliable tuning system requires a good quality reduction gear DC motor. The maxi-

mum revolution speed of the shaft shouldn't exceed 5 rpm. Vacuum capacitors have a built-in reduction gear, reducing the speed further. If you use a different type of capacitor, you must add an external reduction gear to obtain a shaft speed of 1 rpm. The electronic device in **Figure 4** provides an alternative method of controlling motor speed. The DC motor should have a wide range of operating voltages (3 to 12 volts). Use the lower voltage for fine tuning and the higher voltage for band changes. The Buhler motor in my tuning system is a surplus unit from a computer printer.

It's important to position the motor on the tuning board precisely. Make sure the capacitor shaft and motor shaft are perfectly aligned, or eccentricity and subsequent forcing will be introduced along with the possibility of damage

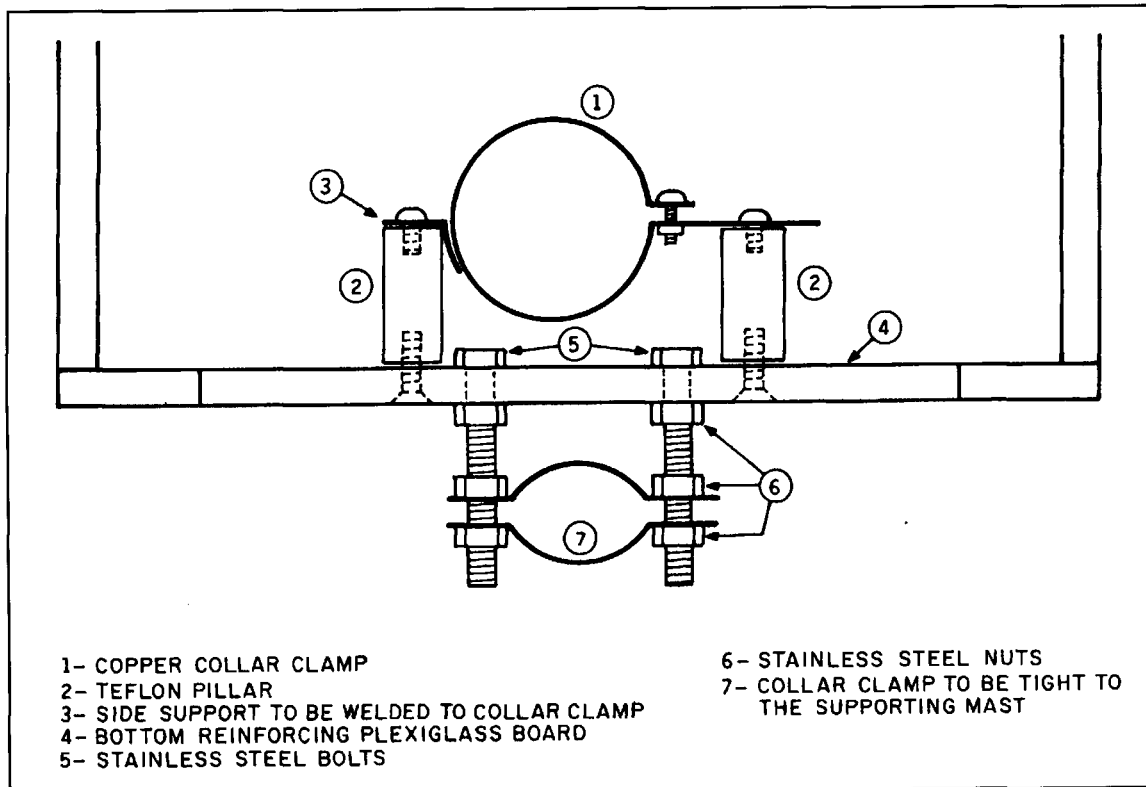


Figure 2. Tuning unit detail

to the components. The ceramic shaft coupler should be circular with center springs, so some small eccentricity can be absorbed. For details see the tuning motor and indicator assembly in **Figure 5**.

Considering that the vacuum capacitor has two stop positions (fully open and fully closed), it's important to know what the tuning position is from inside the shack to avoid any forcing at the stops. You could reverse the turning sense using a microswitch; however, I used another method that gives me a visual indication of the capacitor's tuning position. I inserted a threaded shaft with a nut on it between the motor shaft and the ceramic coupler. This nut is soldered to a small piece of Vetronite circuit board connected to a linear potentiometer. Calculate the length of the threaded shaft according to the number of turns needed to reach the opposite stop ends of the capacitor. The nut travels back and forth according to the revolution sense of the motor and, at the same time, moves the wiper of the resistor. With an adequate circuit in the shack, you can read the variation of the resistance by looking at a meter; consequently, with proper plotting, you can determine the tuning position. Use shielded twin lead for the wiring between the tuning box and the bottom of the loop and ground the external braid at the bottom. The same system must be used for the motor. Bypass the motor contacts to the motor body and the braids with two 10-nF bypass

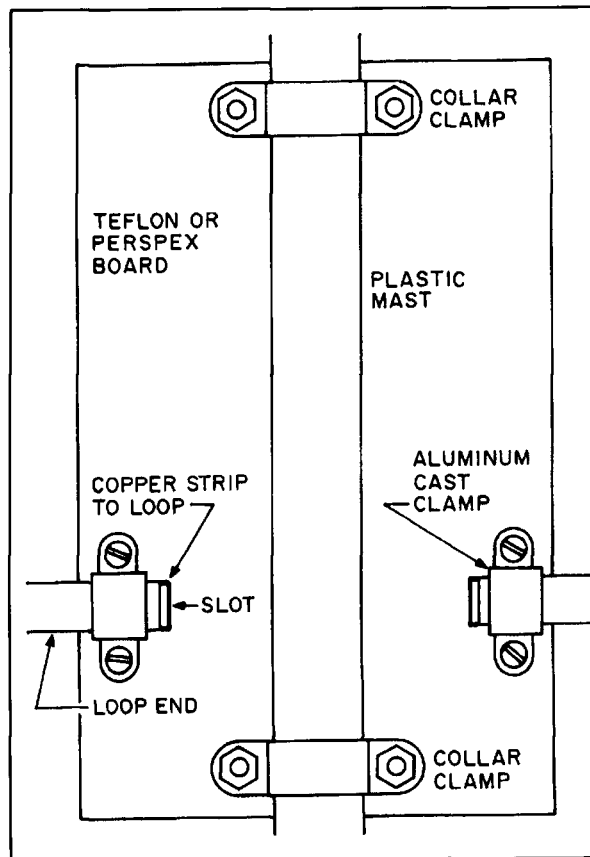


Figure 3. Back view of the tuning board showing mountings. Figures 3-7 first appeared in *Radio Communication*.

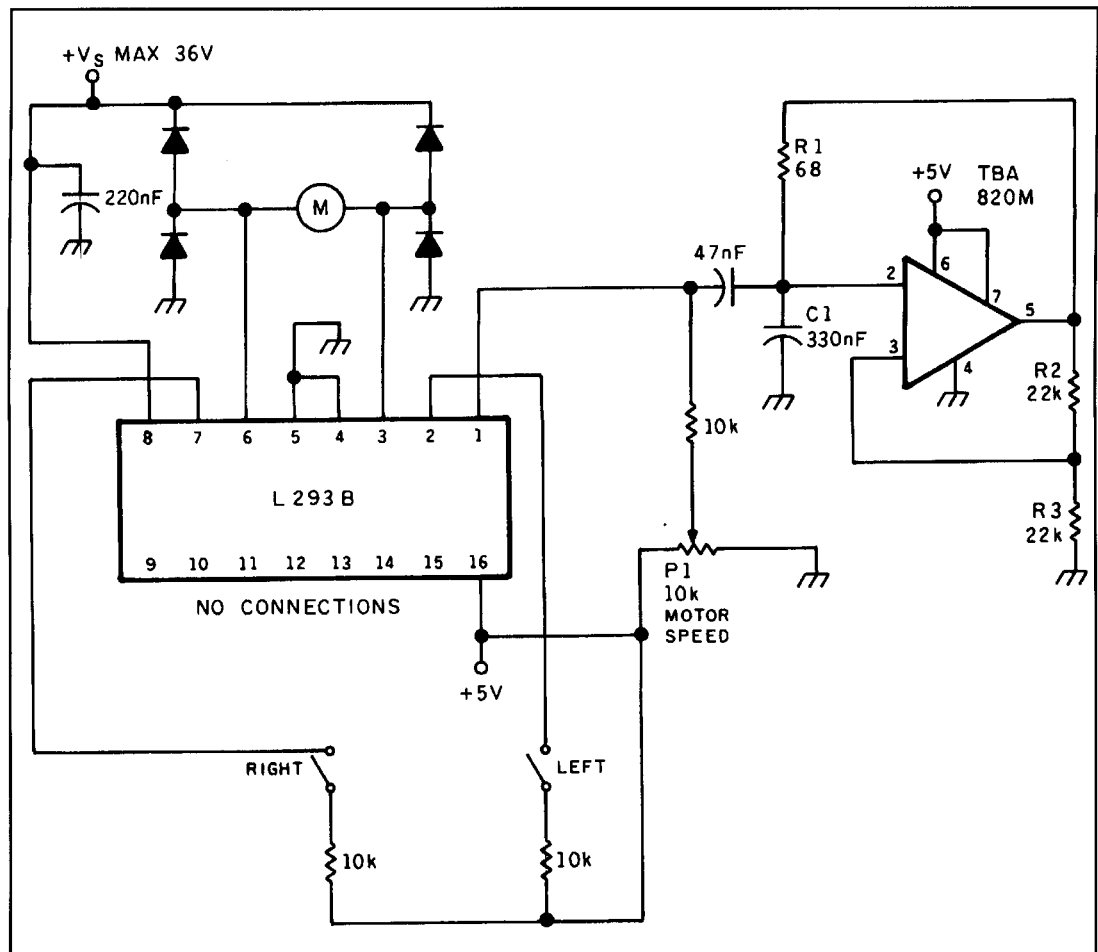


Figure 4. Tuning motor speed regulator.

capacitors (see **Figure 6**). Fix the twin lead to the loop supporting pipe with plastic clamps. You can use conventional twin lead from the bottom of the loop to the shack.

## The feed loop

My experience, and the experience of others, indicates that a short loop antenna is best fed by inductive coupling.

A closeup of the feed loop is shown in **Photo B**. You can make it out of either RG-213 or RG-8 coax. Determine the circumference experimentally to obtain the lowest SWR. This length should be approximately 1/5th of the length of the main loop. Because the antenna is a balanced system in respect to the center line of the loop, make sure the feed loop is symmetrical. The coax line forming the feed loop is open at the top center. The braid on the input side of the opening is unconnected. The inner conductor and braid on the other side of the opening are shorted and grounded together with the braid of the input side.

Affix the feed loop at the bottom of the main

loop in the same plane, and keep it very close to the main conductor. Use two stainless steel hose clamps interconnected at 90 degrees, to keep the feed loop in place. Attach one clamp tightly to the steel or aluminum pipe covering the bottom of the plastic supporting mast across the loop. Screw the other clamp tightly to the bottom terminal joint of the feed loop before inserting it into the first clamp. This mounting system (see **Figure 7**) ensures that all the bottom parts of the loop and feed loop are grounded. An efficient earth connection is essential to obtain a low SWR and good antenna performance.

## Use of an antenna rotator

You can turn the antenna with a TV rotator. The ability to rotate the loop can be very helpful if there's local interference. In fact, on ground wave, the loop acts as a direction finder. The attenuation between maximum and minimum is over 20 dB. A local received interfering signal can almost be eliminated. The same is true when transmitting—if you wish to radiate in a certain direction. However, with

elevation angles of a few grades or more, directivity is reduced and, in practice, skywave operation with the antenna is nearly omnidirectional (see **Figure 8**). Therefore, it isn't imperative that you use a rotator.

## Tuning procedure

Perform all adjustments with the antenna in its final position. Make sure the height of the lower side of the loop is between 3 and 6 feet away from metallic masses (fences, pipes, plates) lying on the same plane of the loop. A ground plane at the bottom improves the antenna's intrinsic pattern gain, even if this plane isn't connected to the antenna mast. Dress the coax feed line and motor line along the supporting mast no less than 3 feet from the lower end of the loop. If you don't there may be coupling with the loop and you may have difficulty adjusting for a low SWR. Make all adjustments on 21 or 18 MHz, as an average frequency between 14 and 28 MHz.

First find the loop resonance. Keep the receiver on receive and act on the tuning motor command. A sharp increase of noise in the

receiver indicates the loop is resonant at the required frequency. An S2 signal off resonance comes up to S9 when in resonance! The bandwidth is very narrow, and if the tuning speed is too fast, you risk passing over the resonance point without realizing it. When you find resonance, start the transmission adjustment. Use very low power to start—just sufficient to calibrate your SWR meter when set at maximum sensitivity. For greatest accuracy, insert your meter at the base of the loop. The resonance point produces a very sharp dip on the meter in the reflected position.

Don't use an antenna tuner. The adjustments you make will be permanent. If the SWR is too high, try deforming the feed loop slightly by applying pressure from the top downward. If the SWR decreases, the small loop is too long, and you'll have to cut off a small piece. Repeat the procedure until you find an acceptable SWR value. The SWR should be about 1:1 on 14 and 21 MHz, and 1:1.4 on 28 MHz. At this stage, try to slowly increase the input power and check the SWR again. You may need to make additional adjustments. When operating, don't keep the transmitter in the tune position

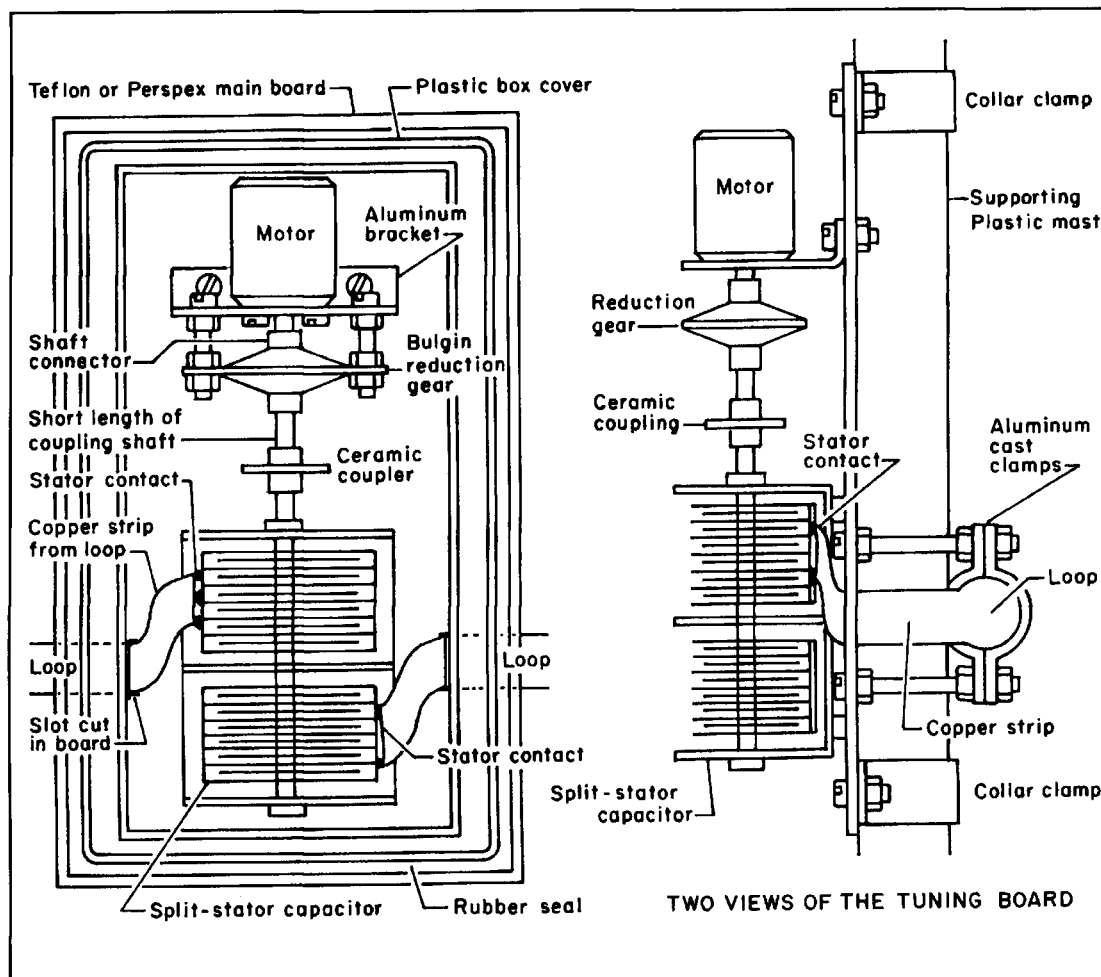


Figure 5. Two views of the tuning board.

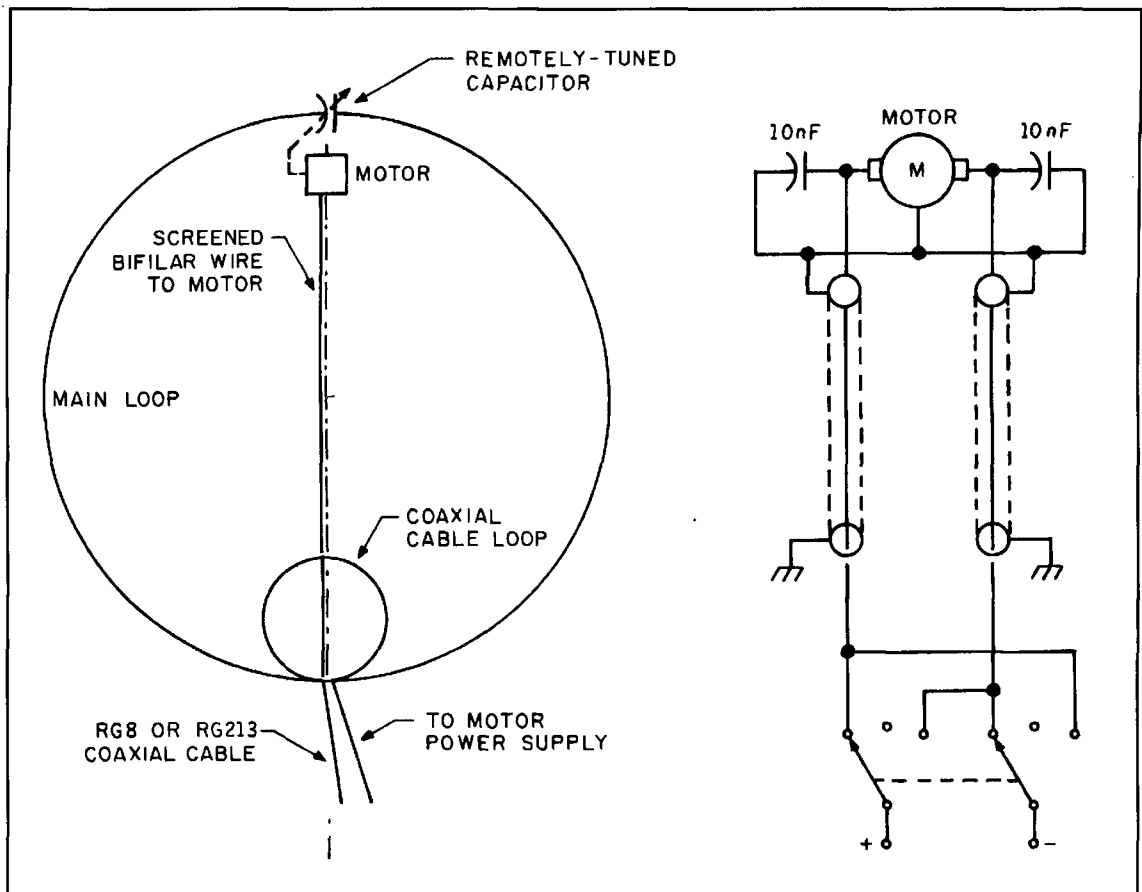


Figure 6. Electric diagram of the antenna with detail of the motor feed system.

for too long. It's better to tune on receive and than fine tune on transmit.

### An alternative method of construction

I realize it isn't easy to find suitable capacitors or the special heliax coax cable, unless you are prepared to spend a small fortune to obtain them new from the manufacturers. Nevertheless, it's possible to built excellent short loop antennas without such sophisticated elements.

The loop conductor can be made from 0.9 inch diameter copper tubing. Such tubing is used for home heating systems and is generally sold coiled. This greatly simplifies loop construction. All you need to do is mark the circumference of the loop with chalk on a flat floor, then slowly enlarge the coiled form following the chalk track. If you're unable to obtain coiled tubing, the job will be a bit more complicated because you'll have to bend the tubing yourself. Do this by first filling the tube with sand and bending it around a circular object, like a round table or car tire. You can also take the tubing to a metal shop for bending.

Make sure the tuning capacitor is a split stator

of good quality for transmitting. Remember that the calculated capacitance value must be doubled for each section of the capacitor. Both capacitor sections are in series with the loop and therefore the capacitance is automatically halved. By the same token, putting two capacitors in series doubles their voltage rating; therefore, less spacing is required to attain the desired voltage rating.

### Conclusion

To build an efficient loop antenna, I strongly recommend you use good quality materials and careful construction techniques. A good earth system is also very important. Welding must be very accurate or you'll experience unacceptable losses and very low radiation efficiency, which will lead you to believe the antenna doesn't work. However, if your work is accurate, the results will be outstanding—considering the small size of the system.

To summarize, here are the main points to consider when building an efficient short loop antenna:

1. Maximum radiation efficiency is obtained when the circumference of the loop is 1/3 wavelength. Do not exceed this length or the



loop will become self-resonant and act as an "electric" antenna with reduced efficiency.

When the loop is  $1/3$  wavelength long, a very small tuning capacitance is needed to resonate the antenna. In such a situation the distributed capacitance of the loop itself plays an important role. Remember to deduct the distributed capacitance from the calculated tuning capacitance value, otherwise it will be very difficult or impossible to obtain resonance.

If you wish to build a monoband loop, you can use a VHF butterfly or split stator capacitor. Alternatively, you can use a fixed high quality ceramic capacitor (for transmission) in parallel with a small variable capacitor covering the single band.

2. For the high bands (28 to 10 MHz), the loop material (copper tubing) should be a minimum of 0.9 inches diameter. For the low bands (7 to 1.8 MHz) the diameter should be 1.5 inches or more. Before installing the antenna, carefully polish the conductor so the copper is shiny. Apply two or three layers of marine flatting or another inert product to protect the conductor from oxidation. Copper oxide is a bad conductor and will cause the radiation efficiency to deteriorate due to increased ohmic resistance.

3. For best results, carefully weld the loop ends to the capacitor contacts.

4. Always use capacitors (split stator or butterfly) with plates welded to the supporting bars and shaft.

5. To avoid difficulties with the capacitor end stops, use capacitors with rotors that rotate freely for 360 degrees instead of 180 degrees.

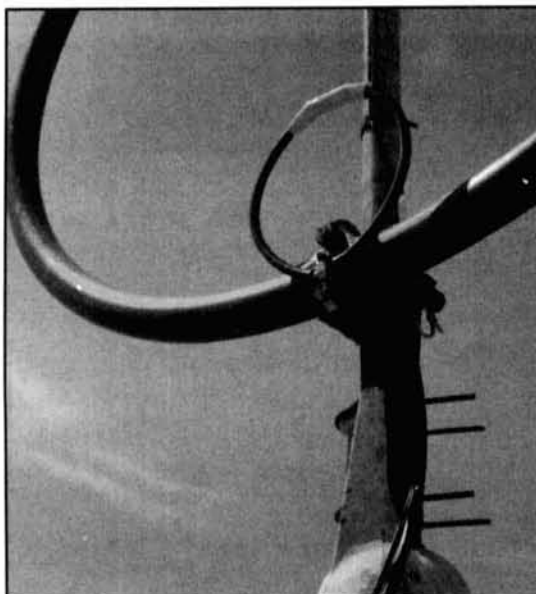


Photo B. A closeup of the feed loop.

6. The tuning motor feed line must be dressed along the center line of the loop. Use shielded wire.

7. Mount the loop at a minimum height of 3 feet. The coax and motor feed line must run vertically at the bottom of the loop.

8. The loop supporting mast must be encased in plastic; however, the lower part should be inserted in a metal tube so the center bottom portion of the loop is grounded together with the ground connections of the coax feed loop.

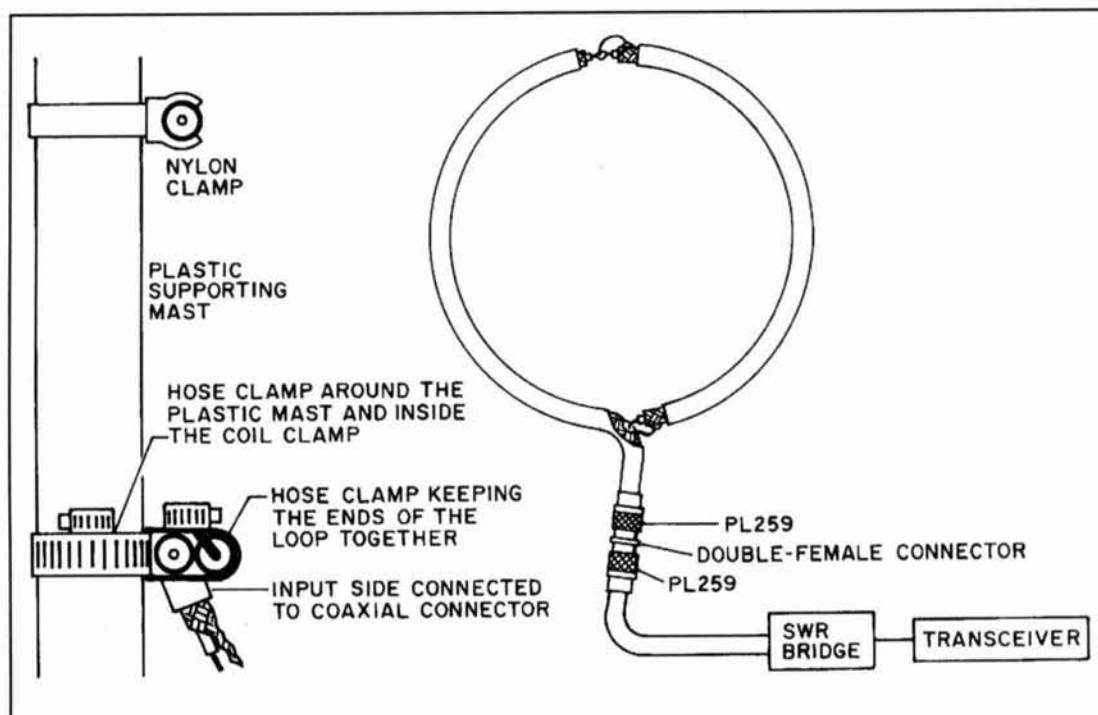


Figure 7. Details of feed loop mount and connections.

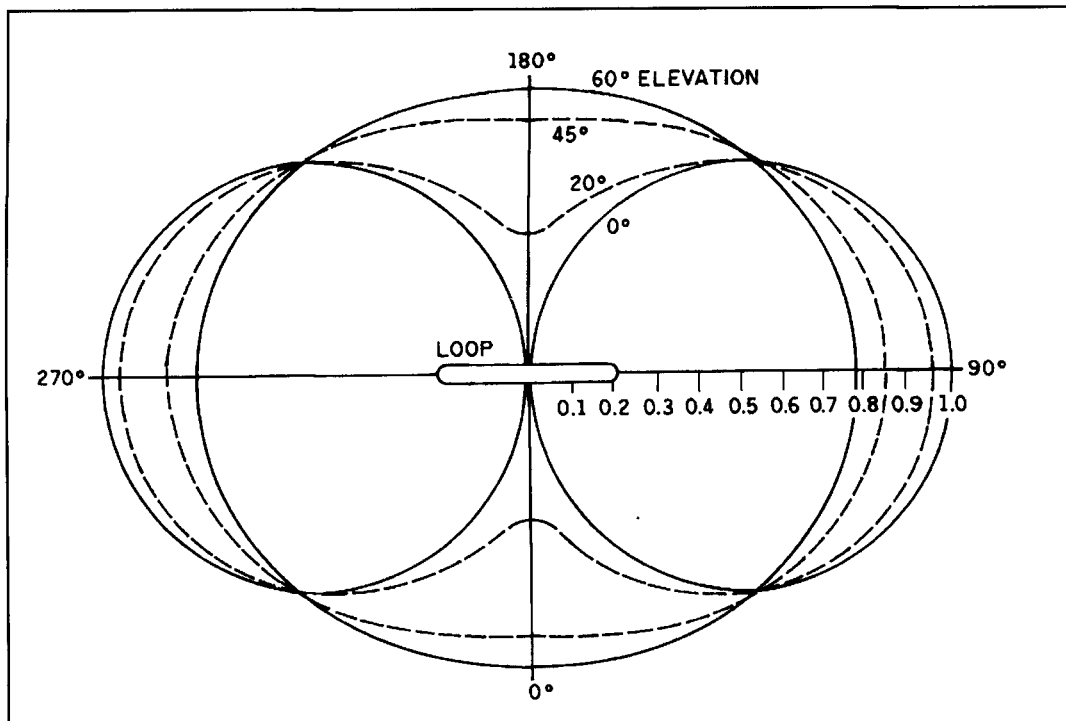


Figure 8. Radiation pattern of the short loop antenna.

It's important that the entire antenna have a solid earth connection to obtain good results.

9. Antenna adjustments must be made when the antenna is mounted in place. The loop must be mounted vertically—away from metallic masses (fences, wires, parked cars, etc.) lying in the plane of the loop.

10. If you experience RF return in the shack, insert some ferrite toroids in the coax line at the bottom of the loop. ■

#### REFERENCE

1. Roberto Craighero, I1ARZ, "Transmitting Short Loop Antennas for the HF Bands: Part 1," *Communications Quarterly*, Summer 1993, page 63.

#### BIBLIOGRAPHY

1. Kenneth H. Patterson, "Down to Earth Army Antenna," *Electronics*, August 1967.
2. Lewis McCoy, W1ICP, "The Army Loop in Ham Communications," *QST*, March 1968.
3. James Taylor, W2OZH, "The Mobiloop," *QST*, November 1968.
4. Joseph Boyer, "Surprising Miniature Low Band Antennas," *73 Amateur Radio*, August/September 1975.
5. Sven Ramstrom, "HF Loop for Transmitting and Receiving," *IERE Conference Proceedings n. 50*, "Radio Receivers and Associated Systems," 1981, pages 455-487.

6. Hans Wuertz, DL2FA, "DX Antennen mit Spiegleden Flacchen—Magnetische Antenne," *CQ-DL*, February/June 1983.
7. Guenther Schwartzbeck, DL1BU, "Rahmen und Ringantennen," *CQ-DL*, May 1985.
8. J.R. Killen, G3KPV, "A Compact HF Antenna for Portable or Base Operation," *Radio Communication*, September 1983.
9. Jerry Hall, K1TD, and Bob Schetgen, KU7G, "Six Winners Emerge from ARRL Competition," *QST*, February 1985.
10. Ted Hart, W5QJR, "Small, High Efficiency Loop Antennas," *QST*, June 1986.
11. Ted Hart, W5QJR, *Small, High Efficiency Antennas—Alias the Loop*, 100-page booklet published by Antenna Radio Products, P.O. Box 334, Melbourne, Florida 32902.
12. Roberto Craighero, I1ARZ, "Le Antenne Loop di Piccola Dimensione per Trasmissione e Ricezione in Bande Decametriche," *Radio Rivista*, January/May 1988.
13. Roberto Craighero, I1ARZ, "Electrically Tunable HF Loop," *Radio Communications*, February 1989.
14. Ted Hart, W5QJR, "The Convoluted Loop," *Ham Radio*, April 1989.
15. Roberto Craighero, I1ARZ, "Loop di Piccole Dimensioni a Grande Efficienza," *Radio Rivista*, December 1989.
16. Roberto Craighero, I1ARZ, "A Magnetic Loop for the Low Bands (40-80-160 meter)," *Radio Communication*, February 1991.
17. A.J. Henk, G4XVG, "Loop Antennas, Facts Not Fiction," *Radio Communication*, September/October 1991.
18. Praheep K. Wahi, "HF Miniloop Antennas," *RF Design*, January 1992.
19. Roberto Craighero, I1ARZ, "More on Short Loop Antennas," and "A Portable Short Loop Antenna for the 20 Meter Band," *Radio Communication*, March/April 1992.
20. Recent editions of *The ARRL Antenna Book*, American Radio Relay League, Newington, Connecticut.

## PRODUCT INFORMATION

### Brian Beezley's NEC/Yagis 2.0 Software

Brian Beezley, K6STI, announces NEC/Yagis 2.0. This software for IBM PC compatibles models Yagis using the Numerical Electromagnetics Code. One feature synthesizes the pattern and gain for an array of Yagis from the pattern of a single antenna. This

feature allows for determination of optimal E- and H-plane spacing for EME arrays of unlimited size. NEC/Yagis includes a facility for plotting polar and rectangular radiation patterns. The amateur version costs \$100. For more information, contact Brian Beezley, K6STI, at 507-1/2 Taylor, Vista CA, 92084, or call 619-945-9824.

## ASTRON POWER SUPPLIES

• HEAVY DUTY • HIGH QUALITY • RUGGED • RELIABLE •

### 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)



MODEL VS-50M

### 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 3/4 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			
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

# LETTERS

## Why not try pseudocode?

Bah humbug! Of *course* you should publish program listings—not for people to *type in*, but for people to *read*, to see how the program works. Publishing a computer project without the listings is like publishing a circuit without the schematic. The reader just has to take it on faith that the black box works as claimed.

But Dave Barton has a point: Listings can be bulky. I suggest you do what computer scientists do, and publish *pseudocode*. That is, publish precise, to-the-point descriptions of algorithms, written in a mixture of programming language and English, as appropriate. That way we don't have to read every REM and PRINT statement.

I enclose an example (see **Listing 1**) based on K4IPV and K0RLT's code in the Summer 1993 issue, page 71. The language is a mix of English and Pascal, with some "begins" and "ends" left out where the indentation makes them unnecessary. Keywords that specify program logic are in boldface, variables are in italics, and English is in Roman type. There is no strict standard for pseudocode; you can use any notation that programmers will understand.

Detailed pseudocode is like the schematic of a circuit; translating it into Pascal, C, or BASIC is like laying out a circuit board. Computer science journals normally publish pseudocode, leaving the actual program implementation to the reader.

**Michael Covington, N4TMI**  
Athens, Georgia

## Spurred on by "the lost art of homebrewing"

During the last years, I ceased to be an active radiotransmission amateur. I found that the principal reasons of this hobby: "The satisfaction of making their own equipment and the possibility of discussing and checking them on the air, with other radiotransmission amateurs," had disappeared because the good running of commercial equipment is no wonder at all, and it doesn't attract anybody's attention after the third communication!

As soon as *Communications Quarterly* appeared, I decided to subscribe since it made note of this thing that I had lost: "The courage of making my own communication equipment."

Algorithm to calculate inductance ( $L$ ) of loop antenna  
given  $A$ =diameter,  $B$ =width of winding,  $N$ =turns:

**case shape of**

*triangle*:  $K1 := .006$ ;  $K2 := 1.1547$ ;  $K3 := .65533$ ;  $K4 := .1348$ ;

*square*:  $K1 := .008$ ;  $K2 := 1.4142$ ;  $K3 := .37942$ ;  $K4 := .3333$ ;

*hexagon*:  $K1 := .012$ ;  $K2 := 2.0000$ ;  $K3 := .65533$ ;  $K4 := .1348$ ;

**end case;**

$NN := N + 1$ ;

$XA := K1 * N^2 * A$ ;

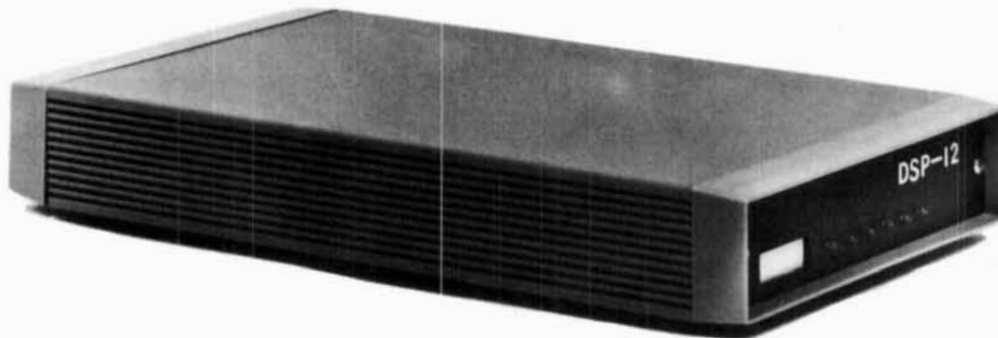
$XB := \log((K2 * A * N) / (NN * B)) + K3 + (K4 * NN * B) / (A * N)$ ;

$L := XA * XB$ ;

**result is  $L$ .**

Listing 1

# 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

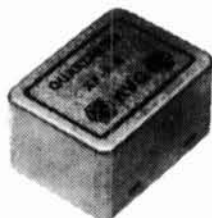
FAX: (609) 751-9705

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.

# K.V.G. CRYSTAL PRODUCTS



## 9 MHz CRYSTAL FILTERS

MODEL	Appli- cation	Band- width	Poles	Price
XF-9A	SSB	2.4 kHz	5	\$ 75.00
XF-9B	SSB	2.4 kHz	8	105.00
XF-9B-01	LSB	2.4 kHz	8	145.00
XF-9B-02	USB	2.4 kHz	8	145.00
XF-9B-10	SSB	2.4 kHz	10	185.00
XF-9C	AM	3.75 kHz	8	110.00
XF-9D	AM	5.0 kHz	8	110.00
XF-9E	FM	12.0 kHz	8	120.00
XF-9M	CW	500 Hz	4	80.00
XF-9NB	CW	500 Hz	8	165.00
XF-9P	CW	250 Hz	8	199.00
XF-910	IF noise	15 kHz	2	20.00

## 10.7 MHz CRYSTAL FILTERS

XF-107A	12 kHz	\$105.00
XF-107B	15 kHz	105.00
XF-107C	30 kHz	105.00
XF-107D	36 kHz	120.00
XF-107E	40 kHz	125.00
XF-107S139	100 kHz	175.00

## 41 MHz CRYSTAL FILTER

XF-410S02 \$199.00

Write for full details of crystals and filters  
Shipping \$8.00

Prices subject to change without notice.



# si

Spectrum International, Inc.  
P.O. Box 1084 Dept. Q  
Concord, MA 01742 U.S.A.  
Phone: 508-263-2145  
FAX: 508-263-7008

## Looking to challenge your building skills?

## Want to explore beyond Amateur communications?

Well here's the publication for YOU!

COMMUNICATIONS QUARTERLY, the journal of communications technology. You are invited to become part of an elite group of Radio Amateurs and technical professionals as a subscriber. Circulation will be strictly limited.

This new quarterly publication has been put together with YOU in mind. Articles will explore, in-depth, areas that no Amateur Magazine is currently covering.

State-of-the-art electronics like direct synthesis, digital signal processing, and computer control are fully investigated and explained so that you can take advantage of these new technologies. COMMUNICATIONS QUARTERLY articles cover high speed data communications, the latest in antennas using computer designed models, plus much more. Articles examine many areas of professional electronics in a way that trade publications can't. Insights into areas that are only whispered about are normal fare. A major effort is made to show how applications can be translated into Amateur Radio operations.

Each quarterly has approximately 100 pages of technical material; advertising is limited. Articles flow from page to page, without the interruption of ads placed in the middle. The open layout accentuates each author's work and lets you make notes, calculations, or comments for later reference. Graphs, photos, computer program listings, and charts are presented in an easy-to-use format. COMMUNICATIONS QUARTERLY is printed on a high-quality stock and "perfect bound" (square backed). It is such a valuable resource, you'll want to file each copy away in your technical archives for future reference. In fact, over time you'll find much of what is presented in COMMUNICATIONS QUARTERLY will become the standard in the Amateur and professional communications field.

## Act now. Subscribe today!

1 year—\$29.95  Foreign—1 year \$39.95  Foreign Air Mail—1 year \$ 60.00   
2 years 56.95  2 years 76.95  2 years 118.95

Name \_\_\_\_\_ Call \_\_\_\_\_

Address \_\_\_\_\_

City \_\_\_\_\_ State \_\_\_\_\_ Zip \_\_\_\_\_

Check  Money Order  MasterCard  Discover

VISA  AMEX

Card No. \_\_\_\_\_ Expires \_\_\_\_\_

Signature \_\_\_\_\_

Mail to: CQ Communications, Inc.  
76 North Broadway, Hicksville, NY 11801  
Order Toll Free 1-800-853-9797/ FAX: 516-681-2926

# 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

## 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

## 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	\$ 89.65
AN779R 20W	\$ 93.20	EB27A 300W	\$136.80
		EB104 600W	\$371.65

### 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 1/4) \$ 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 • Beaver Creek, Ohio 45434-5840  
(513) 426-8600 • FAX (513) 429-3811



24 HOUR SHIPPING

ELESCO • HITACHI • B&K PRODUCTS  
GUARANTEED LOWEST PRICES

TO ORDER CALL TOLL FREE  
1-800-292-7711  
1-800-445-3201 (Con.)

AFFORDABLE - HIGH QUALITY  
2 YEAR WARRANTY



STANDARD SERIES  
S-1325 25MHz \$349  
S-1340 40MHz \$495  
S-1365 60MHz \$649

- High Luminance 6" CRT
- 1mV Sensitivity
- X-Y Operation
- Voltage, Time, Frequency differences displayed on CRT thru the use of cursors (S-1365 only)
- Plus much, much more
- TV Sync
- 2 - x1, x10 Probes
- Complete Schematic

ELESCO OSCILLOSCOPES



DELUXE SERIES  
S-1330 25MHz \$449  
S-1345 40MHz \$575  
S-1360 60MHz \$775

- Delayed Sweep
- Automatic Beam Finder
- Z Axis Modulation
- Built-in Component Test
- Plus all the features of the "affordable" series
- Dual time base
- Illuminated internal grid/cule

HITACHI Compact Series Scopes

V-212 - 20MHz Dual Trace	\$399
V-525 - 50MHz, Cursors	\$995
V-523 - 50MHz, Delayed Sweep	\$949
V-522 - 50MHz, DC Offset	\$895
V-422 - 40MHz, DC Offset	\$795
V-222 - 20MHz, DC Offset	\$649
V-660 - 60MHz, Dual Trace	\$1,149
V-665A - 60MHz, DT, w/cursor	\$1,325
V-1060 - 100MHz, Dual Trace	\$1,395
V-1065A - 100MHz, DT, w/cursor	\$1,649
V-1085 - 100MHz, DT, w/cursor	\$1,995
V-1100A - 100MHz, Quad Trace	\$2,495
V-1150 - 150MHz, Quad Trace	\$2,895

### B&K OSCILLOSCOPES

2120 - 20MHz Dual Trace	\$389
2125 - 20MHz Delayed Sweep	\$539
1541B - 40MHz Dual Trace	\$695
2160 - 60MHz Dual Trace, Delayed Sweep, Dual Time Base	\$949
2190 - 100MHz Three Trace Dual Time Base, Delayed Sweep	\$1,395
2522A - 20MHz / 20MS/s Storage	\$875

Digital Capacitance Meter  
CM-1550B by Elenco  
Big 1" Display \$58.95  
9 Ranges  
1µF-20,000µF  
3% basic accuracy  
Zero control w/Case

Digital LCR Meter  
LC-1801 by Elenco  
\$125  
Measures: Capac, Induct, Res.  
1µF-200µF  
Res. 0.1-20M

Digital Multimeter  
DVM-638 by Elenco  
\$39.95  
11 Functions with Case

FLUKE MULTIMETERS  
(All Models Available Call)  
Model 93 \$1,225.00  
Model 95 \$1,549.00  
Model 97 \$1,795.00  
Model 10 \$62.95  
Model 12 \$79.95

Digital Triple Power Supply  
Elenco XP-765  
\$289  
0-20V @ 2A  
0-20V @ 1A  
5V @ 5A  
Fully regulated, Short circuit protected with 2 limit control, 3 separate supplies XP-660 with Analog Meters \$195

Video Head Tester  
Elenco HT-200  
\$44.95  
Tells you if VHS head is defective or worn

Digital Multimeter w/ Inductance & Capacitance  
LCM-1850 by Elenco  
\$75.00  
Ten Functions

Color Convergence Generator  
Elenco SG-250  
\$89.95  
Kit \$69.95  
Finest in the industry  
10 rock steady patterns  
RF & video output

Quad Power Supply  
XP-580 by Elenco  
\$79.95  
2-20V @ 2A  
12V @ 1A  
5V @ 3A  
-5V @ .5A  
Fully regulated and short circuit protected. Made in USA by Elenco

Triple Power Supply  
XP-620 by Elenco  
Assembled \$75  
Kit \$50  
2 to 15V @ 1A  
-2 to -15V @ 1A  
(or 4 to 30V @ 1A and 5V @ 3A)  
Made in USA by Elenco  
All the desired features for doing experiments. Features short circuit protection, all supplies.

AM/FM Transistor Radio Kit  
Elenco AM/FM 108  
\$27.95  
14 Transistors • 5 Diodes  
Makes a great school project  
AM Radio Kit AM550 \$18.95

True RMS 4 1/2 Digit Multimeter  
M-700T by Elenco  
\$135  
95% DC Accuracy  
1% Resistance with Freq. Counter  
Data Hold

Sweep/Function Generator  
with Freq. Counter  
Elenco Model GF-8026  
\$239  
Sine, Square, Triangle, Pulse, Ramp  
2 to 2MHz, Freq. Counter, 1-10MHz  
Internal Linear & Logic Sweep

Function Generator  
Blox #9600 by Elenco  
\$28.95  
Provides sine, triangle, square wave from 1Hz to 1MHz  
AM or FM capability  
\$26.95

XK-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  
■ +1.25 to 15VDC @ 1 Amp  
■ +1.25 to 20VDC @ 1 Amp  
■ +1.25 to 15VDC @ 1 Amp  
■ +12VDC @ 1 Amp  
■ -12VDC @ 1 Amp  
■ -5VDC @ 1 Amp  
■ 30VAC Center Tapped @ 15VAC @ 1 Amp  
Analog - Section  
■ Function Generator Sine  
■ Frequency adjustable in Hz  
■ 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 1 to 100kHz  
■ Clock amplitude 5Vpp square wave  
Breadboards \$159.95  
Assembled \$129.95 Kit

Learn to Build and Program Computers with this Kit  
Includes: All Parts, Assembly and Lesson Manual  
Model MM-8000  
\$129.00  
by Elenco

Elenco Wide Band Signal Generators  
SG-9000 \$119  
RF Freq 100K-450MHz AM Modulation  
with 100Hz Variable RF output  
SG-9500 w/ Digital Display & 150MHz built-in counter \$239

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.

WE WILL NOT BE UNDERSOLD  
UPS SHIPPING, 48 STATES 5%, OTHERS CALL  
IL RES add 7.75% TAX  
PROBES INCL. ALL SCOPES & METERS

C&S SALES INC.  
1245 ROSEWOOD, DEERFIELD, IL 60015  
FAX: 708-520-0885 • (708) 541-0710

15 DAY MONEY BACK GUARANTEE  
FULL FACTORY WARRANTY  
WRITE FOR FREE CATALOG  
Prices Subject To Change

# NEC/Wires 1.0

Why Fool Around With MININEC  
When You Can Have The Real Thing?

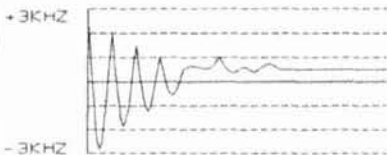
NEC/Wires 1.0 is the first general implementation for amateurs of the powerful and sophisticated Numerical Electromagnetics Code. NEC is more accurate than MININEC, runs faster, and handles much larger models. Only NEC can accurately predict the performance of antennas close to earth by modeling true ground losses. NEC can tell you how many elevated radials you need and how high to put them. Analyze delta loops and wire beams. Find optimal height for NVIS arrays. NEC has no frequency-offset error like MININEC, nor does it require tapered segmentation for complex models with bent wires or multiwire junctions. NEC can take advantage of left/right antenna symmetry over ground and two planes of symmetry in free space to greatly reduce calculation time. NEC can quickly and accurately model huge, interlaced, multiband, cubical quads. It automatically detects and converts tapered, telescoping Yagi elements to uniform-diameter conductors. NEC can model transmission lines in log periodics without using any segments. With 1000 segments available, it's likely you can model every antenna and conductor in your antenna farm to analyze all interactions. (NEC provides 2000 segments for free-space models with left/right symmetry.)

NEC/Wires reads AO/MN antenna files so it's easy to analyze existing models. The program features automatic wire segmentation, symbolic dimensions, symbolic expressions, transcendental functions, voltage or current sources, RLC, Laplace transform, and impedance loads, transmission lines, and skin-effect modeling. NEC/Wires plots azimuth and elevation patterns in polar or rectangular coordinates with overlays.

NEC/Wires 1.0 is \$100 for amateur use only. Visa, MasterCard, U.S. check, cash, or money order. Add \$5 overseas. NEC/Wires requires a 386+ 387 or 486DX, a hard disk, and VGA.

Brian Beezley, K6STI  
507 1/2 Taylor, Vista, CA 92084 • (619) 945-9824

## Transmitter Finger Printing is here!



Every radio transmitter has a unique frequency versus time characteristic. When keyed, a "finger-print" unique to each transmitter is produced that can be captured, stored and analyzed. Using patented technology--available from **MoTron Electronics**--and our advanced software, you can identify the abusers on your repeater!

The MoTron TxID includes a sophisticated circuit board that plugs into your IBM compatible computer and our exclusive software.

Additional features of the MoTron TxID include:

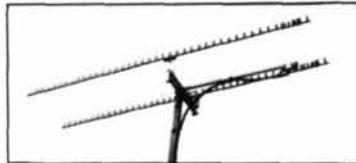
- ✓ Measures deviation
- ✓ Measures frequency of the received signal relative to your receiver's center frequency
- ✓ Has CI-V output for Icom receiver control
- ✓ Tape recorder output/control with digitally delayed audio
- ✓ Decodes DTMF

Call or write for a brochure with full details, technical specifications, and prices.

## MoTron Electronics

310 Garfield St., Suite 4 • Eugene OR 97402  
Info: (503) 687-2118 • Fax: (503) 687-2492

## 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 Feed		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	high gain 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

	900, 1269, 1296, 2304, 2400, 3456, 5760 MHz		
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

## The Latest Addition to CQ's Video Library!

### Getting Started in Contesting



Yours for only \$19.95!

Whether you're a newcomer to contesting or an experienced veteran, CQ's **Getting Started in Contesting** is for you! You'll get advice and operating tips from some of contesting's most successful competitors, including Ken Wolff, K1EA, Dick Newell, AK1A, and CQ's own contest columnist, John Dorr, K1AR.

Here's just a sample of what you'll see:

- What contesting's all about
- Explaining contest jargon
- Tips for beginners
- Competing from a small station
- Operating secrets from the "pros"
- Live QSOs from world class stations
- VHF/UHF contesting

Send for your contest video tape today and watch your scores climb!

Call our order department today at  
**(800) 853-9797**

or write to: CQ Communications  
76 North Broadway, Hicksville, NY 11801.



## 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





## LUKE POWER SUPPLIES

**CONTINUOUS DUTY  
AMPERE RATINGS**

**24 Hours A Day  
7 Days A Week  
52 Weeks A Year**



S40-40AMP-13.8V.....	\$275
S55-55AMP-13.8V.....	\$310
S65-65AMP-13.8V.....	\$425
S80-80AMP-13.8V.....	\$540
S100-100AMP-13.8V.....	\$585
S35H-35AMP-28V.....	\$445
S55H-55AMP-28V.....	\$540
S25VH-25AMP-50V.....	per quote
S55VH-55AMP-50V.....	per quote
OPTIONAL RACK MOUNT.....	\$75
OPTIONAL LCD METER.....	\$75

- |  |   |
|--|---|
| <ul style="list-style-type: none"> <li>• Electronic Regulated</li> <li>• Fold Back Current Limit</li> <li>• Crowbar Protection</li> <li>• Over Temp Protection</li> <li>• Over Temp Indicator</li> <li>• Input Surge Protection</li> <li>• Digital LCD Volt/Amp Meter w/display hold (optional)</li> </ul> | <ul style="list-style-type: none"> <li>• Soft start on most models</li> <li>• Made in U.S.A.</li> <li>• One Year Warranty</li> <li>• Rack Mount Option</li> <li>• Crowbar Indicator</li> <li>• 120/240v all models</li> <li>• Ripple Low as 2mv</li> <li>• Industrial transformer manufactured in U.S.A.</li> </ul> |
|--|---|

**LUKE CO.**

7113 North 9 Mile, Lake City, MI 49651  
(616) 229-4593

VISA

M.C.

## FANTASTIC FM TRANSCEIVER KITS!

**2 meters • 223 MHz  
• 440 MHz • 6 meters  
Only \$149.95!  
\$169.95 for 440 MHz  
Fully Synthesized  
No Crystals**



Ramsey breaks the price barrier on FM rigs! The FX is ideal for shack, portable, mobile or a dedicated packet rig. Wide frequency coverage (20MHz) and programmable repeater splits makes the FX the perfect rig for CAP, MARS or business band use. Packeteers really appreciate the built-in packet I/O connector, true FM signal, instant T/R switching and 9600 baud capability. You get twelve channels easily diode programmed anywhere you want, 5 watts RF output (3-4 for FX-440), sensitive dual conversion receiver with crystal and 6 pole ceramic IF filters, plus proven easy assembly.

You build your FX in stages of a few parts at a time, logically working your way along through each section, testing your work as you build. The only test equipment you'll need is a voltmeter and another radio or frequency counter. Our 120 page instruction manual will carefully guide you to a finished unit that will work first time — and you'll learn, too! Why pay more for a used foreign-made rig when you can have an AMERICAN MADE one (by you) for less? Kit comes complete with all parts, quality epoxy PC board with printed parts layout, and extensive manual. You only need to add an ICOM/Radio Shack/Yaesu style mike and suitable enclosure. The Ramsey matching case set includes all hardware, knobs and rugged steel case (weights 3 lbs!) with durable black powder coat finish.

FX-50 (6 meters)	\$149.95	FX-223 (1 1/4 meters)	\$149.95
FX-146 (2 meters)	\$149.95	FX-440 (70 cm)	\$169.95
		CFX matching case set	\$24.95

FX manual, specify band desired - \$10.00 refundable with purchase of transceiver kit.

**YOUR SATISFACTION IS GUARANTEED.** Examine your Ramsey product for 10 days. If you're not pleased, return in original condition for refund. Add \$3.95 for shipping handling and insurance. Foreign orders add 20% for surface mail. COD (US only) add \$5.00. Orders under \$20 add \$3.00. NY residents add sales tax. Warranty: 90 days on kit parts; 1 year parts and labor on pre-wired units.

LOCAL INFO 716-924-4560 FAX 716-924-4555

**ORDERS ONLY 1-800-446-2295**

**RAMSEY ELECTRONICS, INC.**

793 Canning Parkway • Victor, NY • 14565

CALL

# NEMAL

FOR

# RF



- Connectors
- Adapters
- Cable Assemblies
- Coaxial Cable

**Manufacturer Of Custom  
Electronic Wire And Cable.**

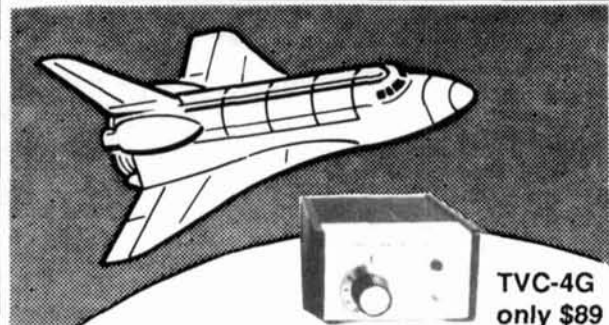
- Low Minimums • Quick Delivery

CALL US AT  
1-800-522-2253  
OR FAX YOUR REQUIREMENTS TO  
1-305-895-8178

Call for your copy of our new 44-page  
Cable & Connector Selection Guide.  
More than 2,500 commercial and QPL  
cable and connector products in stock.

**NEMAL ELECTRONICS, INC.**  
12240 N.E. 14th AVE., N. MIAMI, FL 33161  
(305) 899-0900

## 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)

# AMIDON Associates, INC. W2FMI BALUNS AND UNUNS TRANSFORMERS

How to put MORE power into your antenna and get better and higher signal strength ???

Use the all new **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	\$34.95
2.1-HU50	50:25-OHMS	\$34.95
2.1-HDU50	50:22.22-OHMS	\$49.95
	50:25-OHMS	
2.25-1-HU112.5	112.5:50-OHMS	\$34.95
2.1-HU100	100:50-OHMS	\$34.95
2.1-HDU100	112.5:50-OHMS	\$49.95
	100:50-OHMS	
1.78-1-HU50	50:28-OHMS	\$34.95
1.5-1-HU50	50:32-OHMS	\$34.95
1.5-1-HU75	75:50-OHMS	\$34.95
4.1-HRU50	50:12.5-OHMS	\$49.95
4.1-HCU50	50:12.5-OHMS	\$49.95
9.1-HRU50	50:5.56-OHMS	\$49.95
9.1-HU150	50:5.56-OHMS	\$34.95
1.78-1-HDU50	50:28-OHMS	\$49.95
	50:12.5-OHMS	
1.56-1-HDU50	50:32-OHMS	\$49.95
	50:18-OHMS	
HMMU50(5 diff ratios)	MULTIMATCH UNUN	\$69.95
1.78-1-HMMU50 (8 different ratios)	MULTIMATCH UNUN	\$69.95

W2FMI HIGH POWER BALUNS	IMPEDANCE MATCH	COMPLETE UNIT PRICE
4.1-HB50	50:12.5-OHMS	\$39.95
1.1-HB150	50:50-OHMS	\$39.95
1.5-1-HB75	75:50-OHMS	\$59.95
2.1-HB100	100:50-OHMS	\$59.95
4.1-HBM200	200:50-OHMS	\$39.95
4.1-HB/U200	200:50-OHMS	\$59.95
4.1-HBHT200	200:50-OHMS	\$59.95
6.1-HB300	300:50-OHMS	\$59.95
6.1-HB/U300	300:50-OHMS	\$119.95
9.1-HB450	450:50-OHMS	\$119.95
12.1-HB600	600:50-OHMS	\$199.95

NOTE: OTHER BALUN & UNUN TRANSFORMERS AVAILABLE. PLEASE CALL OR WRITE FOR ADDITIONAL DETAILS.  
BOOKS: "1" Transceiver Line Transformers Design Handbook", by Jerry Sevick, W2FMI, Amidon Associates, Inc. 1991. \$9.00 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 Gadwick Street, Dominguez Hills, CA 90220  
TEL: (310) 763-5770 • FAX: (310) 763-2250



Unconditional money back guarantee for 1 year on completed unit.

**SPECIAL KIT PRICE \$28.00!**  
**CALL TODAY!**

# 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. \$100. AO-Pro 6.0 (5700 pulses), \$600. MNC+MNH 4.5 (assembly language, 480 pulses, no optimizer or 3-D patterns), \$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 reflection, 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. \$75. YOC 5.0 (assembly language, much faster), \$100.

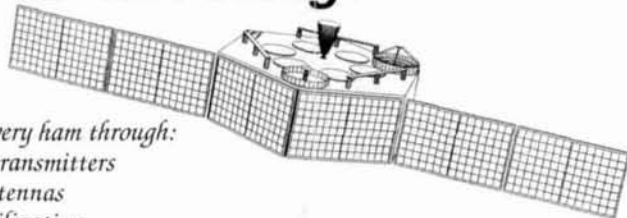
**NEC/Wires 1.0** models true ground losses and complex arrays. Optimize above-ground radials, delta loops, etc. 1000 segments. \$100.

**NEC/Yagis 2.0** provides highest-accuracy Yagi analysis. Quick pattern synthesis for EME arrays of unlimited size. 2000 segments. \$100.

AO and NEC require a 386+387 or 486DX and VGA. Others come with both coprocessor and noncoprocessor versions for use on any PC. Extensive documentation. Visa, MasterCard, U.S. check, cash, or money order. Add \$5 overseas.

Brian Beezley, K6STI  
507 1/2 Taylor, Vista, CA 92084 • (619) 945-9824

## 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

## ADVERTISER'S INDEX

AEA/Adv. Elec. Applications.....	5
Alinco Electronics .....	Cov.II
Amidon Associates .....	112
AMSAT.....	112
Astron Corporation .....	105
Beezley, Brian, K6STI.....	110,112
C & S Sales .....	109
Communications Quarterly.....	108
Communications Concepts .....	109
CQ Books & Videos .....	8
CQ Videos.....	110
CQ T-Shirts.....	9
Down East Microwave.....	110
HAL Communications .....	10
Hewlett Packard .....	7
Japan Radio, JRC.....	12
Kenwood .....	Cov.IV, 2
L.L. Grace Comm. Products .....	107
Lewallen, Roy, W7EL .....	109
Luke Co.....	111
Motorola.....	11
Motron Electronics.....	110
Nemal Electronics .....	111
OPTOElectronics.....	1
Palomar Engineers .....	109
PC Electronics.....	111
Ramsey Electronics.....	111
Spectrum International, Inc.....	108
TX RX Systems Inc .....	110
Yaesu.....	Cov.III

We'd like to see your company listed here too. Contact Arnie Sposato, N2IQO, at (516)681-2922 or FAX at (516)681-2926 to work out an advertising program to suit your needs.

NEW

"What a great field radio. Mobile, too! I couldn't afford an HF rig until now."

"What a great price! Terrific features, high performance – and within my budget."

"Yaesu did it again!"



## FT-840 Compact HF Transceiver

- Direct Digital Synthesis (DDS)
- Frequency coverage:  
RX: 100 kHz-30 MHz  
TX: 160-10 m
- IF Shift
- 100 Memory Channels (Independent TX/RX per memory)
- Twin Band Stacking VFOs
- FM\* Repeater Operation Automatic 10-Meter Repeater Offset w/Selectable CTCSS Encode
- CW Reverse Feature
- Choice of Two Optional Antenna Tuners:  
FC-10 Matching External Antenna Tuner  
FC-800 External Remote Antenna Tuner
- Accessories:  
Contact your Dealer for full details.
- \* Optional

# It's a small price to pay for such a wealth of features.



If you're trading up from an older rig, but have a budget, you want the most you can afford in top-notch HF. Then the FT-840 is for you. It's right on the money! Considering a mobile HF or field radio and doubt the quality and features of tiny HF rigs? Then the FT-840 is for you. It won't disappoint you!

Built to handle rigorous field operation, the new intense LCD display affords sharp visibility in bright sun-

light. Die-cast heat sink and internal thermally switched fan keep the FT-840 running cool. Modular design circuit boards ensure operating efficiency – manufacturing excellence you'd expect in much higher priced radios.

For high performance, the FT-840 features a low noise front end that uses the latest in FET RF amplifier design. Two DDSs and magnetic encoder for silent, smooth tuning and fast switching. Twin band-stacking VFOs. And,

automatic 10-m FM (optional) repeater offset with selectable CTCSS. Even two optional external antenna tuners to customize your rig.

Top of the line quality and features at a remarkably low price. Just what you'd expect from Yaesu! For high-tech performance, and a wealth of features that won't break your budget ask your dealer about the FT-840.

# YAESU

*Performance without compromise.™*

# 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