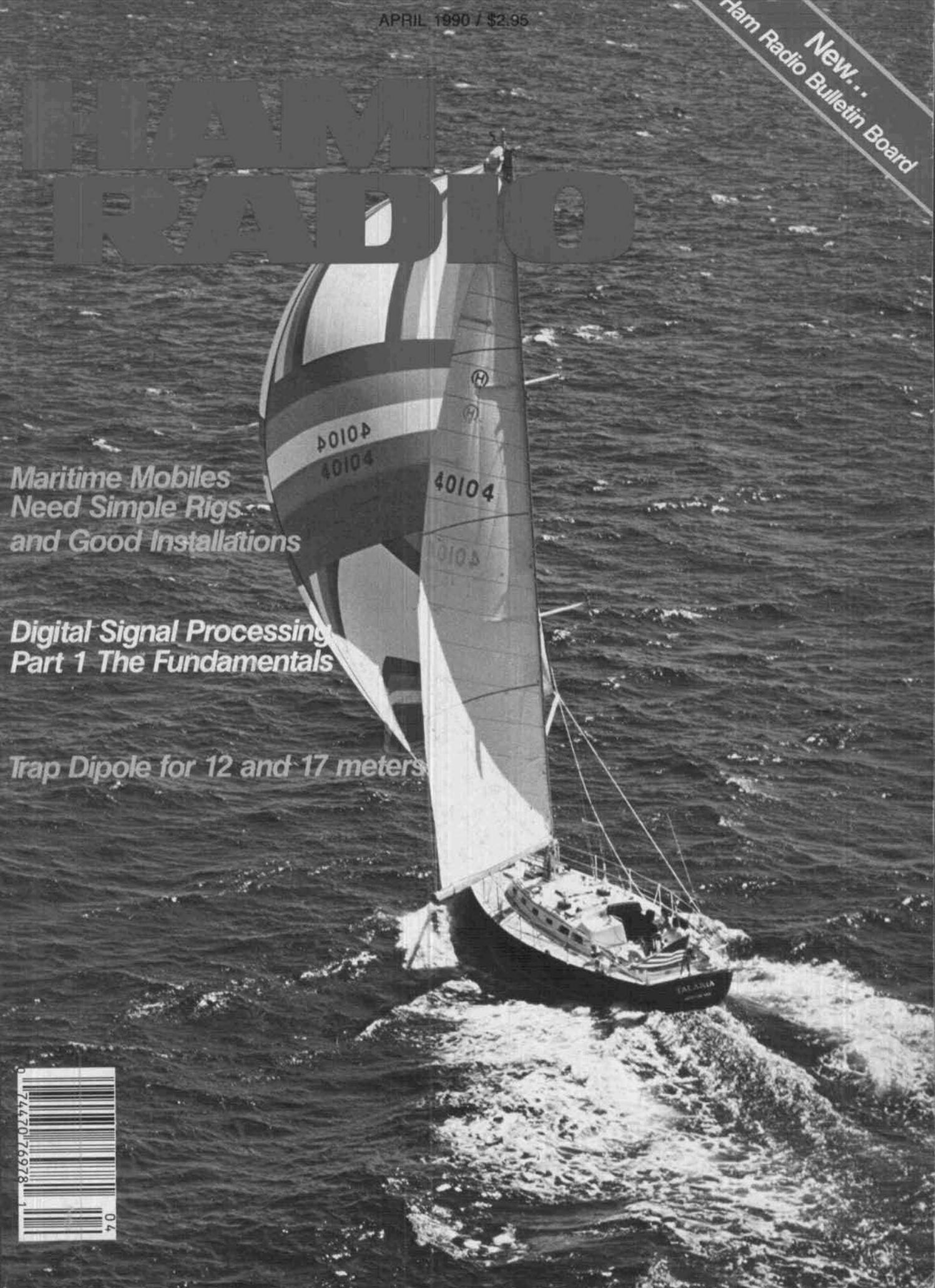
APRIL 1990 / $2.95

New...
Ham Radio Bulletin Board

# HAM RADIO




## Maritime Mobiles Need Simple Rigs and Good Installations

## Digital Signal Processing Part 1 The Fundamentals

## Trap Dipole for 12 and 17 meters

ICOM

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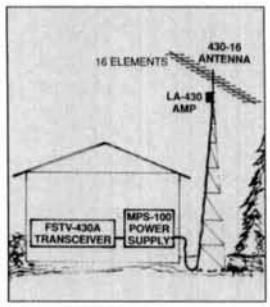
**First in Communications**

# INTRODUCING AEA'S NEW ATV SYSTEM

Add a new dimension to your amateur radio communications with AEA's Amateur Television (ATV) system. If you hold at least a technician-class license, you can transmit and receive live or taped audio and video Fast-Scan TV (FSTV) information that rivals broadcast quality. Now you can share more than conversation over the air with this new mode of "personal communications."



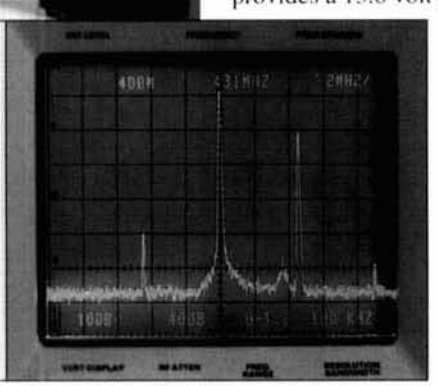

**It's Easy and Inexpensive.** If you have a video camera or camcorder and a standard TV set, you may already own the most expensive components of an ATV system. AEA's ATV system includes a transceiver and antenna. Simply connect the camera, TV and the antenna to the transceiver, and you're on the air LIVE with one watt P.E.P.! Your TV set will monitor your transmitted and received pictures. If you want to broadcast with more power, AEA also offers a 50 watt mast-mounted linear amplifier with power supply.

**The FSTV-430A Transceiver** features a low-noise UHF GaAsFET preamp with a typical noise figure of less than 1.5dB and a crystal-controlled or variable tuning down converter. Output is available on channel 3 or 4 for signal reception AND monitoring transmissions. Two frequencies can be selected from the front panel for transmission (one crystal is included). The AEA design is also optimized for superior video and audio quality without sync buzz even with weak signals. The FSTV-430A is the only transceiver you need to work ATV and it also allows you to use the same TV set to monitor your transmitted and received pictures.

**The LA-430/50 Amplifier with Power Supply** gives a boost to your ATV signal. It includes a 50W P.E.P. mast-mounted Linear Amplifier (**patent pending**) covering 420 to 450 MHz and a GaAsFET preamp which utilize the antenna feedline for DC power. The mast- mount eliminates the line loss between the amplifier/preamplifier and the antenna to improve both transmission and reception, and is the equivalent of a 100W amplifier in the shack with a 3dB line loss. The amplifier is housed in a weather-resistant alodized aluminum case. The MPS-100 power supply also provides a 13.6 volt output for the FSTV-430A.

**What is the advantage of Vestigial Sideband (VSB)?**

AEA's FSTV-430A Vestigial Sideband operation drastically reduces adjacent-channel interference. VSB requires much less bandwidth than existing double-sideband designs; it's the standard method of modulation required by the FCC for all U.S. broadcast TV stations. Similar in principle to SSB, VSB puts all of the audio energy and most of the video in ONE sideband instead of two. Using about half the spectrum space of competitive units, the FSTV-430A is the ONLY ATV unit that conserves spectrum space by using VSB. Even with AEA's LA-430/50 amplifier, one sideband is reduced more than 30dB. VSB presents an obvious advantage to the bandwidth-conscious ATV operator.



**The 430-16 Antenna** is a high-performance, computer-optimized yagi specifically designed for ATV operation. It features broadband frequency coverage from 420 to 440 MHz, 14.3dB gain, O-ring sealed connectors, 28 degree E plane and 32 degree H plane beam widths and 16 elements on a 10-foot boom.

**See AEA's FSTV System** at your local authorized AEA dealer. Put yourself in the ATV picture and join the fun!


**Advanced Electronic Applications, Inc.**
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**AEA Brings You A Better Experience.**

# KENWOOD

# The DXpeditioner!

## TS-440S

**Compact high performance HF transceiver with general coverage receiver**

**Portable reliable performance and ease of use makes the TS-440S your obvious "low bands" choice. It is "Every Ham's" rig to go – ham shack, portable or mobile. But don't let the small size fool you – there's lots of "big rig" performance packed into this package. Built-in antenna tuner option. Continuous duty transmitter. Super DynaMix™ front end. Five filter functions. The TS-440S is at your service wherever you wish to operate.**

- **Covers all Amateur bands** General coverage receiver tunes from 100 kHz–30 MHz. Easily modified for HF MARS operation.
- **Direct keyboard entry of frequency**
- **All modes built-in** USB, LSB, CW, AM, FM, and AFSK. Mode selection is verified in Morse Code.
- **VS-1 voice synthesizer (optional)**
- **Built-in automatic antenna tuner (optional).** Covers 80–10 meters.
- **5 IF filter functions**
- **Superior receiver dynamic range** Kenwood DynaMix™ high sensitivity direct mixing system ensures true 102 dB receiver dynamic range. (500 Hz bandwidth on 20 m.)
- **100% duty cycle transmitter** Super efficient cooling permits continuous key-down for periods exceeding one hour. RF input power is rated at 200 W PEP on SSB. 200 W DC on CW, AFSK, FM, and 110 W DC AM. (The PS-50 power supply is needed for continuous duty.)
- **Computer interface port**
- **Adjustable dial torque**
- **100 memory channels** Frequency and mode may be stored in 10 groups of 10 channels each. Split frequencies may be stored in 10 channels for repeater operation.
- **TU-8 CTCSS unit (optional)**

- **MC-43S UP/DOWN mic. included**
- **Superb interference reduction** IF shift, tuneable notch filter, noise blanker, all-mode squelch, RF attenuator, RIT/XIT, and opt. filters fight QRM.
- **Dual SSB IF filtering** A built-in SSB filter is standard. When an optional SSB filter (YK-88S or YK-88SN) is installed, **dual** filtering is provided.
- **VOX, full or semi break-in CW**
- **AMTOR compatible**

**Optional accessories:**

• **AT-440** internal auto. antenna tuner (80 m – 10 m) • **AT-250** external auto. tuner (160 m – 10 m) • **AT-130** compact mobile antenna tuner (160 m – 10 m) • **IF-232C/IC-10** level translator and modem IC kit • **PS-50** heavy duty power supply • **PS-430** DC power supply • **SP-430** external speaker • **MB-430** mobile mounting bracket • **YK-88C/88CN** 500 HZ/270 Hz CW filters • **YK-88S-88SN** 2.4 kHz/1.8 kHz SSB filters • **MC-60A/80/85** desk microphones • **MC-55** (8P) mobile microphone • **HS-4/5/6/7** headphones • **SP-41/50B** mobile speakers • **MA-5/VP-1 HF** 5 band mobile helical antenna and bumper mount • **TL-922A** 2 kw PEP linear amplifier • **SM-220** station monitor (no pan display) • **VS-1** voice synthesizer • **TU-8** CTCSS tone unit • **PG-2C** extra DC cable.


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# Backscatter

## Life, Liberty, and the Right to Privacy

Have you ever wondered about a QSO in progress? Have you ever thought to yourself in the middle of a transmission, "Is this subject really appropriate for Amateur Radio?" Obviously, business communications are forbidden. But common sense also dictates that there are plenty of subjects best left to other methods of communication — like the telephone. Or are they?

An interesting case has come to light in Iowa concerning police monitoring of cordless telephone conversations. The November 29, 1989 edition of *The Wall Street Journal* reported that the Scott County Iowa Sheriff monitored a family's cordless telephone conversations over a 9-month period. This monitoring led to a 10-year sentence for conspiracy and theft. The family, in turn, has sued the sheriff for violating their constitutional right to privacy. In fact, they asked the Supreme Court to determine whether or not Fourth Amendment protection against unreasonable searches and seizures applied. Though the Supreme Court has decided not to hear the case, this story isn't over yet.

According to the *Journal,* lower court precedents don't favor Fourth Amendment protection for cordless telephones. They do apply, however, for more *technically sophisticated* cellular and wireless telephone conversations. Because the technology used for cordless telephones makes the calls easy to intercept, users shouldn't expect protection. The cellular telephone industry was able to ram legislation through Congress that could bring an all expenses paid "vacation" at the Allenwood, Pennsylvania Federal Correctional Facility for anyone hearing even the mere whisper of a cellular telephone conversation. To me this conjures up images of police raids on families huddled around TV sets scanning channels 70 to 83 in search of "neat" cellular conversations. While I shudder at the thought of people with paper pads and tape recorders ready to copy juicy tidbits of information, I hate to think that "Big Brother" might be looking over their shoulders while they do so.

This legislation is an example of a law which criminalizes an act that can't, in any way, be enforced. As a young ham, I feared the FCC. I was afraid that any indiscretion, even my beginner's sloppy fist, would bring the "radio police" to my door with a warrant for my arrest. In all practicality it wasn't going to happen, but I was still scared.

So what's the real scoop? Where should the line be drawn regarding what is and what isn't constitutionally protected in Amateur Radio conversations? For us radio Amateurs, the fight for privacy in communications will have little impact. However, it's always good to be reminded that when we talk, many are listening. A casual comment about going away for a weekend could result in a house break-in, or worse. There are millions of radios, scanners, and other types of equipment out there that can tune into Amateur communications.

It will be interesting to see if this matter is brought before the Supreme Court again. Hypothetically, if the case were to receive a favorable judgment, one could conjecture that all radio communications would enjoy protection. That would mean that you could call, but nobody could listen. Just think of it; the ham bands would be filled with endless CQs, and not one answer! Is this scenario off the wall? Yes. Is it probable? No. Well...maybe?

**de NX1G**

# KENWOOD

# Stacked in your favor!

## TM-231A/ 331A/431A/531A

### FM Mobile Transceiver

**Looking for a compact transceiver for your mobile VHF and UHF operations? KENWOOD has a compact rig for each of the most popular VHF/ UHF bands.**

- **20 multi-function memory channels.**
- **High performance — high power! 50W (TM-231A), 35W (TM-431A) with a 3 position power switch.**
- **Optional full-function remote controller (RC-20).**
  A full-function remote controller can be mounted in any convenient location. Using the IF-20 interface the RC-20 may be connected to four mobile transceivers. (TM-231A/ 431A/531A or the TM-701A).
- **Multi-function microphone supplied.**
  Various controls are provided on the mic. for increased utility.
- **Auto repeater offset on 144 and 220 MHz.**
- **Built-in digital VFO allows selection of the frequency step.**
  (5, 10, 15, 20, 12.5, 25kHz; TM-531A: 10, 20, 12.5, 25kHz.)
- **Selectable CTCSS tone built-in.**
- **Tone alert system — for true "quiet monitoring"!**
  When enabled this function will activate a tone when squelch opens.
- **DRS (Digital recording system).**
  The optional DRU-1 can store received and transmitted messages for up to 32 seconds, allowing the operator to check or return any call using the tone alert system.
- **Automatic lock tuning function (TM-531A).**
- **Repeater reverse switch.**

TM-231A
136–174 MHz receive,
TX on Amateur bands only.
Modifiable for MARS/CAP.
Permit required.

**Optional Accessories:**
• **RC-20** Full-function remote controller • **RC-10** Handset • **IF-20** Interface unit handset • **DRU-1** Digital recording unit • **MC-44** Multi-function hand mic. • **MC-44DM** Multi-function hand mic. with auto-patch • **MC-48B** 16-key DTMF hand mic. • **MC-55** 8-pin mobile mic. • **MC-60A/80/85** Desktop mics. • **MA-700** Dual band (2m/ 70cm) mobile antenna (mount not supplied) • **SP-41** Compact mobile speaker • **SP-50B** Mobile speaker • **PS-430** Power supply • **MB-201** Mobile mount • **PG-2N** Power cable • **PG-3B** DC line noise filter
• **PG-4H** Interface connecting cable
• **PG-4J** Extension cable kit
• **TSU-6** CTCSS unit

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# Comments

## Lose some..win some!

**Dear HR**

I wanted to write to you and express my pleasure that you have picked up Bob Atkins, KA1GT, to write a new microwaves column. I was very sorry to see W1JR drop out, but I'm sure Bob's column will be every bit as interesting. This was the first column I would turn to in *QST* to read. So their loss is your gain!

Keep up the good work at publishing the best ham magazine available.

I might also suggest that possibly an ATV column would be nice. Perhaps W6ORG, Tom O'Hara, or KB9FO, Henry Ruh, would consider being contributing editors.

Again I'm very happy to see a microwaves column.

**Dave Williams, WBØZJP**
**St. Louis, Missouri**

## Early radar buff

**Dear HR**

"Understanding Over-the-Horizon Radar," *Ham Radio*, February 1990, brought back memories of field testing early warning radar SCR270 in the Signal Corps the summer of 1941 on Sandy Hook, New Jersey. It operated around 106 MHz and if the planes and thunderstorms were high enough it would reach out 100 miles or so. Not exactly over-the-horizon but a good radar application at the time. I understand from Dick Aspinwall, W7PV, that there is a museum where this equipment is on display along with the SCR584, an "S" band microwave radar Gunn director, that I later worked on.

Keep up the good work.

**Wayne W. Cooper AG4R**
**Miami Shores, Florida**

## A good time to upgrade

**Dear HR**

I feel compelled to respond to N6SWA's "pet peeve" expressed in the February "Comments," regarding higher class licensees operating in the Novice/Technician 10-meter phone band.

I'm sorry, but I can't sympathize with him. Since the expansion of the 10-meter phone band, the majority of the "rarer" DX stations have been operating below 28.5 MHz. With the excellent propagation we are experiencing at the peak of the sunspot cycle, Novice/Techs have an unprecedented opportunity to take part in some of the best worldwide communications conditions we have had in many years. In six months it should be possible to work DXCC with 100 watts and a dipole with a little effort.

However, these conditions are not going to last, and will decline over the next couple of years. As an avid DXer, I am certainly not about to miss taking part in the peak activity and will operate wherever the DX is, with 100 watts if that will do, or a kilowatt if necessary.

Even under very good propagation conditions 28.3 to 28.5 has plenty of elbow room. And 200 kHz is more than most of us can operate on the other HF phone bands (except 15 meters). I have never heard QRM on 10 that comes close to what you'll find on 20 meters.

The best solution for the Novices and Techs that feel inhibited with their band allocations is, of course, to upgrade. I empathize with the Techs that have a problem with code (note that I don't have a 2X1 call) but with effort most people can make 13 wpm. Thanks to the question pool being "Bashed" a few years ago, the General Class (and higher) theory is now easily passed by anyone who is willing to read and memorize for a dozen or so evenings, with no real need for comprehension.

As further incentive, newcomers to the HF bands should be aware that it won't be very long until 10 meters will only be good for talking across town, and a sporadic E opening to another state will be a big deal. So "kwitcherbellyakin," upgrade, and be ready to put up with some real QRM when 20 is the only band open and everyone is crowded into 200 kHz.

**Bruce Sanborn, KB2WN**
**Rochester, New York**

## On the mark

**Dear HR**

I wish to comment on "Elmer's Notebook" in the January isue of *Ham Radio*. It is by far the best review on "resistance" I have ever read and has a very positive outline on the basics of the subject.

Thanks again for this and many other articles you have written. I look forward to "Elmer's Notebook" each month.

**Stuart J. Tuma, W1QXS,**
**Melrose, Massachusetts**

**Dear HR**

I receive *Ham Radio* magazine every month, and look forward to each magazine with anticipation. I am a project builder and enjoy all the articles.

My wife gives me the magazine every year as a Christmas present. You and your staff are to be congratulated on a very fine magazine. Keep up the fine work.

**Phil Fuglsaug, WDØHXY**
**Imperial, Missouri**

# KENWOOD

## TM-731A/631A
### 144/450 and 144/220 MHz FM Dual Banders

- **Extended receiver range** (136.000 – 173.995 MHz) on 2 m; 70 cm coverage is 438.000 – 449.995 MHz; 1-1/4 m coverage is 215 – 229.995 MHz. (Specifications guaranteed on Amateur bands only. Two meter transmit range is 144 – 148 MHz. Modifiable for MARS/CAP. Permits required.)
- **Separate frequency display for "main" and "sub-band."**
- **Versatile scanning functions.** Dual scan, and carrier and time operated scan stop.
- **30 memory channels.** Stores everything you need to make operating easier. Two channels for "odd splits."
- **50 Watts on 2 m, 35 watts on 70 cm, 25 watts on 1-1/4 m.** Approx. 5 watts low power.
- **Automatic offset selection.**
- **Dual antenna ports.**
- **Automatic Band Change (A.B.C.)** Automatically changes between main and sub-band when a signal is present.
- **Dual watch function allows VHF and UHF receive simultaneously.**
- **CTCSS encode/decode selectable from front panel or UP/DWN keys on microphone.** (Encode built-in, optional TSU-6 needed for decode.)
- **Balance control and separate squelch controls for each band.**
- **Full duplex operation.**
- **Dimmer switch.**
- **16 key DTMF/control mic. included.**
- **Frequency (dial) lock.**

**Optional Accessories:**
• **PG-4H** Extra interface cable for IF-20 (for three to four radios) • **PG-4J** Extension cable kit for IF-20 DC and audio • **PS-430** Power supply • **TSU-6** CTCSS decode unit • **SWT-1** 2 m antenna tuner • **SWT-2** 70 cm antenna tuner • **SP-41** Compact mobile speaker • **SP-50B** Deluxe mobile speaker • **PG-2N** DC cable • **PG-3B** DC line noise filter • **MC-60A, MC-80, MC-85** Base station mics. • **MA-700** Dual band 2 m/70 cm mobile antenna (mount not supplied) • **MB-11** Mobile bracket • **MC-43S** UP/DWN hand mic. • **MC-48B** 16-key DTMF hand mic.

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# MFJ gives you *all 9* digital modes

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**No** 3 year old technology at MFJ!

**Using** the latest advances, MFJ brings you 9 exciting digital modes and *keeps on* bringing you state-of-the-art advances.

**You** get tons of features other multi-modes just don't have.

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**Count** 'em -- you get 9 fun modes -- Packet, AMTOR, RTTY, ASCII, CW, FAX, SSTV, Navtex and full featured Contest Memory Keyer.

**You** can't get all 9 modes in *any* other multi-mode at *any* price. Nobody gives you modes the MFJ-1278 doesn't have.

### The best modem you can get

**Extensive** tests in ***Packet Radio Magazine*** prove the modem used in the MFJ-1278 copies HF packet more accurately than all other modems tested.

**MFJ-1278** is the *only* multi-mode with a ***true*** DCD circuit for HF. This dramatically reduces sensitivity to noise and dramatically increases completed QSOs.

### *Exclusive* Built in Printer Port

**Only** the MFJ-1278 has a dedicated printer port that lets you plug in your printer.

**You** don't need to buy an optional $40 cable just to plug in your printer.

### New Easy Mail™ Personal Mailbox

**You** get MFJ's new Easy Mail™ Personal Mailbox with soft-partitioned memory so you and your ham buddies can leave messages 24 hours a day.

### 20 LED Precision Tuning Indicator

**MFJ's** unequaled tuning indicator makes it really easy to work HF packet.

**And** unlike others, you use it exactly the same for all modes -- not differently for each mode. Just tune your radio to center a single LED and *you're precisely tuned in to within 10 Hz* -- and it shows you which way to tune!

### Multi-Gray Level FAX/SSTV Modem

**You'll** see tomorrow's news today when you copy outstanding FAX news photos with crisp, clear details.

**MFJ-1278** is the *only* multi-mode with a built-in multi-gray level modem. It lets you transmit *and* receive high resolution multi-gray level FAX/SSTV pictures with an appropriate terminal program.

## New MFJ-1278T *Turbo*

### with *fast* 2400 baud modem

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# MARITIME MOBILES

# NEED SIMPLE RIGS AND GOOD INSTALLATIONS

## Sound tips keep maritime operations afloat

*By Clyde Kirlin, WB6VPX, P.O. Box 2116, Walnut Creek, California 94595*

Sailboats and Amateur Radio have been mates afloat for many years. Sailors have frequently explored ocean-going Amateur Radio as a hobby, but in recent years it has been used more as a lower cost substitute for a single sideband (SSB) marine radio. Today's technology makes it easier than ever for the multitasked sailor to use ham gear.

Dial twisting is a thing of the past. Memory bands now hold the frequencies we most commonly use. Microprocessor-controlled automatic antenna tuners quickly adjust for the chosen frequency. Push buttons, rather than analog rotation devices, simplify RF gain, mode selection, and tuning dial adjustments.

While operations are streamlined, good installation practices are often forgotten in the haste to "get on the air." Successfully fitting a radio to a sailboat involves creating a good ground plane, installing an exceptional antenna, and placing the equipment in proper relationship to antenna, ground plane, and power source. Additionally, the operator's transceiver selection process is often affected by brand preference, peer pressure, and impulse buying. Buyers frequently ignore the basic requirements and specifications that comprise a good receiver, as well as available power sources at sea.

### Is the best really the best?

Take the case of my good friend Sam, who was about to fit his sailboat with an ICOM-781. Knowing he was a pragmatist at heart, I felt he had been swayed by the unsurpassed quality and features of the IC-781. Granted, it is a superb transceiver in a base station environment, but it wasn't designed for a sailboat. The IC-781's receiver performance is highly rated and I don't know of any radio that can equal the tricks performed with the IC-781's built-in keyer, filters, and twin passband tuning. Consider, however, what's required to maintain this performance level on a sailboat. Would Sam be using RTTY, AMTOR, and CW at sea? I doubt it. Good voice communications meet most objectives — no need for full break-in, noise blankers, and CW filters in that mode.

Because the 781 operates only on 120 volts AC, a 12-volts DC inverter or fuel-driven AC generator would be needed on a sailboat. A major maker of inverters agreed with my observations that a "demand-type" inverter wouldn't respond quickly enough to the fluctating power demands of a CW or SSB signal. And who wants to listen to a noisy generator on board while enjoying the pleasure of Amateur Radio? Further, the IC-781's automatic antenna tuner is designed for a 50-ohm output and not suited to match the varying impedance of an unloaded whip or random wire, as would be used on a sailboat.

Several other models using the same upbeat technology, but with simplified operation, stood at the head of the selection line. ICOM's compact IC-725 with its AH-3 automatic tuner* would be an excellent choice, as would economical performers from Kenwood† and Yaesu‡.

### Many questions and some answers

Sam continued his quest for the best maritime setup with questions on how to create an antenna ground plane, the use of a mobile center-loaded whip or the boat's backstay as an antenna, and antenna length.

In my opinion, receiver specifications come first. The old adage applies, "If you can't hear 'em, you can't talk to 'em." Receiver sensitivities of less than 0.5 $\mu$V are quite common and most adequate for the quiet atmospheric environment of radio work at sea. Automatic antenna tuners will match almost any wire span that isn't near a half-wave in length in less than three seconds! All the operator needs to do is pick up the microphone and talk.

### What makes a good installation?

A cruising sailboat is usually between 35 and 40 feet long, made of fiber glass, and rigged with one or two masts. Most sailboats carry a set of 12-volt DC batteries which are charged intermittently by the boat's engine alternator. Preplanning the overall installation layout is important. Because the antenna system will be similar to that of a quarter-wave ground plane vertical antenna, the antenna, antenna tuner, and ground plane must be in a reasonable vertical alignment. Mount the antenna tuner in a dry location in the lazarette area under the cockpit. Connect the antenna to the tuner here. The radiation current node

* ICOM America, 2380 116th Avenue, NE, Bellevue, Washington 98004.
† Kenwood U.S.A. Corporation, PO Box 22745, Long Beach, California 90801-5745.
‡ Yaesu USA, 17210 Edwards Road, Cerritos, California 90701.

TABLE 1

**Antenna Length Calculations**

Commonly used Amateur frequencies and the decimal portion of a wavelength that various antenna lengths represent

1. One-half wavelength = 492/Frequency (MHz)
2. One-half wavelength x 2 = one wavelength (W/L)
3. Antenna length/one wavelength (W/L) = decimal portion of W/L

| | Antenna Lengths in Feet | | | | | |
|---|---|---|---|---|---|---|
| | 23.00 | 30.00 | 35.00 | 40.00 | 45.00 | 50.00 |
| Frequencies (MHz) | Wavelengths | | | | | |
| 7.25 | 0.17 | 0.22 | 0.26 | 0.29 | 0.33 | 0.37 |
| 14.35 | 0.34 | 0.44 | 0.51 | 0.58 | 0.66 | 0.73 |
| 21.35 | 0.50 | 0.65 | 0.76 | 0.87 | 0.98 | 1.08 |
| 28.35 | 0.66 | 0.86 | 1.01 | 1.15 | 1.30 | 1.44 |

begins at the tuner output. Keep the portion of antenna wire inside the boat as short as possible to increase power radiation outside the hull.

Radiation resistance is composed of several factors, including ground resistance. Unfortunately, the antenna tuner usually is situated at some point above ground. Make the tuner think that it's sitting in sea water (ground) by using multiple copper ground straps between the tuner and the ground plane. Keep the straps as short as possible to reduce the ground resistance. The transceiver should be located at a convenient operating position, as close as possible to the batteries.

## The antenna selection

Choose an antenna that fits the boat's rigging configuration, but is as long as possible. With today's high aspect ratio of tall mast and short boom, the boat's boom doesn't overhang the stern. This leaves an area for mounting a 23-foot vertical unloaded whip, like a Shakespeare Model 390,* or a mobile center-loaded whip designed for a specific frequency. Of course, this leaves open for debate the question of insulating the backstay for longer antennas, creating higher efficiency on lower frequencies — versus the possibility of backstay failure in heavy weather.

Being safety minded, I opt for the 23-foot unloaded whip or the mobile center-loaded whip. With either of these, there are cautions to be considered. The 23-foot whip works well on 10, 20, and 40 meters. However, at 15 meters, it's close to one-half wavelength and won't load properly. The mobile whip must be mounted at the stern, using the metallic stern pulpit as a mounting base and possible ground plane. This creates a dangerous situation for the sailor attempting to change loading coils in a heavy seaway. Also, with the pulpit as a ground plane, a portion of the radiation — the lower 2 or 3 feet of the antenna mounting staff — is often wasted because it's sheltered by the ground plane itself.

## How long is a vertical antenna?

The actual antenna length includes the distance from the tip of the antenna, the antenna lead-in to the tuner, through the ground lead from tuner to ground. Because antennas don't load very well at multiples of one-half wavelength, the formula in **Table 1** outlines the calculation of antenna lengths to be avoided.

For 10, 15, 20, and 40-meter operation, insulate the backstay to produce an antenna and ground lead length of 30, 40, or 45 feet. Note that a 35-foot antenna is close to a quarter wavelength at 40 meters — an ideal vertical antenna length. Yet the same antenna crowds the half-wavelength point on 20 meters. The total length from tip to ground should not be a multiple of one-half wavelength for any frequency being used.

## The backstay

The backstay, when used as an antenna, must have strain insulators inserted in the wire. A good backstay insulator must be able to withstand 15-kV breakdown voltage and have the physical strength to endure the rigors at sea. Any backstay insulator work is best left to a professional rigging shop, for both selection and fitting. They must assume the responsibility for insulator failure and possible dismasting.

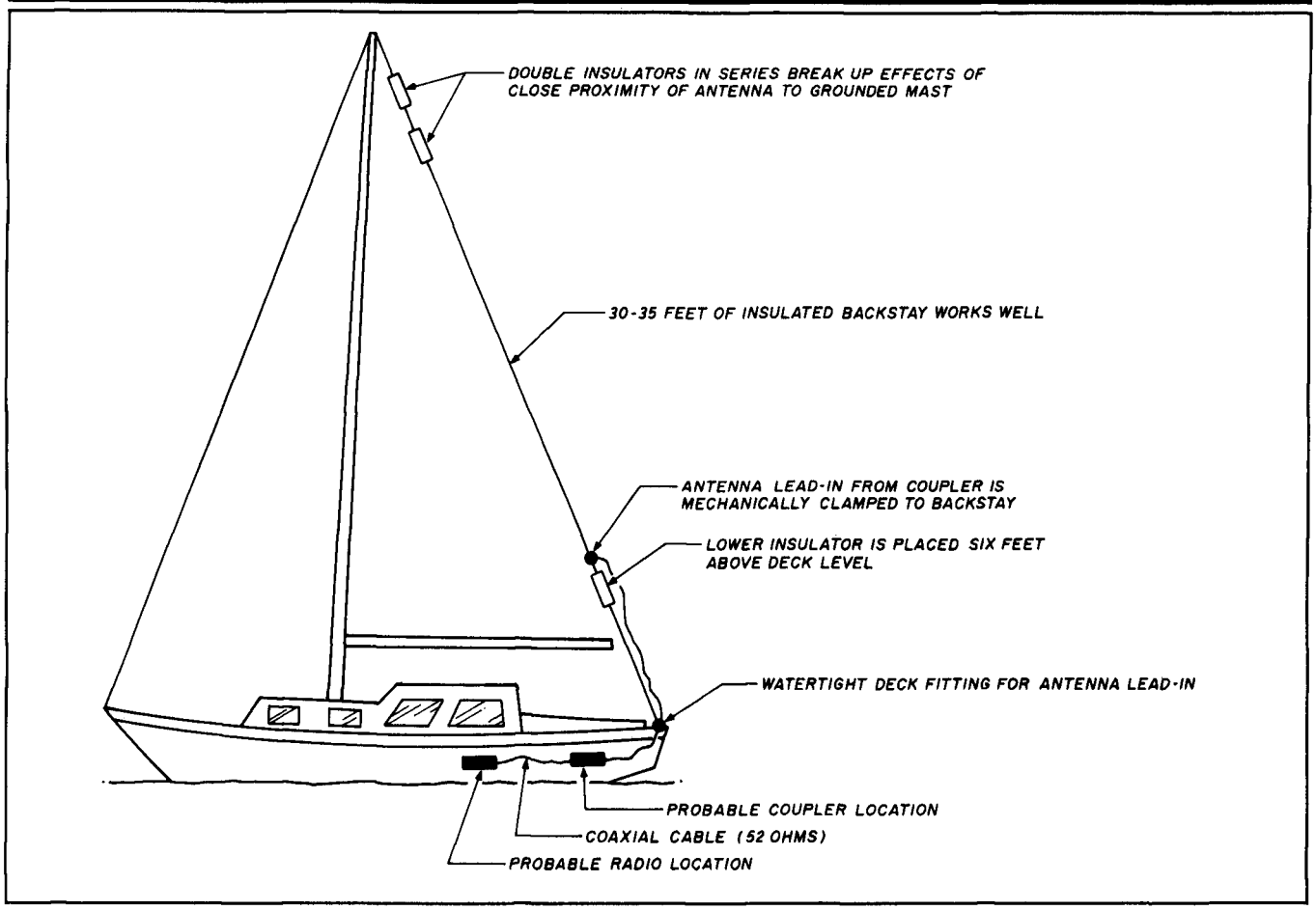
Have the lower insulator placed in the backstay at a height of 6 feet above the deck level, so those looking for a steadying grip at sea don't grab onto an antenna that's radiating power. Make sure there's high voltage insulation (25 to 30 kV) on the antenna lead from the coupler to the insulator.

The placement of the upper insulator depends on the distance between the point where the antenna tuner connects to the ground plane and the upper end of the desired antenna length. To offset any detrimental ground effects that a metal mast introduces to the antenna, a third insulator may be placed at the upper end of the backstay, within a foot or so of the masthead. "Living dangerously?" you ask. "A greater chance for dismasting," you observe. Yes, but this is why the insulator task is best left to professional riggers.

The antenna lead-in from the tuner to the backstay is a piece of high voltage insulated wire like GTO-15 neon sign wire or the polyethylene dielectric core of RG-8/U. One end of the lead-in is secured to the insulated portion of the backstay with U-bolts mounted close to the lower insulator. Wrap the U-bolt clamp with chafe protection. The tuner end of the lead-in is fed through a watertight deck fitting into the lazarette area near the antenna tuner (see **Figure 1**). If you have a metal deck, use a porcelain feedthrough insulator to keep the antenna lead-in as much as 6 inches from the

* The Shakespeare Company Antenna Group, PO Box 733, Newberry, South Carolina 29108.

FIGURE 1




**Pre-planning the overall layout of the antenna system — the antenna, antenna tuner, and ground plane — for best results.**

metal. Also, if the antenna lead-in is to be fixed in place on a metal hull, dress it 6 inches from all metal.

## Creating the ground plane

What constitutes an adequate ground plane when afloat? Where should it be positioned? How large should it be? What material should be used? The only reasonable location for an antenna on a sailboat is far aft. The antenna, antenna tuner, and ground plane should be in reasonable vertical alignment. The only logical place for the ground plane is aft.

It's difficult to relate the ground plane area afloat with the number and length of radials used ashore. A coupling effect to the water exists if a ground plane is mounted low in the boat. The higher the ground plane is above the water, considering its necessarily limited size, the less efficient the system will be.

The best material for a ground plane on a boat is copper window screen, 16 mesh, found at most hardware stores. Cut the screen to the shape of the area in which it is to be mounted. Seam solder the wires of two edges at right angles to each other with a 2-inch wide, cold-rolled and annealed copper strap of 0.005-inch thickness. This way you'll know that all of the individual wires are a part of the ground plane. The copper strap readily allows other portions of the screen to be attached together with soldered joints. Secure the antenna tuner to the ground plane with several copper straps attached at different points on the ground plane. Look for copper straps at your local sheet metal supply house.

A good ground plane will lie high in the bilge (out of bilge water) in close proximity to the sea water. There's a degree of capacitive coupling to the sea water through the non-metallic hull. As an alternative layout, use the underside of quarterberths — those berth areas generally found on either side of the cockpit or engine — for part of the total ground plane. Bond these portions of the screening with copper strap soldered to the bilge portion of the ground plane (see **Figure 2**).

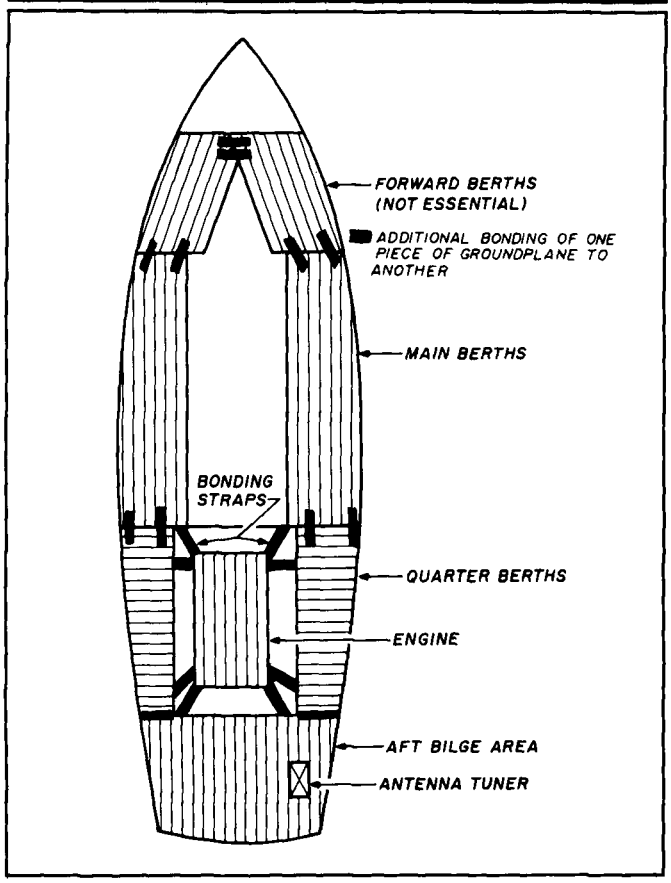
In **Figure 3** you'll see that tankage, metal hydraulic tubing, fire extinguisher piping, water lines, lifelines, toerails, engine, shaft, through-hull fittings, propeller, and rudder constitute the rest of the ground plane when bonded to other ground plane portions with copper strap and soldered joints.

With any boat installation, everything is a compromise. One could claim that 400 square feet of ground area is required, but a horizontal area that large is difficult to obtain on a boat. Yet a ground plane of 10 by 15 feet is the absolute minimum for good efficiency. The limitations of boat size and layout face you. Do the best you can.

## Wrapping it up

With the antenna and the ground plane in place, and the antenna tuner positioned and attached to the ground plane,

FIGURE 2




**What the antenna would see, if it were to search out possible candidates for ground plane or counterpoise areas. The areas of the cabin sole would add to this, but only if neatness could be maintained by covering screening with carpeting. However, in heavy weather cruising carpeting isn't always practical because it will get wet.**

the transceiver can be wired to the boat's battery system. Good practice dictates that a dedicated electronic wiring bus be used to bring full power to an electronic distribution circuit breaker panel.

One hundred-watt transceivers draw 20 A when transmitting. The wire size between the battery and transceiver should be selected to limit the voltage drop to about 3 percent.

The American Boat and Yacht Council (ABYC)* recommends using AWG no. 6 wire if the length of the conductors from the battery to the transceiver **and back** to the battery is between 25 and 40 feet. Any longer wiring run will require AWG no. 4 wire.

Most manufacturers will provide a power cable less than 6 feet in length of an AWG no. 10 wire size. This may be used only if the battery is located within the length of cable provided. Otherwise, this cable should be shortened to less than 2 feet and terminated on a barrier terminal strip using crimp lugs. Use heavier wiring of a size that follows the ABYC dictates between the terminal strip and the distribution panel.

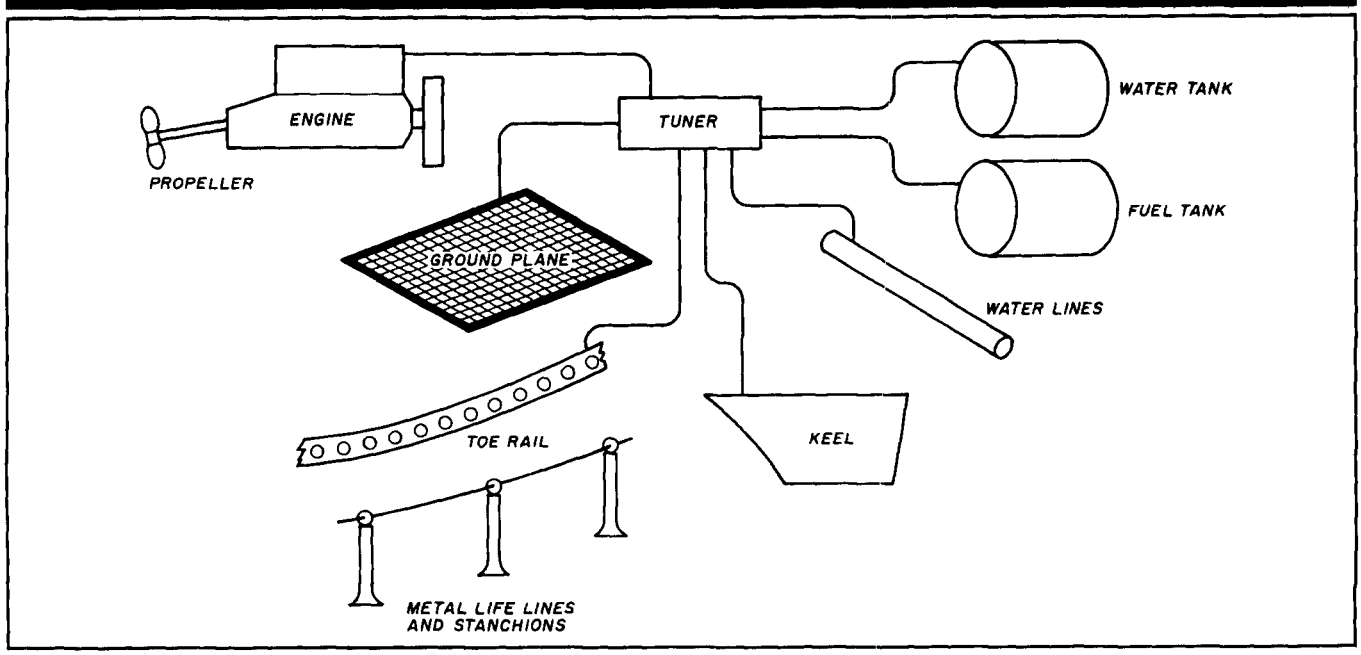
Wire the tuner and transceiver as directed and connect the two units with a piece of RG-58/U antenna lead-in cable to complete your system.

## Time to get on the air!

All antenna and ground elements are secured in near vertical alignment. The ground plane is as large and as horizontal as you can make it, and located close to the bilge in the after section of your boat. Finally, due to your professional wiring, there's a full 12 volts DC at your rig when transmitting. If all the work has been done with care, there's not much left to do except to get on the air and listen to those gratifying comments from 10,000 miles away, "You're 20 over 9!" hr

* The American Boat and Yacht Council, PO Box 806, Amityville, New York 11701.

FIGURE 3




**Make sure you bond to everything for a good ground plane, or counterpoise.**

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# Microwaves

*By Bob Atkins, KA1GT*

## OPTICAL RECEIVER

Last month I described the requirements for the transmitter end of an optical communciations link. This month I'll take a look at the other end of the link — the receiver.

### The optical receiver

Just as the equipment used in the laser transmitter has RF analogs, so do the components used in an optical receiver like the one shown in **Figure 1**. Probably the best analogy is that of a crystal set. The antenna of a crystal set would correspond to the lens (or mirror) of the optical receiver. The crystal set's tuned circuit corresponds to the optical filter, the crystal itself is analogous to the optical detector, and the headphones of the crystal set are replaced by the amplifier section of the optical receiver.

The first function of the optical receiver is to capture as much of the transmitted light as possible and to direct that light towards the optical detector. The function is the same as that of a telescope, with the receiver as the eye, so it's no surprise that an optical receiver closely resembles a telescope. Either a large lens or mirror can be used to capture the light. For visual use, the optics of a telescope system must be shaped with extreme accuracy. In fact, they must corresond with their designed shapes to better than one-quarter the wavelength of light (a few millionths of an inch). Though such optics would work well in an optical receiver, they are very expensive in large sizes. Rather than form a detailed image, the optics of the optical receiver need only concentrate the incoming light into an area less than or equal to the area of the optical detector; thus lower quality optics can be used to good effect.

### The fresnel lens

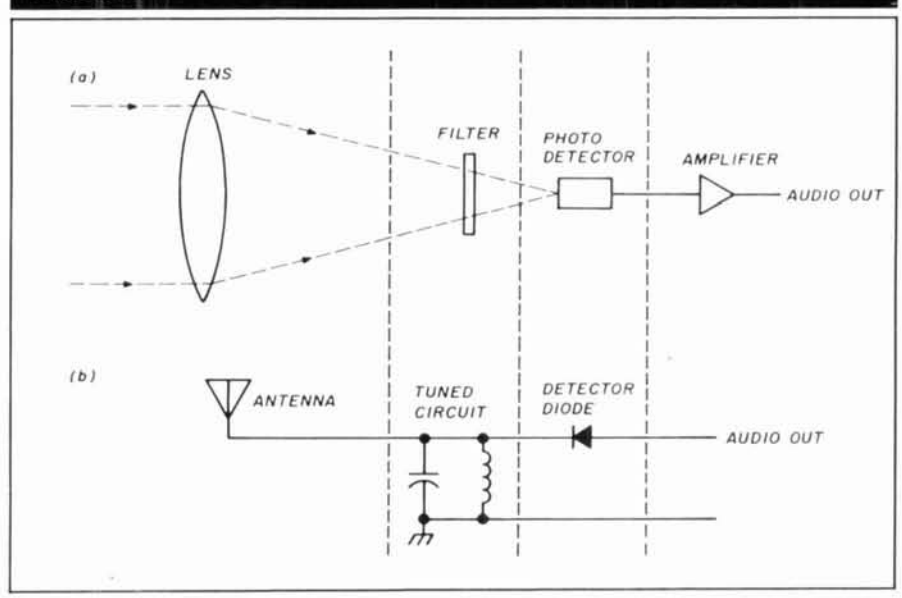
One type of lens which has been used in optical receivers is known as the fresnel lens. This lens is comprised of concentric rings of prismatic elements, each of which bend the incoming light towards the lens's focal point as shown in **Figure 2**. The fresnel lens has advantages over conventional optics because it is much cheaper, much thinner, and lighter. Unfortunately, the fresnel lens produces a very poor quality visual image. But, as I pointed out earlier, this may be of no consequence in an optical receiver. To compare prices, a 6-inch diameter optical lens might cost $700 and a 6-inch parabolic mirror might run $150, but a 12-inch fresnel lens (which captures 4 times as much light as the 6-inch optics) can be purchased for around $30.

### Interference and absorption filters

As mentioned last month, the light from a laser is monochromatic (i.e., composed of only one color). The background light entering the optical receiver will, in general, be composed of all visible colors plus some infrared and ultraviolet wavelengths. These unwanted background optical signals can be thought of as "optical QRM." A filter can be used to select out only the laser light, just as an RF filter is used to select one specific RF frequency.

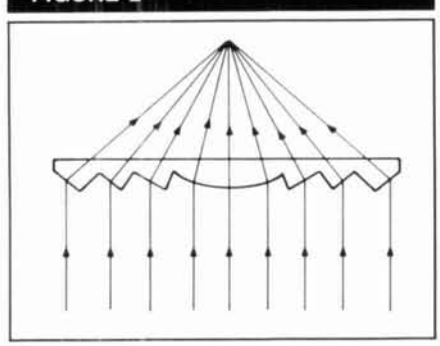
There are two commonly used filter types — interference and absorption filters. Interference filters are composed of many layers of different dielectric

**FIGURE 1**




**(A)Schematic of a simple optical receiver. (B) Analogous components of a crystal set.**

**FIGURE 2**



**Cross-section of a fresnel lens.**

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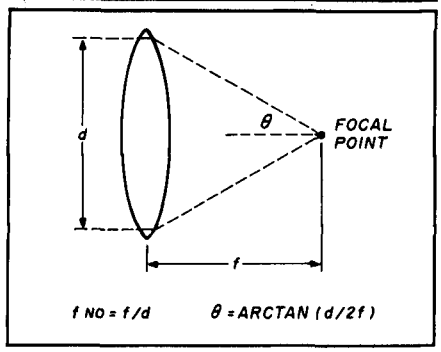
materials and work on the principle of the constructive and destructive interference of light. These filters have a very narrow passband, analogous to high Q cavity filters at RF. With interference filters, the wavelength of maximum transmission depends on the angle at which the light strikes the filter. This is of consequence when such a filter is used in conjunction with small f-number receiving optics.

The f-number of a lens is given by the focal length of the lens divided by its diameter. As you can see in **Figure 3**, light captured by the edge of such a lens will arrive at its focal point at a large angle (theta). This may be sufficient to cause the transmission maximum of an interference filter to shift to such an extent that light of the laser's wavelength is no longer transmitted. For a 10-nm wide filter at the He-Ne wavelength of 632.8 nm, light coming from the edge of an f2 lens will hit the filter at a 14-degree angle of incidence. This is enough to shift the filter peak response from 632.8 nm to around 628.4 nm, reducing the transmission by around 50 percent. Because only the light from the edges of the lens suffers this much attenuation, this configuration is acceptable. However, if an f1 lens is used (and this is not unusual for a fresnel-type lens), 632.8-nm light from the outer edges of the lens won't pass through the filter at all, and only light collected by the central region of the lens can be used.

What this all boils down to is that a 6-inch diameter f2 lens (12-inch focal length) will be as efficient at light collecting as a 12-inch f1 lens (12-inch focal length ) when a 10-nm bandwidth interference filter is used. If you have a lens whose f-number is less than 2, there are two ways to use it efficiently. A small diverging (convex) lens can be placed before the filter. This converts the converging beam from the main lens to a parallel (or much less converging) beam, lessening the angle at which the light enters the filter. The other alternative is to use an absorption filter instead of an interference filter. Absorption filters absorb light of certain wavelengths and allow all others to pass through. In RF terms, they can operate like low pass, high pass, or bandpass filters of low to moderate Q.

In summary, the interference filter has the advantage of narrow bandpass (and hence high background light rejection) at the expense of sensitivity to incidence angle and cost. On the other hand, while absorption filters are low in cost and insensitive to incidence angle, they do not provide the background rejection of interference filters. For daytime operation the use of an interference filter is highly desirable, perhaps even essential, while nighttime operation is possible with an absorption filter, or even no filter if there's little or no background light.

FIGURE 3




**Definition of the f number of a lens. Note: When the f number is small, angle θ is large. This can be a problem if an interference filter is used in the optical path (see text).**

For He-Ne lasers, a 632-nm interference filter 0.5 inches in diameter with a 10-nm bandpass is available for $38 from Edmund Scientific (see last month's column). A red photographic filter (Kodak types 29, 26, or 25) would be acceptable as an absorption filter for use with He-Ne lasers. A 2-inch diameter filter should be available from local photographic equipment suppliers for less than $10.

After capturing and filtering the light, the next step is to convert the optical energy into electrical energy using an optical detector. There are two detector types of practical interest, the photomultiplier and the photodiode.

The photomultiplier is an electron tube device which contains a number of internal electrodes (dynodes). An overall supply voltage of around 1000 volts is split evenly between the electrodes. When a photon enters the tube it impacts on a photocathode and ejects an electron (this is known as the photoelectric effect). This electron is then accelerated to the nearest electrode by the potential difference between them. When it hits the electrode, it ejects several more electrons. These electrons are accelerated to the next electrode, where they strike it and emit even more electrons, and so on. Finally, after passage via the multiple electrodes, the initial electron release is effectively multiplied by up to 1 million times, and a measurable current is generated.

In contrast to the high voltage/amplification characteristics of the photomultiplier, the photodiode is a low voltage device with no intrinsic gain. It operates much like a solar cell, generating a small current when exposed to light. To be useful at low light levels, the current generated is converted to a voltage and amplified. This is accomplished most simply using an op amp configured as a transimpedance amplifier. Some photodiodes are available with a built-in high gain amplifier in the same package. This is a desirable configuration because, owing to the very high gains needed ($>10^5$), pickup of stray signals is a problem. These signals can be minimized by housing the photodiode and amplifier together in the same shielded package.

## Choosing an optical detector

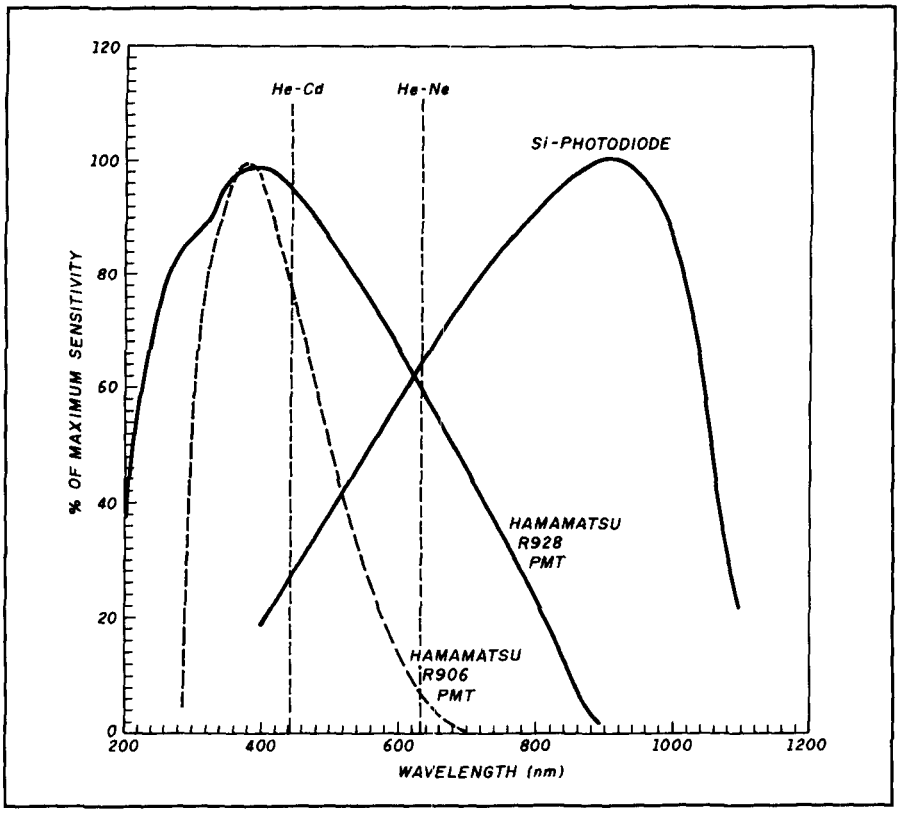
The choice of optical detector depends on a number of factors including spectral response, size, sensitivity, and convenience of use. Most photomultipliers are more sensitive in the blue region of the spectrum than the red. A typical tube might be 10 times more sensitive at 442 nm (He-Cd) than at 632 nm (He-Ne). Some tubes do have extended red response, but they are less common on the surplus market and more expensive. Photodiodes usually show maximum response in the near infrared at around 900 nm. They are usually about 1.5 times more sensitive at 632 nm than at 442 nm. The shapes of the response curves for several photodetectors are shown in **Figure 4**.

Detector area is also an important parameter. For maximum receiver efficiency, all the light collected by the lens must hit the active area of the detector. Using a fresnel lens will result in a larger image than would be obtained using high quality optics. It's evident, therefore, that a larger area optical detector is desirable because it will increase the probability that all the collected light will hit the detector. For the same reason, a large detector area also lowers the required pointing accuracy. A potential drawback of a very large area detector is that, given

reasonable imaging optics, it will collect light from a wider area than a small area detector (hence the easier pointing). This means that if there's a lot of background light, it could swamp out the laser signal. This shouldn't be a problem at night and/or when efficient narrowband filters are used. A typical photomultiplier has quite a large active area of around 1 cm × 2 cm.

Photodiodes are available in a range sizes from an active area of several square centimeters down to less than 1/10 of a square millimeter. Very small area detectors, like those of the Motorola MRD series (MRD 360 photodarlington, MRD 510 photodiode, and MRD 300 phototransistor) are quite sensitive and inexpensive ($2), and even include photodarlington transistor detectors which effectively show intrinsic gain. Unfortunately, for the reasons given previously, their extremely small active area (0.03″ × 0.03″ maximum) makes them a poor choice for most free space DX optical receivers. However, they are good and inexpensive devices for initial experimentation. Large area detectors are normally more expensive ($10 to $30), but will provide much better results in most cases. They do generate a little more internal noise, though in this application it shouldn't be enough to cause a problem.

FIGURE 4




**Relative response curves for photodetectors.**

## Noise equivalent power

Ultimate detector sensitivity is limited by the intrinsic noise generated internally by the device. There is a measure of this noise known as the noise equivalent power (NEP). This is roughly comparable to the noise figure of a diode or transistor. NEP is defined as the amount of optical power required to generate a signal equal to the intrinsic noise level of the detector. Therefore, it's also the amount of optical power required to give a 0-dB signal-to-noise ratio. Because noise is a function of bandwidth, it's expressed in units of W/Hz$^{1/2}$. The NEP of a good photomultiplier is about 10 times lower than that of a good photodiode. Some typical NEP values for devices which would be suitable for use in Amateur optical receivers are:

Hamamatsu R906 photomultiplier
at 440 nm = 3e-16 W/Hz$^{1/2}$
(blue sensitive, $70)
at 632 nm = 6e-15 W/Hz$^{1/2}$

Hamamatsu R928 photomultiplier
at 440 nm = 1e-16 W/Hz$^{1/2}$
(red sensitive, $300)
at 632 nm = 2e-16 W/Hz$^{1/2}$

Hamamatsu S2386-8K photodiode
at 440 nm = 8e-15 W/Hz$^{1/2}$
(area = 33mm$^2$, $30)
at 632 nm = 3e-15 W/Hz$^{1/2}$

UDT-020D photodiode/amplifier
at 440 nm = 2.5e-14 W/Hz$^{1/2}$
(area = 20mm$^2$, $43)
at 632 nm = 5.1e-14 W/Hz$^{1/2}$

If the laser transmitter is tone modulated before keying (MCW or A2A modulation as described last month), all that's required to produce an audible CW signal is to amplify the output of the optical detector and feed the signal into an audio amplifier. A narrowband audio filter centered on the modulation frequency can be included for improved performance. If the laser is not tone modulated (CW or A1A modulation), and if DC coupling is then used throughout, the output of the photodetector after the transimpedance amplifier is a voltage level which rises and falls as the transmitter is keyed. To produce an audible signal, this voltage level must be used to key a tone on and off. This can be achieved by feeding the amplifier output into one input of a comparator. A reference voltage is then fed into the other comparator input and adjusted so the comparator output is low when the laser is off and high when the laser is on. The output of the comparator can then be used to key a tone on and off — or its digital state can be read directly by a computer.

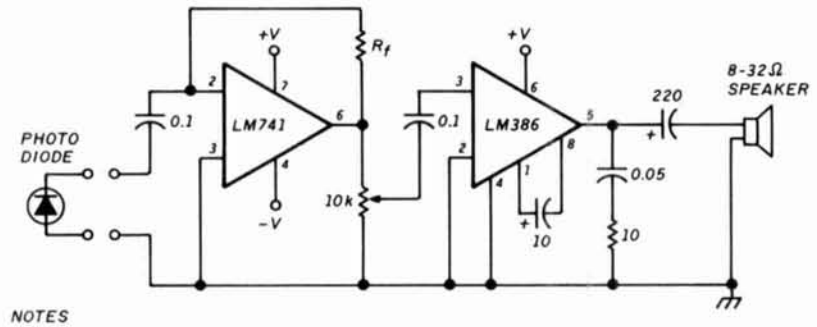
**Figure 5** shows the electrical schematic for a complete optical receiver suitable for the reception of keyed tone-modulated transmissions. If an integrated detector/amplifier is used, this would replace the first amplifier stage. Note that such detector/amplifiers are DC coupled, which means that their output saturates (reaches the supply voltage) with fairly low levels of incident light. For example, the EG&G HUV 2000B, using a ±15 volt supply, will saturate at about 3×10$^{-7}$ watts of incident light energy when operating at maximum "gain." (Strictly speaking, "gain" is an inappropriate term because the amplifier is operating as a current-to-voltage converter.) In a well-constructed optical receiver,

and with a low level of background light, the detection of limit will probably be in the range of $10^{-10}$ to $10^{-12}$ watts of modulated laser energy — depending on the quality of construction, noise levels generated in the amplifier stages, the receiver bandwidth, and how well you can copy weak CW! For those interested in a detailed mathematical analysis of the noise components of optical detectors, check out the EG&G Photon Devices catalog cited at the end of this article.

**EG&G Photon Devices** — Manufacturer of photodiodes. Small photodiodes from $17, diode/amplifier combinations from $70. No minimum order. Catalog available. Contact EG&G Solid State, 35 Congress Street, P.O. Box 506, Salem, Massachusetts 01979-6526.

**Motorola** — Manufacturer of small diodes and $21 for small photodiode/amplifier combinations. Minimum order $100. Data sheets available. Contact UDT, 12525 Chadron Avenue, Hawthorn, California 90250.

**FIGURE 5**




**Schematic of the electronic stages of an optical receiver.**

## Suppliers

Here are a few sources for the components required to construct an optical receiver. The catalogs and data books referenced are free to commercial users, but there may be charges to individuals who order them. This is not an exhaustive list; inclusion or omission from the list does not imply endorsement or otherwise.

**Hamamatsu** — Manufacturer of photomultipliers and photodiodes. Large range of products includes PMTs from $70, photodiodes from $10. Minimum order $25. Will take credit cards. Catalogs available. Contact Hamamatsu, Order Service Department, 360 Foothill Road, Box 6910, Bridgewater, New Jersey 08807-0960.

**United Detector Technology** — Manufacturer of photodiodes. Good selection, from $10 for small photoarea optoelectronic devices (photodiodes, phototransistors, photodarlingtons), mostly in the $2 to $15 price range. Optoelectronics Data book available from Motorola Literature Distribution, P.O. Box 20912, Phoenix, Arizona 85036. Devices available through distributors (e.g., Newark, Hamilton-Avnet).

See last month's column for suppliers of optics and filters.

Next month, in the final article on optical communication, I'll discuss the factors (principally scattering) which influence optical communication range and I'll show you how to calculate the DX potential of a laser communication system. Please send comments, questions and any microwave activity reports to Bob Atkins, KA1GT, 103 Division Avenue, Millington, New Jersey 07946.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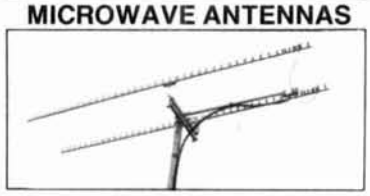

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| 2345LYK | 45el | loop Yagi Kit | 1296 MHz | 21 dBi | $89.00 |
| 2445LYK | 45el | loop Yagi Kit | 1269 MHz | 21 dBi | $89.00 |
| 1844LY | 44el | loop Yagi (assem.) | 1691 MHz | 21 dBi | $99.00 |
| 2355LYK | 55el | Superlooper Kit | 1296 MHz | 22 dBi | $99.00 |
| 1345 LYK | 45el | loop Yagi Kit | 2304 MHz | 21 dBi | $75.00 |
| 945LYK | 45el | loop Yagi Kit | 3456 MHz | 21 dBi | $75.00 |

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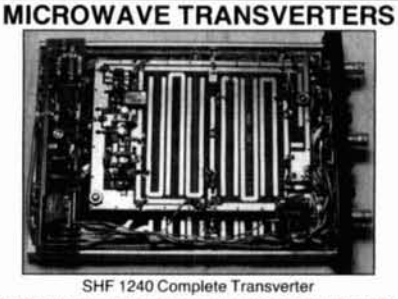

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| **SHF 1269K** | **1268-1272 Oscar Mode L** | **10mW** | Kit $130 | Built $240 |
| **SHF 2304K** | **2304-2308 MHz** | **10mW** | Kit $195 | Built $320 |
| **SHF 2401K** | **2400 MHz Mode S rcv Conv** | | Kit $150 | Built $250 |
| **SHF 3456L** | **3456-3460 MHz** | **10mW** | Kit $195 | Built $320 |
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| 2340 PA | 1W in 35W out | 1240-1300 MHz | $355 |
| 2370 PA | 5W in 70W out | 1240-1300 MHz | $695 |
| 3318 PA | 1W in 20W out | 902-928 MHz | $265 |
| 3335 PA | 14W in 40W out | 902-928 MHz | $320 |
| 1303 IPA | 10mW in 0.4W out | 2304 MHz | $150 |
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| 1691LNAWP | preamp 1 dB NF | 1691 MHZ mast mounted | 13.8V | $140 |
| 4017LNAK | preamp kit | 400-1700 MHz | .6 dB | $ 40 |

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# A 220-MHZ RECEIVING ADAPTER

*By Peter J. Bertini, K1ZJH, 20 Patsun Road, Somers, Connecticut 06071*

Catch the wave — 1-1/4 meter waves, that is — with your 2-meter handheld and this simple converter! This weekend project allows casual 1-1/4 meter monitoring on your 2-meter handheld or VHF scanner. The self-contained converter is small enough to mount directly on the antenna jack of most BNC outfitted handhelds. Many converter designs have been published, but I wanted a circuit that didn't need an etched pc board. I also wanted to make this project as simple as possible!

## Works both ways!

The converter uses a passive diode mixer and is bilateral; you can also receive 2-meter signals on a 220-MHz receiver! In this mode the converter is connected "backwards." My service monitor, an older Cushman CE-3, covers 144 MHz but not 220. With this converter, I can use the monitor to perform accurate frequency/deviation monitoring and limited signal generation on the 1-1/4 meter band (more on this later).

A VHF converter may use several stages. You might find an LO oscillator with one or two multiplier stages, a mixer, and usually an RF amplifier. This is hardly something you could build in a weekend or would want sitting on top of your 2-meter HT! I make do with only a mixer and oscillator. The mixer has fairly high losses and the ultimate sensitivity is determined by the 2-meter IF receiver.

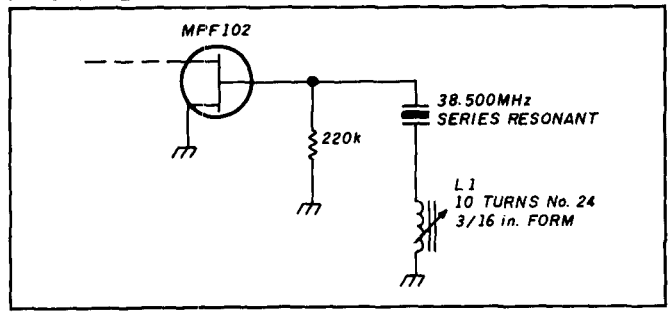
## Construction starts with plastic case

The converter is housed in a plastic Radio Shack project case. A matching perfboard with 0.1-inch spaced component mounting holes comes with the case. A VHF converter would normally be built on an etched pc board. I avoided this by mounting a piece of sheet brass, slightly less than 0.7 inches wide, transversing the center length of the perfboard. All component leads going to ground connect directly to this strip. The brass stock is available at hobby shops and can be cut easily with tinsnips. A thickness of 0.01 inch will do. When mounting the trimmer capacitors, connect the rotor plates to ground. This lets you use metallic tuning tools during alignment, except when you're working with the trimmer on the oscillator secondary winding where both plates are at RF potential.

The BNC connector ground lugs are connected to the ground plane. This also holds the finished converter board in place. Remove the two plastic standoffs inside the case, so the converter board will sit flush with the case bottom. Carefully break away the standoff material with a small pair of needle-nose pliers. You may clean any remaining rough edges with a penknife. With care, everything — including the 9-volt battery — will fit inside the case. Parts layout isn't critical.

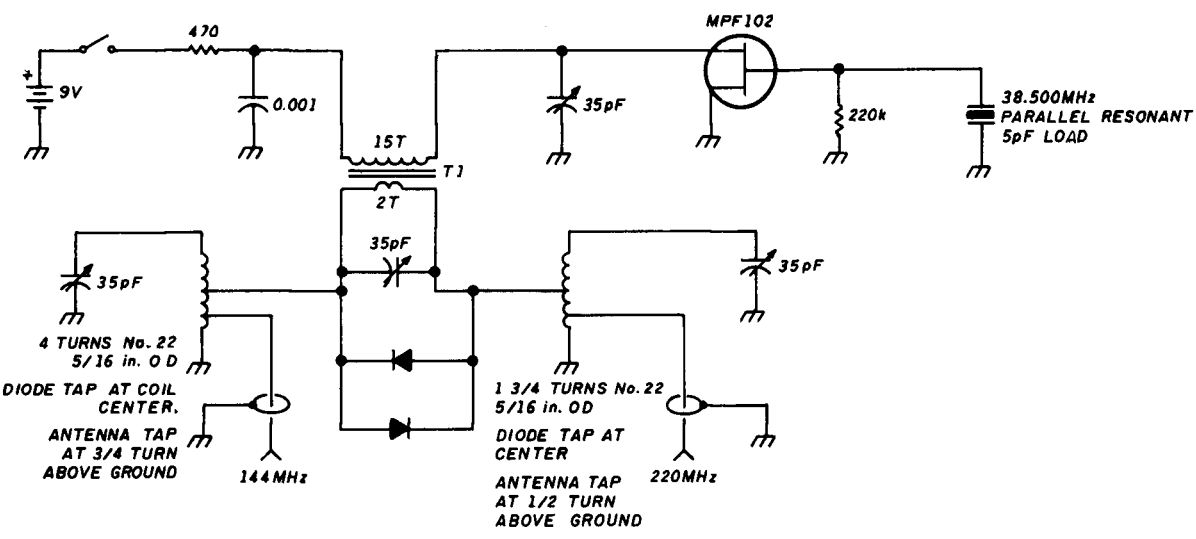
I used chassis mount BNC female connectors. You'll need a double male BNC adapter to connect the converter to most handhelds. You can also use a cable end male connector on the case for the radio connection. Use the rear cable nut to mount the connector through the enclosure. Prepare a short piece of bare hookup wire and solder it to the male pin of the BNC connector. Insert the pin so it's properly seated in the connector. Several drops of epoxy will keep the pin from pushing back when mated.

FIGURE 1




**Variation of oscillator circuit uses series-resonant crystal with adjustable series inductance to permit precise setting of oscillator frequency.**

FIGURE 2




**Receiving converter adapter circuit. Oscillator is shown with parallel-resonant crystal cut for 5-pF load and is accurate to within a few hundred cycles. Coil details are given in the test and parts lists.**

PARTS LIST

Capacitors

4 *Trimmers, between 30-40 pF maximum, plastic or ceramic miniature.*
1 *Ceramic disc, 25 or greater working volts 1000 pF (0.001 μF)*

Coils

1 *220-MHz antenna coil, 1-3/4 turns no. 22 tinned wire, air wound. 5/16-inch OD, center tap for diode connection. Antenna tap is at 1/2 turn above ground.*
1 *144-MHz antenna coil, 4 turns no. 22 tinned wire, 5/16-inch OD, center tap for diode connection. Antenna tap is at 3/4 turn above ground.*
1 *L1 warping inductor. About 12 turns AWG no. 24 magnet wire solenoid wound on 1/4 to 3/16-inch slug-tuned form. Used only with series resonant cut crystal.*

Crystal

1 *38.5000-MHz, third overtone cut, HC-18u wire lead case. Note: order ICM factory no. 4713601 for general use. Parallel resonant, 5-pF load.*

*Order ICM factory no. 471360 if it is desirable to set frequency. Series resonant mode. Used with series inductor to set frequency.*

Diodes

2 *High speed switching, silicon 1N914, 1N4148 or hot carrier type.*

Project case

1 *Radio shack 270-283 or similar*

Resistors

1 *1/4 watt carbon or film composition, 20 percent or better*
1 *470 ohm*
1 *220 k*

Transistor

1 *FET, MPF102 or 2N5486*

Transformer (T1)

1 *Toroid, T-37-6, yellow SF mix, 0.37-inch OD primary winding 15 turns AWG no. 24 enamel magnet wire, secondary winding 2 turns AWG no. 24 enamel magnet wire wound in phase over primary winding.*

## Oscillator is only active stage

The third overtone oscillator is the only active stage in this converter. Because I wanted 223 MHz to correspond to 146 MHz, I needed an LO frequency of 77 MHz. The third overtone oscillator operates at one-half the LO frequency. For casual monitoring, use an International Crystal catalog no. 4713601 38.5000-MHz crystal. It will be within several hundred cycles and will work fine. This parallel resonant crystal is cut for a 5-pF load.

For critical applications — like using the converter with a service monitor to set frequencies — I suggest an International Crystal catalog no. 471360 38.500-MHz crystal. This series resonant crystal can be set, or "warped," to exactly 38.5000 MHz by placing a small variable inductor in series with the crystal. See **Figure 1** for details. About ten turns of no. 24 wire on a 3/16 to 1/4-inch slug-tuned form will suffice. Be careful, too much inductance will cause spurious oscillations; some cut-and-try experimentation may be necessary.

## Innovative mixer simplifies circuit

The mixer is driven directly by the 38.5-MHz oscillator. This unique passive mixer stage uses two diodes paralleled back to back (see **Figure 2**). Each diode conducts on alternating crests of the LO injection. This halves the required LO injection and eliminates the need for a frequency doubler. While this isn't the ultimate mixer, it serves my needs. I spent a lot of time optimizing the mixer circuitry, particularly when trying to couple the LO energy into the diodes. The best performance was achieved with the diodes connected across the output link over the oscillator tank circuit; however, losses were still very high. I discovered (entirely by accident!) that by resonating the coupling link to 2 meters the mixer losses dropped by several dB — hence the trimmer capacitor across the link winding. Why this improves performance I honestly don't know, but I surmise the coupling link loads the mixer diodes until they are transformed into

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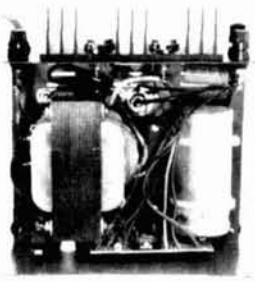

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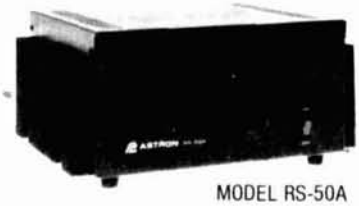

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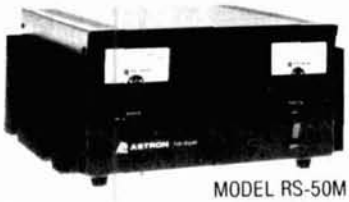

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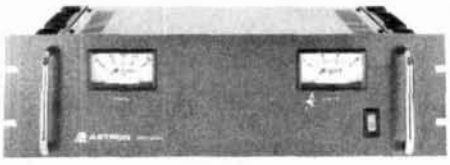


MODEL RM-35M

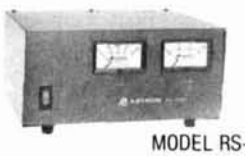
| MODEL | Continuous Duty (Amps) | ICS* (Amps) | Size (IN) H × W × D | Shipping Wt. (lbs.) |
|---|---|---|---|---|
| RM-12A | 9 | 12 | 5¼ × 19 × 8¼ | 16 |
| RM-35A | 25 | 35 | 5¼ × 19 × 12½ | 38 |
| RM-50A | 37 | 50 | 5¼ × 19 × 12½ | 50 |
| • Separate Volt and Amp Meters | | | | |
| RM-12M | 9 | 12 | 5¼ × 19 × 8¼ | 16 |
| RM-35M | 25 | 35 | 5¼ × 19 × 12½ | 38 |
| RM-50M | 37 | 50 | 5¼ × 19 × 12½ | 50 |

## RS-A SERIES

MODEL RS-7A

| MODEL | Continuous Duty (Amps) | ICS* (Amps) | Size (IN) H × W × D | Shipping Wt. (lbs.) |
|---|---|---|---|---|
| RS-3A | 2.5 | 3 | 3 × 4¾ × 5¾ | 4 |
| RS-4A | 3 | 4 | 3¾ × 6½ × 9 | 5 |
| RS-5A | 4 | 5 | 3½ × 6⅛ × 7¼ | 7 |
| RS-7A | 5 | 7 | 3¾ × 6½ × 9 | 9 |
| RS-7B | 5 | 7 | 4 × 7½ × 10¾ | 10 |
| RS-10A | 7.5 | 10 | 4 × 7½ × 10¾ | 11 |
| RS-12A | 9 | 12 | 4½ × 8 × 9 | 13 |
| RS-12B | 9 | 12 | 4 × 7½ × 10¾ | 13 |
| RS-20A | 16 | 20 | 5 × 9 × 10½ | 18 |
| RS-35A | 25 | 35 | 5 × 11 × 11 | 27 |
| RS-50A | 37 | 50 | 6 × 13¾ × 11 | 46 |

## RS-M SERIES



MODEL RS-35M

| MODEL | Continuous Duty (Amps) | ICS* (Amps) | Size (IN) H × W × D | Shipping Wt. (lbs.) |
|---|---|---|---|---|
| • Switchable volt and Amp meter | | | | |
| RS-12M | 9 | 12 | 4½ × 8 × 9 | 13 |
| • Separate volt and Amp meters | | | | |
| RS-20M | 16 | 20 | 5 × 9 × 10½ | 18 |
| RS-35M | 25 | 35 | 5 × 11 × 11 | 27 |
| RS-50M | 37 | 50 | 6 × 13¾ × 11 | 46 |

## 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) @13.8VDC | @10VDC | @5VDC | ICS* (Amps) @13.8V | Size (IN) H × W × D | Shipping Wt. (lbs.) |
|---|---|---|---|---|---|---|
| VS-12M | 9 | 5 | 2 | 12 | 4½ × 8 × 9 | 13 |
| VS-20M | 16 | 9 | 4 | 20 | 5 × 9 × 10½ | 20 |
| VS-35M | 25 | 15 | 7 | 35 | 5 × 11 × 11 | 29 |
| VS-50M | 37 | 22 | 10 | 50 | 6 × 13¾ × 11 | 46 |
| • Variable rack mount power supplies | | | | | | |
| VRM-35M | 25 | 15 | 7 | 35 | 5¼ × 19 × 12½ | 38 |
| VRM-50M | 37 | 22 | 10 | 50 | 5¼ × 19 × 12½ | 50 |

## RS-S SERIES

MODEL RS-12S

• Built in speaker

| MODEL | Continuous Duty (Amps) | ICS* Amps | Size (IN) H × W × D | Shipping Wt. (lbs.) |
|---|---|---|---|---|
| RS-7S | 5 | 7 | 4 × 7½ × 10¾ | 10 |
| RS-10S | 7.5 | 10 | 4 × 7½ × 10¾ | 12 |
| RS-12S | 9 | 12 | 4½ × 8 × 9 | 13 |
| RS-20S | 16 | 20 | 5 × 9 × 10½ | 18 |

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

a high impedance parallel resonant circuit. It's possible that dividing the coupling link into two series parallel resonant circuits, one at 144 MHz and the other at 220 MHz, might further reduce the mixer losses.

I tried both hot carrier and high speed switching diodes, and noted no discernible difference. The HCDs may offer the advantage of better performance with lower LO drive as the battery voltage falls off. Diodes like the 1N914 or 1N4148 work well in this mixer.

## Mixer limitations

The mixer has 12 dB of insertion loss. This is a little high and is a fair compromise between cost and circuit simplicity. Despite the losses, I achieved a respectable 0.5-$\mu$V sensitivity with the ICOM 2-AT. Repeaters within 20 miles of my QTH are full quieting with only a rubber duck antenna. With just one tuned stage used at each mixer port, unwanted converter image responses and high levels of LO and related harmonic energy leakage may occur. The converter offers about 20-dB attenuation port to port. Thus, for example, a strong signal on 146.82 MHz may compete with a repeater on 223.82 MHz.

## Where to find the parts

I made every effort to use common parts. Many of the components can be found at the local Radio Shack. Most of the other parts (trimmer caps, toroids, FETs, etc.) can be obtained via mail order. I recommend you try Radiokit or Circuit Specialists. I used handwound coils and capacitive tuning, and avoided the high cost of slug-tuned inductors, which are hard to find anyway. I also used inexpensive 35-pF plastic trimmer capacitors. Radio Shack currently offers 50-pF trimmers. You can use these in a pinch, but you may find the tuning overly sharp. The airwound coils were made from no. 22 gauge hookup wire stripped of its insulation and wound over a pencil. The wire gauge isn't especially critical; neither are the coil dimensions — within reason.

## Alignment

Oscillator tuning is performed by monitoring the current drawn by the converter. When oscillating, the current will drop sharply — typically from 8 mA to about 2 to 4 mA. The trick here is to trade maximum oscillator output for reliable oscillator startup when power is removed and reapplied. The oscillator should maintain oscillation and start reliably with voltages as low as 6 volts. If you use it, set the warping inductor to minimum inductance (slug out) during testing. You may use a frequency counter to set the oscillator to exactly 38.500 MHz, or you can simply adjust for minimum distortion on monitored signals. The oscillator's fourth harmonic can also be zero beat to 144.000 MHz on an SSB transceiver.

Tune the 144 and 220-MHz tuned stages for best sensitivity. You can use a signal generator or strong local 220-MHz signal. If a peak occurs at minimum or maximum capacity, try spreading or compressing the coil windings to compensate. Finally, you can tweak the oscillator tank slightly for best sensitivity. You may have to retweak the input circuit a bit if you change antennas. Use a 220-MHz antenna for best sensitivity.

I use a 9-volt alkaline battery for power. The battery will last a long time with only a 2 or 3-mA draw. For continuous monitoring use a wall plug-type power supply. The converter is forgiving of accidental transmitter keying, but it is not burnout proof. Always set the handheld to low power before connecting the converter.

## Using the converter with a service monitor

Need 220-MHz coverage on your service monitor? This converter makes that possible. First, calibrate the converter. Set the preselector and CE-3 to monitor 144.000 MHz. Connect the 2-meter port to the preselector input and adjust the oscillator frequency to exacly 144.000 MHz. This sets the converter to the limits of the CE-3's own calibration.

To monitor a 220-MHz signal, connect the converter 2-meter port to the preselector input. Set the Cushman to the 2-meter frequency corresponding to the 220-MHz frequency to be monitored. You can now make very accurate 220-MHz frequency and deviation checks. Connecting the 2-meter port to the signal generator output will produce 220-MHz signals. Although the generator calibration won't be very accurate because the conversion losses will affect the readings, you can still perform meaningful relative comparisons and receiver peaking. hr

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A well-stocked junkbox is the best source of parts. I suggest writing the above companies for current catalogs; prices, minimum orders, and stock vary continually.

# DIGITAL SIGNAL PROCESSING
## PART 1: THE FUNDAMENTALS

*By Bryan P. Bergeron, NU1N, 30 Gardner Road, Apt. 1G, Brookline, Massachusetts 02146*

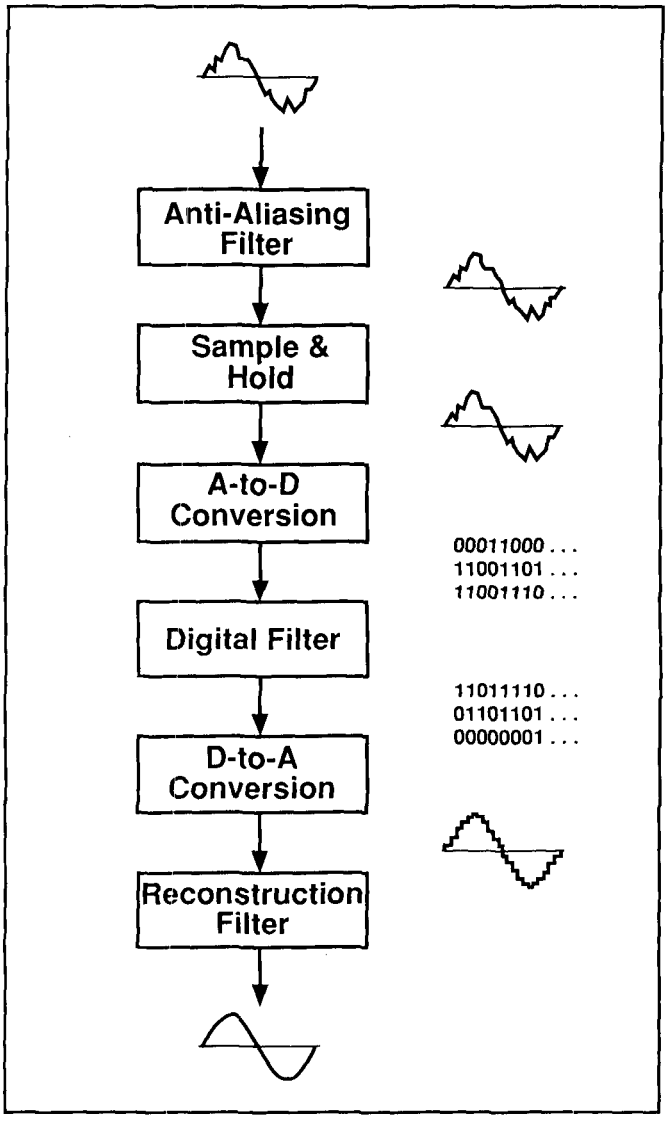
The introduction of transistors and other solid-state devices over the past 20 years has changed the way most of us think about and work with electronic circuitry. As radical as the change from tube-type to solid-state devices and circuits seemed, the current progress in digital signal processing (DSP) technology suggests that an even greater revolution is at hand. Signal processing itself isn't new to electronics or communications — it was used extensively during World War II to remove the noise and clutter from radar signals.[1] Most early signal processing was done in analog form and was of limited value. The introduction of specialized DSP chips has kindled a new interest in signal processing using powerful *digital* techniques. Just as the digital compact disk (which relies on DSP techniques) has redefined the standards in the field of music recording and reproduction, DSP technology promises to drastically enhance the power and flexibility of Amateur and commercial communications systems. This article, the first in a series, examines the fundamentals of digital signal processing and provides experiments in DSP techniques that you can perform on your home computer.

### DSP fundamentals

To compare and contrast DSP and classic circuit design techniques, I'll approach a simple problem from both perspectives. For discussion's sake, assume that a low frequency audio signal is accompanied by a significant amount of high frequency noise — for example, a 1000-Hz CW signal accompanied by considerable QRN, most of which is higher than 1000 Hz in frequency at the receiver audio stage. The task is to provide low pass filtering to minimize the amplitude of the high frequency components of the offending noise, without appreciably degrading the desired CW signal.

The standard approach to this problem is to design either an RC or RL low pass filter comprised of discrete components. The DSP approach is considerably more involved and requires several components including an analog antialiasing filter, an analog to digital (A/D) converter, a digital filter and, in the form of hardware and computer instruc-

FIGURE 1




**A digital signal processing system configured to provide low-pass filtering of an analog signal.**

tions, a digital to analog (D/A) converter and an analog reconstruction filter (see **Figure 1**).

## Antialiasing filter

The analog antialiasing filter removes extraneous noise or other higher frequency signals that, when sampled by A/D conversion circuitry, could appear as false low frequency signals. This effect of folding higher frequencies down to lower frequencies is often called aliasing, hence the name antialiasing filter.

## A/D converter

The A/D converter can be thought of as a combination sampler, quantizer, and encoder. The A/D converter samples the analog input signal at evenly spaced intervals of time, and produces an accurate, digital representation of the instantaneous signal amplitude.

Of the wide variety of techniques available for analog-to-digital signal conversion, the most common include parallel encoding, successive approximation, voltage-to-frequency conversion, and single slope integration.[2] Each of these basic techniques has its own tradeoffs in terms of relative accuracy, speed, and expense. A/D converters vary in sampling rate, quantization noise, dynamic range, and digitization technique used.

The Nyquist sampling theorem states that the minimum sampling frequency is at least twice that of the signal to be sampled. The sampling rate effectively limits the upper frequency range of signals that can be digitized accurately. To adequately digitize a 50-kHz analog signal, the digitizer should sample the input waveform at least 100,000 times/second. Sampling at a lower rate results in aliasing, a significant source of noise in A/D conversion. One way to avoid aliasing is to use a low pass (antialiasing) filter that cuts out frequencies greater than half of the sampling rate.

Both the quantization noise and the dynamic range of an A/D converter are related to how the digital data is represented. Eight-bit digitizers, which represent amplitude variations with integers from 0 to 255, generally have more quantization noise than 12 or 16-bit digitizers. Quantization noise can be thought of as the high frequency noise introduced into a digitized waveform due to the "staircase effect" inherent in the A/D process. For example, even though the amplitude of an analog signal smoothly varies from 2 to 3 volts, an eight-bit digital representation of the signal might show sudden jumps from 2.0 to 2.5 to 3.0 volts. A 16-bit digitizer, on the other hand, might record less drastic changes. The less jagged representation provided by the 16-bit digitizer should result in less associated quantization noise. You can also think of quantization noise as the difference between the true value and the actual value of a waveform that would be observed on an oscilloscope if the digital representation were immediately applied to a D/A converter.

The digital representation of signal amplitude defines the maximum dynamic range (the difference between the largest and smallest signal that can be recorded) of a digitized signal. The dynamic range of an A/D converter, equivalent to the signal-to-noise ratio of a converter, is related to the number of bits used to represent a waveform by the following equation:[3]

$$SNR = 6.02\ n + \log_{10}(1.5)\ dB$$

where n is the number of bits used in the digital representation of the sinusoidal waveform, not including the sign bit. That is, an eight-bit (N = 8; n = 7) A/D converter uses one bit for sign (+ or −) and seven bits (n = 7) for magnitude. For example, with N = 8, 12, 16, and 32, the SNR = 42, 66, 90, and 187 dB, respectively.

## Digital filter

The digital filter (in actuality a computer program or algorithm) works with the digitized signal data to produce the desired output, which is also in digital form. Unlike a conventional system based on discrete components, all waveform manipulations in a digital filter are implemented in software. In many respects, this operation is not unlike everyday data processing. It just so happens that the data represents signal amplitude levels, as opposed to financial account balances or the current value of stocks and bonds. The other qualification is that DSP deals with *real time* analog signals which are processed digitally.

Because of the computational demands of digital filtering, systems based on general purpose microcomputers typically are limited to applications where the signals have bandwidths of only a few hundred hertz. Specialized DSP chips, which can be interfaced to a microcomputer system, extend the working bandwidth by several orders of magnitude. I'll discuss a working example of a microcomputer-based algorithm that implements digital low pass filtering for signals in the audio frequency range later in this article.

## D/A converter

The D/A converter performs the inverse function of the A/D converter described earlier. The output of the digital filter is converted by the D/A conversion circuitry into an analog signal. Like A/D converters, D/A converters suffer from the staircase effect, in that a continuous output signal is represented as abrupt changes in amplitude.

## Reconstruction filter

Like the antialiasing filter, the reconstruction filter is typically an analog low pass filter composed of discrete components. Its purpose is to remove the high frequency staircase artifact introduced by the D/A conversion process.

## Advantages of DSP

Given the above signal processing environment, the digital filter designer would still be faced with the problem of specifying the sample rate for the A/D and D/A converters, selecting the implementation hardware, and creating or locating the appropriate algorithm for the desired characteristics of the digital filter. The designer also has to work around the memory and computational limits of the hardware platform.

With these demands, why would anyone ever consider using DSP techniques instead of conventional methods of circuit design? The major reason is flexibility, especially when complex filtering operations are considered. For simple problems, the DSP solution is much more complex than the conventional alternative. However, this distinction fades when you consider complex filtering involving multiple inputs. While the DSP system would be able to perform all needed operations without a change in physical configuration, a conventional analog system would probably require numerous discrete components for each filtering

operation. A DSP system can be transformed into a specific tool with changes only in programming.

Digital signal processing has additional advantages over analog systems developed with discrete components. DSP techniques are stable, repeatable, and predictable. DSP systems don't degrade with time, suffer from temperature drift, or change operating characteristics with age. The mathematical relationships and binary operations within a digital system remain constant, making the outcome completely predictable and reproducible from one DSP system to another.

Using DSP, the conventional hardware design task is transformed into a programming problem. Once you know the algorithms, it's relatively straightforward to get a process working. The bulk of the development cycle is typically spent improving the speed and performance of the algorithms. Many operations that are difficult or nearly impossible to achieve with conventional analog processing techniques, like linear phase filtering, are almost trivial with today's DSP techniques. DSP promises to minimize the design and testing time inherent in circuit design and maximize the designer's flexibility when solving particular problems.

## DSP limitations

DSP techniques are not without their limitations. In addition to quantization noise and limited dynamic range, DSP suffers from round-off and truncation errors, along with limited bandwidth.

## Round-off and truncation errors

All digital computations, especially those that use multiple operations on floating point numbers (3.1415926535898 or $\pi$, for example) are prone to errors. Because of the way in which the two components of a floating point number (the mantissa and the exponent) are handled, computations involving numbers in floating point format aren't exact. Even when hardware is dedicated to floating point operations, the cumulative results of multiple mathematical manipulations will become significant, given enough iterations.

*Round-off errors*, like quantization noise and limited dynamic range, are due to computer hardware limitations. They can be minimized by the judicious use of appropriate data types and algorithms. Another type of error, independent of the computer hardware, can be attributed to the algorithms used. *Truncation errors* occur when algorithms use approximations to arrive at an answer. Instead of computing the sum of an infinite series, a practical algorithm might stop after a sufficient number of elements in the series have been added. Truncation error can best be thought of as the difference between the actual answer and the answer obtained via a practical calculation. Unlike round-off errors, which are inherent to the hardware platform, truncation errors can be minimized by selecting the appropriate algorithms for a given problem.

## Limited bandwidth

Although there are DSP systems which operate in real time on microwave signals (massively parallel over-the-horizon radar receiving installations[4]), most of us will have to contend with the limited bandwidth of current commercial DSP devices.

The best of today's generally available DSP chips use clock frequencies of up to 42 MHz to realize cycle times of as low as 50 nanoseconds.[3] Depending on how many signals must be handled concurrently, the usable bandwidth of these devices approaches only a few hundred kHz.

## DSP hardware

For the past several decades, the use of digital signal processing has been limited by the inherent limitations of available computer hardware. Although digital signal processing can be done with a general purpose computer system, only the relatively recent introduction of special purpose digital hardware has moved DSP from the research laboratory into the hands of engineers and developers. The introduction of VLSI chips that can perform signal processing algorithms in parallel, with many chips operating simultaneously, makes it possible to perform many signal processing tasks up to a thousand times faster than present computers — including supercomputers.[1]

## Dedicated integrated circuits

The first generation of dedicated DSP devices, like the INTEL 2920 introduced in 1979, had on-chip scratch pad (RAM) and program memory (EPROM), sample and hold circuitry, and a high speed digital processor, along with D/A and A/D converters.[5] With a 10-MHz clock, this 40-pin DIP chip was fast enough to perform real time processing on analog signals. It was used extensively to implement frequency and phase filters, as well as peak detectors.

Since the introduction of the INTEL 2920, there have been two subsequent generations of dedicated DSP devices. Examples of later generation devices are Texas Instruments' TMS320, Analog Devices' ADSP2100, Motorola's DSP5600, NEC's MPD7720, Fujitsu's MB8764, and AT&T's WE DSP32. These later devices typically provide much more program memory than the 2920, allowing for more conventional programming techniques. That is, some devices provide enough memory to permit programmers to work in high level languages — like C, as opposed to ASSEMBLER.

Compared to the INTEL 2920, which was limited by its on-board A/D and D/A converters, these newer devices have been used to implement a wide variety of digital filters for a variety of applications. The TMS320 has been a popular building block of portable spectrum analyzers.[6] All current DSP chips use a single 5-volt supply, and are designed to interface directly with TTL devices.

## Device architectures

DSP chips are designed with one primary consideration: speed of execution. Unlike the serial nature of the general purpose microcomputer, DSP devices use parallel architectures and pipelining to increase execution speed. Unlike the typical microcomputer which handles one instruction at a time, DSP devices perform multiple operations simultaneously. A complex problem that might take 20 instruction cycles to solve in a serial device might take only six instruction cycles in a parallel device.

Most DSP chips make use of the Harvard Architecture, where the program memory is separate from data memory.[3] This allows data to be fetched concurrently with program instructions. The Harvard Architecture contrasts sharply with the architecture of conventional computer systems, in which only a data value *or* a program instruction can be moved from one memory location to another at a time.

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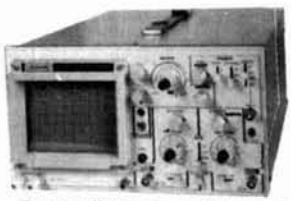

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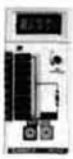

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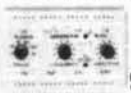

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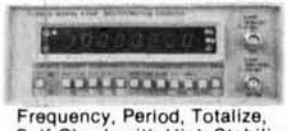

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In addition to separate data and program memory, DSP chips have a number of dedicated processing units that can work in a highly coordinated fashion. One or more arithmetic logic units (ALUs) handle addition, subtraction, logical operations, and address generation. The shifter scales data to or from other processor units, while the multiplier (MAC) performs multiplications.

## DSP chip instruction sets

The instruction sets (an instruction set is the native language of a microprocessor chip) supported by specialized DSP chips are in many respects comparable to those supported by the multipurpose CPUs used in popular microcomputers. DSP devices regularly support an impressive assortment of arithmetic and logic operations, data and program control, condition testing, and system control instructions specifically designed to support signal acquisition and processing.

## DSP software

Every DSP program has two components: one that deals with signal acquisition, and one that's concerned with the actual processing. Of the two components, signal acquisition is usually the most straightforward, perhaps because the alternatives are limited. Designers are typically constrained by the resolution, sampling rate, and dynamic range associated with a particular A/D converter, even in a system designed around a general purpose microcomputer. On the other hand, designers have considerable flexibility in the design of the actual signal processing software.

Perhaps the most significant software design issue, especially in a microcomputer-based DSP system, is whether to work in the frequency or time domain — that is, whether to represent signal amplitude as a function of time or frequency. This decision is akin to deciding between using an oscilloscope or a spectrum analyzer as a diagnostic tool; it depends on your experience with the tools in question and on the problem at hand. Traditionally, digital signal processing in the time domain has been avoided in favor of the more powerful processing available when working in the complex frequency domain. Although there's increased activity in investigating time domain signal processing,[9] frequency domain work continues to be the most popular method of signal processing.

Even so, most of us are more comfortable working in the time domain. After all, recording events relative to the passage of time is part of our everyday lives. Many of our test instruments, including the common oscilloscope, assume that time is the point of reference. Electronic signals are commonly expressed as functions of time. Time usually serves as the point of reference in the analysis of common circuits, like the charging of a capacitor. Were you to view the charging of a capacitor on an oscilloscope screen, you'd see a familiar voltage curve increasing exponentially over time.

Anyone who has worked with a spectrum analyzer appreciates the unique perspective it provides. A conventional time-domain oscilloscope simply isn't useful in assessing quantitatively the spectral quality or bandwidth of a signal. Similarly, there are operations in signal processing that just can't be performed easily in the time domain. For example, consider the problem of constructing a linear phase filter — a low pass filter that has linear phase characteristics over the operating frequency range of the filter. The very concept of phase linearity is foreign to the design of high and low pass filters in the time domain, where we work with magnitude characteristics without regard to phase.

By working in the frequency domain with mathematical techniques like the Fourier Transform, a filter designer can manipulate the phase characteristics of a filter easily. Just as a spectrum analyzer is easy to understand and use, designing filters and other circuits in the frequency domain can become second nature once you understand the nature of the device and the mathematical tools involved.

## DSP with your personal computer

For signals in the audio spectrum, the capabilities of even the most modest personal computer are sufficient to demonstrate basic DSP techniques. Some microcomputer systems, like the Commodore 64, have hardware dedicated specifically to DSP tasks. I'll describe the DSP capabilities of the C-64 hardware in more detail in the next section, along with two BASIC programs that demonstrate the capabilities of this class of hardware environment.

## Audio frequency DSP with the C-64

The 6581 Sound Interface Device (SID) in the C-64 can be used as a simple DSP system. While the 6581 supports several tone oscillators, amplitude modulators, envelope generators, and A/D conversion, the most important characteristic from the perspective of DSP is its programmable filter. The digital filter section of the 6581 can provide high, low, bandpass, and notch filtering, as well as variable resonance and volume control over a range of 30 Hz to 12 kHz.[7] The rolloff is about 12 dB/octave.

The SID chip supports three separate software-definable filters whose outputs can be chained to produce a variety of effects. For example, both low pass and high pass filters can be selected to produce a notch filter response. The volume can be varied from no output to maximum volume in 16 linear steps. Because the SID chip provides no amplification, maximum volume represents no attenuation of the input signal.

The external analog audio input, applied to pin 5 of the C-64's audio/video socket (see **Figure 2**), shouldn't exceed 3 volts peak to peak. The input impedance is on the order of 100 k. The external signal can ride a DC voltage, if necessary, because the external input is AC-coupled to the SID chip through a 1-$\mu$F electrolytic capacitor. The audio output of the SID chip, directed to pin 3 of the audio/video socket shown in **Figure 2**, has an impedance of approximately 1 k. The output level, set by the output volume control on the SID chip, reaches a maximum of 2 volts peak to peak. Like the audio input, the audio output of the SID chip is AC-coupled to the audio/video socket through a 1-$\mu$F electrolytic capacitor.

**Table 1** describes the registers and SID chip addresses pertinent to the discussion of DSP. SID offset addresses 21 and 22 define the cutoff frequency used by the filters. Offset address 23 controls both the filter resonance and voice input controls; offset address 24 selects the filter mode and volume.

To better understand how the SID chip can be manipulated for DSP purposes, see the Commodore BASIC program in **Table 2**. This short routine provides an example of digital low pass filtering, with a cutoff frequency of 1000 Hz. The first step is to define the starting address, S or 54272,

**TABLE 1**

| Register Address | Bits | Function |
|---|---|---|
| S + 21 | □□□□□■■■ | Filter cutoff frequency: low-nibble |
| S + 22 | ■■■■■■■■ | Filter cutoff frequency: high-byte |
| S + 23 | ■■■■□□□□<br>□□□□■□□□ | Filter resonance/voice input control<br>Filter resonance: 0 - 15<br>Filter external input: 1 = Yes; 0 = No |
| S + 24 | □■□□□□□□<br>□□■□□□□□<br>□□□■□□□□<br>□□□□■■■■ | Filter mode and volume<br>High-pass mode: 1 = On<br>Band-pass mode: 1 = On<br>Low-pass mode: 1 = On<br>Output volume: 0 - 15 |

**Description of the registers and SID chip addresses pertinent to the discussion of DSP.**

of the SID device (line 110) and then to clear the SID registers (lines 140 to 160). Next, define the cutoff frequency (lines 190 to 200). The decimal values to be POKEd into memory locations S+21 and S+22 can be determined by the following formula:

$$S+21 = Frequency_{Hz} \; MOD \; 256$$
$$S+22 = Frequency_{Hz} - \{(S+21) \times 256\}$$

For example, with a cutoff frequency of 1000 Hz:

$$S+21 = 1000 \; MOD \; 256 = 3$$
$$S+22 = 1000 - \{(3) \times 256\} = 1000 - 768 = 232$$

The Commodore, like many other microcomputers, stores multiple part numbers with the low part placed in the first memory location and the high part next. You should note that, because the maximum value that can be POKEd into the low bits of filter frequency cutoff (S+21) is 7 or 111 binary,[8] the cutoff frequency range is from 0 Hz to ((7 ×256) + 255), or 2047 Hz.

Now that you've initialized the SID chip and defined the cutoff frequency, you need to route the external audio input to the filter circuitry (line 220). Line 250 defines the resonance characteristics of the filter. In this example 7, a moderate value of resonance, is selected. To vary the amount of resonance provided by the filter, use the following formula:

$$POKE \; S + 24, \; PEEK \; ((S + 24) + (R \times 16))$$

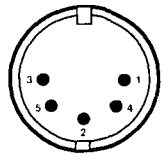
where R is equal to the amount of resonance desired from 0 to 15 (0000 to 1111 binary).

Next, you need to define the output volume of the filter (line 280). Like resonance, output volume can be varied from 0 to a maximum (as in this example) of 15. Remember that maximum volume is really minimum attenuation, as the C-64 doesn't provide amplification of the external audio signal.

Finally, you can define the type of filtering provided by the SID chip — in this case low pass filtering (Line 310). This filter will pass any frequency lower than the cutoff frequency set in addresses S+21 and S+22 without attenuation. See **Figure 3** for a summary of this and other filter mode definitions that can be substituted for line 310 of **Table 2**.

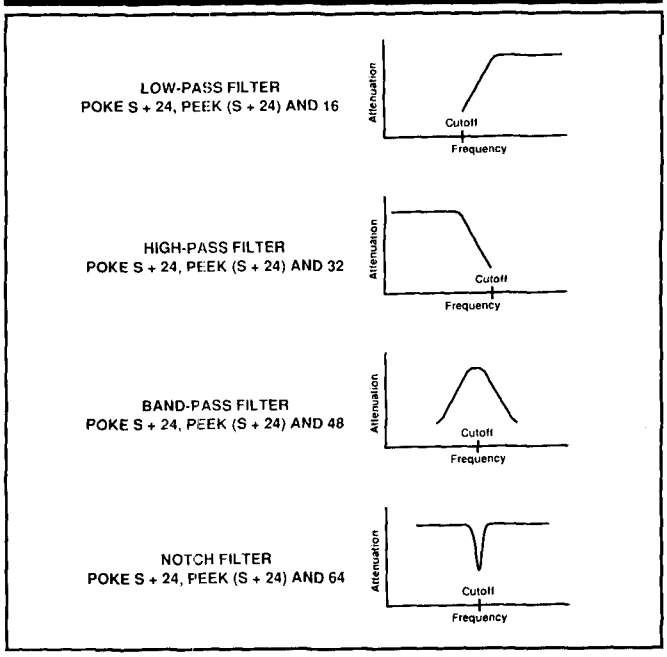
**FIGURE 2**

| Pin | Signal |
|---|---|
| 1 | Luminance |
| 2 | Ground |
| 3 | Audio Out |
| 4 | Video Out |
| 5 | Audio In |




**The audio/video port pinouts for the Commodore-64. Pins 2, 3, and 5 are used in the audio frequency DSP program described in the text.**

**FIGURE 3**




**Filter mode definitions that can be used with the Commodore-64's SID chip.**

**Table 3** provides a more interactive DSP environment on the C-64. A standard game paddle can be used to vary the cutoff frequency of a notch filter in real time, from 0 to 2047 Hz. Line 110 defines the memory address for the start of the paddle routine. Because the paddles are unreliable when read from BASIC alone, a machine language routine (lines 340 to 400) is POKEd into memory (line 130). Next, as in **Table 2**, the SID chip address is defined (line 150) and the chip is reset (line 170). The filter is set up to handle an external audio signal (line 190), filter resonance is set to medium (line 210), and filter volume is set to maximum (line 230). Next, the filter mode is set to notch (line 250).

Because the machine language routine to read the paddle values (lines 340 to 400) returns a value between 0 and 255 when called (line 280), the value returned is multiplied by 8 to cover most of the frequency range supported by the filter, 0 Hz to 2047 Hz. For example, when the paddle value is read as 255, the value variable (V) is set to 8 × 255 or 2040. Because the cutoff frequency must be POKEd into the appropriate memory locations as separate high and low order bits, the value variable is decomposed

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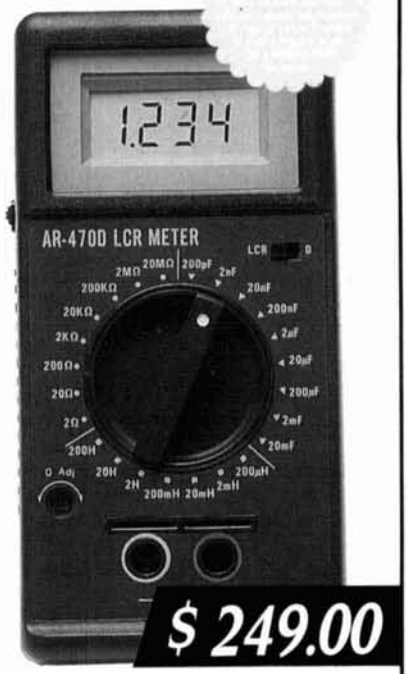


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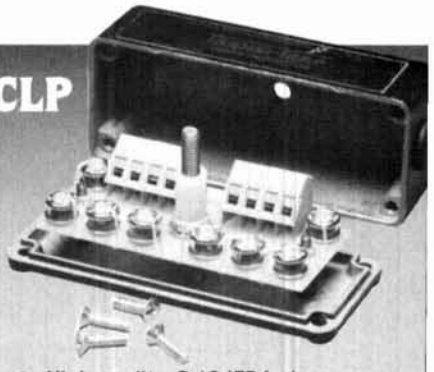

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| UG21DS N plug for RG8,213,214 Silver | 3.3 |
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into V1, the low order three bits, and V2, the high order eight bits (line 290). The high and low order bit values are then POKEd into their corresponding cutoff frequency memory locations (line 300). The endless loop structure set up by line 310 causes the cutoff frequency value to be updated continuously, therefore following the instantaneous game paddle position.

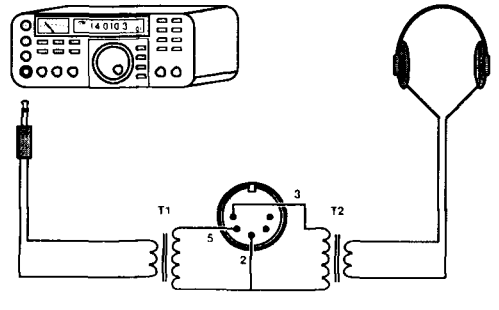
Equipped with the above programs, you can experiment with a variety of audio input and output hardware devices. For example, if you take moderate care to match the input and output impedances of the C-64 hardware with your communications gear, you can experiment with digitally processed CW signals. I found that a pair of common audio transformers, like the popular 1000 to 8-ohm variety (Radio Shack no. 273-1380), are sufficient to couple receiver audio into the C-64 filter system and out to a pair of low impedance headphones (see **Figure 4**). Alternatively, you can experiment with an audio signal generator and your home stereo system or an audio amplifier.

TABLE 2

```
100  REM DEFINE SID CHIP SOUND ADDRESS
110  S = 54272
120  REM
130  REM CLEAR SID CHIP REGISTERS
140  FOR I = 0 TO 28
150  POKE S + I, 0
160  NEXT I
170  REM
180  REM SET CUTOFF FREQ TO 1KHZ
190  POKE S + 21, 3
200  POKE S + 22, 232
210  REM FILTER EXTERNAL INPUT
220  POKE S + 23, PEEK (S + 23) OR 16
230  REM
240  REM SET RESONANCE TO MEDIUM
250  POKE S + 23, PEEK (S + 23) + (7 * 16)
260  REM
270  REM SET VOLUME TO MAXIMUM
280  POKE S + 24, 15
290  REM
300  REM SET FILTER MODE TO LOW PASS
310  POKE S + 24, PEEK (S + 24) AND 16
```

**A BASIC listing for the Commodore-64 that illustrates audio-frequency DSP techniques. This program defines a low-pass digital filter with a cutoff frequency of 1 kHz.**

FIGURE 4




**A suggested configuration for experimenting with DSP and real-time CW signals. Audio output from a receiver or transceiver is coupled to the relatively high impedance external audio input (pin 5) port of the Commodore-64, through an impedance matching transformer (T1). The digitally filtered audio output (pin 3) is coupled to low impedance headphones through a second impedance matching transformer (T2).**

TABLE 3

```
100  REM DEFINE START OF PADDLE ROUTINE
110  C = 12 * 4096
120  REM READ IN PADDLE ML DATA
130  FOR I = 0 TO 63: READ A: POKE C + I, A: NEXT I
140  REM DEFINE SID CHIP SOUND ADDRESS
150  S = 54272
160  REM CLEAR SID CHIP REGISTERS
170  FOR I = 0 TO 28: POKE S + I, 0: NEXT I
180  REM FILTER EXTERNAL INPUT
190  POKE S + 23, PEEK (S + 23) OR 16
200  REM SET RESONANCE TO MEDIUM
210  POKE S + 23, PEEK (S + 23) + (7 * 16)
220  REM SET VOLUME TO MAXIMUM
230  POKE S + 24, 15
240  REM SET FILTER MODE TO NOTCH
250  POKE S + 24, PEEK (S + 24) AND 64
260  REM
270  REM SET CUTOFF FREQ
280  SYSC: V = 8 * PEEK (C + 257)
290  V1 = INT (V/256): V2 = V - V1
300  POKE S + 21, V1: POKE S + 22, V2
310  GOTO 280
320  REM
330  REM MACHINE LANGUAGE DATA
340  DATA 162, 1, 120, 173, 2, 220, 141, 0, 193
350  DATA 169, 192, 141, 2, 220, 169, 128, 141
360  DATA 0, 220, 160, 128, 234, 136, 16, 252
370  DATA 173, 25, 212, 157, 1, 193, 173, 26
380  DATA 212, 157, 3, 193, 173, 0, 220, 9, 128
390  DATA 141, 5, 193, 169, 64, 202, 16, 222
400  DATA 173, 1, 220, 141, 6, 193, 88, 96
```

**A BASIC listing for Commodore-64 that provides an interactive environment for investigating audio DSP techniques. The cutoff (center) frequency of the digital notch filter can be controlled, in real time, by the position of a game paddle attached to the game port of the Commodore-64.**

**Figure 4** shows a suggested configuration for experimenting with DSP and real-time CW signals. Audio output from a receiver or transceiver is coupled to the relatively high impedance external audio output (pin 5) port of the Commodore-64 through an impedance matching transformer (T1). The digitally filtered audio output (pin 3) is coupled to low impedance headphones through a second impedance matching transformer (T2).

## Time domain DSP with generic PC hardware

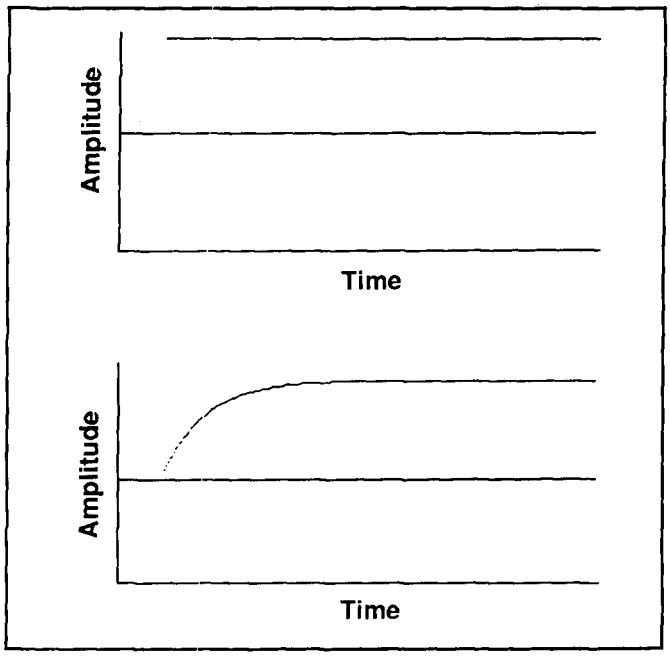
Experimentation with microcomputer-based DSP techniques needn't be limited to computer systems that come factory equipped with DSP hardware, like the Commodore 64, especially if real time processing isn't a necessity. Assuming that there's some way of moving real or simulated signal data into and out of working memory by means of D/A and A/D converter cards or simple data statements written in BASIC, then DSP is simply a matter of programming. (A variety of digital I/O boards and periperals are available for the popular microcomputer systems through a number of sources including MetroByte Corporation, 440 Myles Standish Boulevard, Taunton, Massachusetts 02780.) Although the programs will typically be much more involved than the example listings I've provided for the C-64, they can also be much more powerful. You are no longer constrained to make do with the algorithms provided

TABLE 4

```
100  REM DEFINE VARIABLES, IN/OUT ARRAYS
110  LAST = 0
120  COUNT = 30
130  DIM INPT[COUNT]
140  DIM OUTPT[COUNT]
150  REM
160  REM READ INPUT SIGNAL INTO INPT ARRAY
170  FOR I = 1 TO COUNT: READ INPT[I]: NEXT I
180  REM
190  REM DEFINE FILTER TIME CONSTANT
200  INPUT "Cutoff Frequency (1-100)?", CUTOFF
210  IF CUTOFF < 1 OR CUTOFF > 100 THEN 170
220  X = CUTOFF * 628 / 256
230  ALPHA = (X * X * 35 - 9644 * X + 1000000) / 1000
240  REM
250  REM FILTER INPUT SIGNAL
260  FOR I = 1 TO COUNT
270  TEMP = (1024-ALPHA) * (INPT[I] * 1024)
280  TEMP = (TEMP + (ALPHA * LAST)) / 1024
290  LAST = TEMP
300  OUTPT[I] = INT(TEMP / 1024)
310  NEXT I
320  REM
330  REM PRINT FILTER OUTPUT
340  CLS
350  FOR I = 1 TO COUNT: PRINT OUTPT[I]: NEXT I
360  REM
370  REM STEP SIGNAL OF AMPLITUDE 20
380  DATA 0, 0, 0, 20, 20, 20, 20, 20, 20, 20
390  DATA 20, 20, 20, 20, 20, 20, 20, 20, 20, 20
400  DATA 20, 20, 20, 20, 20, 20, 20, 20, 20, 20
```

**A BASICA listing for the IBM-PC and clones that illustrates time-domain DSP with generic computer hardware. The program simulates a first order recursive low-pass filter, with a user-definable cutoff frequency.**

FIGURE 5




**Input to (top) and output from (bottom) the first-order recursive low-pass filter defined in *Table 4* (bottom). The step function, defined in the DATA statements of *Table 4*, was sent through the filter with the cutoff frequency set to 1 Hz.**

in ROM, but are free to develop your own digital signal processing algorithms.

The program in **Table 4**, developed in BASICA for the IBM PC, provides for low pass filtering in the time domain (frequency domain filtering will be considered in a future article). For those who don't have a digital I/O board for their PC, the input data for the filter is supplied by data statements within the program. Remember that in an actual digital processing system, the input and output would be directed, through the I/O board, to a buffer in preparation for processing.

This program simulates a first order *recursive* (IIR) low pass filter, with a user-definable cutoff frequency. The filter is said to be recursive because at any given moment the output of the filter is a function of the current and previous inputs and previous outputs.[10] Recursive filters, while generally superior in performance to nonrecursive filters (filters which don't feed on their own output), suffer from problems associated with instability.[11] While nonrecursive filters are always stable, recursive filters may not be, so the output of a poorly designed recursive filter may increase exponentially — even though the input has ceased. The challenge associated with designing recursive filters comes in selecting filter coefficients that minimize instability while providing the required filter response.

The underlying equation defining the operation of the low pass filter in **Table 4** is:

$$Output_n = \{(K - å) \times (Input_n \times K) + (å \times Output_{n-1})\} / K$$

Where å is the low pass filter time constant, and the coefficient K is taken to be $2^{10}$, or 1024. As you can see, the output of this recursive filter is due in part to the previous filter output value ($Output_{n-1}$).

In this example, å is related to the filter cutoff frequency by the following equation:

$$å = \{(Cutoff \times K_1)^2 \times K_2 - K_3 \times (Cutoff \times K_1) + K_4\} / K_5$$

where the coefficients K1, K2, K3, K4, and K5 are defined as 2.45, 35, 9644, 1000000, and 1000, respectively, and Cutoff is the cutoff frequency in hertz. For an excellent introduction to how these coefficients are selected, see the texts by Hamming[12] and Rice.[13]

Given the preceding relationships, it's rather easy to define the filter in a BASIC program. As shown in **Table 4**, the first step is to declare the variables and arrays used in the program (lines 110 to 140). The variable LAST holds the previous output of the filter; that is, $Output_{n-1}$. COUNT is the number of data points in the input signal. In line 170, the input data is read into the input data array, INPT. In this example, a step signal of amplitude 20 is used as the input to the filter (lines 380 to 400). Remember that in an actual DSP application, the input data would be routed in real time to the filter input.

Lines 200 to 230 define the filter constant, å in terms of the desired cutoff frequency. Notice that the cutoff frequency range has been limited to between 1 and 100 Hz (line 210). This range will be more than sufficient to examine the effect of digital low pass filtering on the input signal provided. (Note that the cutoff and signal frequencies are relative.)

The actual filter processing takes place in lines 260 to 310. For each input data element, an output value is calculated, based on the filter constant and the previous output value.

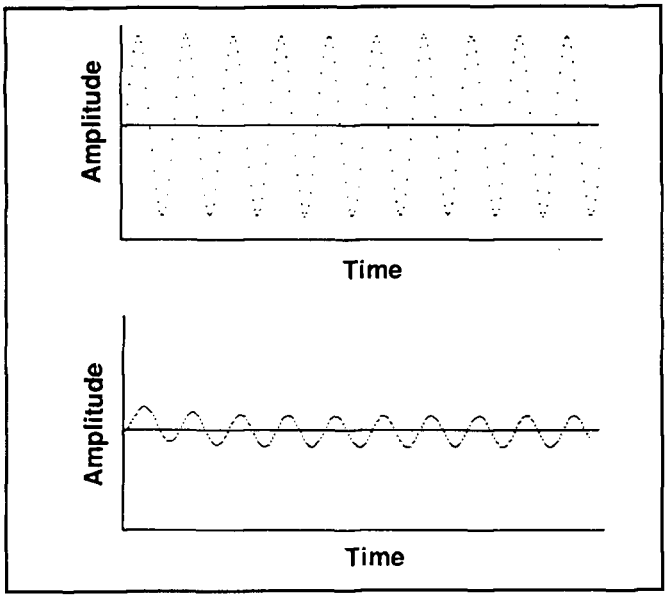
The output of the filter is saved into the LAST variable for the next filter cycle (line 290). In line 300, the filter output is both scaled and converted from a real to an integer value for a more esthetic output. Line 340 clears the screen in preparation for a listing of the filter output values (line 350).

In place of the simple listing of filter output, you can substitute a graphic plotting routine to get a better grasp of the relationship between cutoff frequency and filter output over time. The lower half of **Figure 5** shows the plotted output of the program in **Table 4**, with a cutoff frequency of 1 Hz. To make the plot, the data statement was extended to contain 70 additional input signal points, and the count variable adjusted to 100 accordingly. The filter output was then sent to an Apple Macintosh plotting program for the graphical representation.

The plots shown in **Figure 6** were made in a similar manner by substituting sinusoidal data for the step data in lines 380 to 400 of the program. The input signal (**Figure 6**, top) has a frequency of 10 Hz. Notice the decreased amplitude of the output signal after processing by the digital low pass filter, which has a cutoff frequency of 1 Hz. Upon careful examination of the start of the output waveform (**Figure 6**, bottom), you'll notice that the baseline of the first sine wave is abnormally elevated. The subsequent output waveforms, however, seem to be centered squarely on the baseline.

This aberration is due to the assumptions made during initialization of the LAST variable, corresponding to $Output_0$ in the program. Although an initial value of 0 for $Output_0$ was appropriate for the step function, it wasn't optimum for the sinusoidal signal. If we know a priori about the nature of the input signal (that it always starts with a positive-going sine wave, for instance), then $Output_0$ can be initialized to an appropriate (that is, negative) value, thereby minimizing the baseline distortion.

FIGURE 6




**Input to (top) and output from (bottom) the first-order recursive low-pass filter defined in *Table 4* (bottom). To obtain this plot, the DATA statements in *Table 4* were modified to represent the 10 Hz sine wave with amplitude 40. As in *Figure 5*, the cutoff frequency of the filter was defined as 1 Hz. Note the elevated baseline of the first sine wave processed by the filter.**

Modifying a prototype low pass filter made of discrete components to create a high pass filter would likely entail several minutes of additional soldering, in addition to the time it would take locate components with the appropriate values. Because you're working with algorithms and not components in the DSP domain, circuit modifications are mere keystrokes away. For example, the low pass filter described in **Table 4** can be modified to work as a recursive high pass filter if you substitute the following line for the one in the complete listing:

```
270 TEMP = 1024 * ((INPT[I] * 1024) - LAST)
```

This BASIC statement, together with the supporting code in lines 260 to 300, implements the following relationship:

$$Output_n = \{(K\ ((Input_n \times K) - Output_{n-1})) + (\mathring{a} \times Output_{n-1})\} / K$$

where K and å are defined as in the original low pass filter design.

## Summary

DSP may one day make conventional circuit design obsolete. Designers will no longer be concerned with component values and tolerances, but will deal with algorithms and software. In part 2, I'll consider techniques used for digital signal processing in the frequency domain, including the Fourier Transform. hr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# Practically Speaking

*By Joseph J. Carr, K4IPV*

# SOME ANALOG METER APPLICATIONS

In March I looked at analog meter movements, mostly those of DC meters. This month I'll expand the discussion to include some application issues (including meter movement protection) and a construction project.

## Meter protection

The electrical environment for meters is hostile enough that meter movements seem to be perpetually at risk of total destruction. There can be physical damage. A bent pointer can result when the meter is connected into a circuit where a severe overrange current flows, or if the meter is connected backwards in the circuit. The meter can also fail to perform properly because of the strong RF fields inside a piece of ham radio transmitting equipment. Fortunately, there are a few things you can do to protect your meter.

## Overcurrent protection

The simplest way to prevent meter movement damage from overcurrent is to place a fuse in series with the meter. This approach is reasonable when the meter has a current range compatible with fuse values, but there's no fuse for 1-mA (or less) meters. In fact, meter movements much under 100 mA probably can't be protected easily with a fuse.

**Figure 1** shows a method used in some multimeters to prevent meter movement destruction when an overrange current tries to flow in the movement. Diodes CR1 and CR2 are silicon diodes (e.g., 1N4148) of the type used in low current meters; rectifier diodes (e.g., 1N400x) are used in high current meters. The silicon diode requires a forward bias potential of 0.6 volts or more for proper linear operation, and is virtually ineffective at potentials lower than about 0.3 volts. The voltage drop across the meter is the product of the current flowing in the meter and the meter's internal resistance ($R_m$). In **Figure 1**, the meter movement is 100 $\mu$A full scale and has an internal resistance of 500 ohms. (Always check; the value in your meter will probably be different.) The full-scale current will produce a voltage drop of 500 ohms × 0.000100 A, or 0.05 volts. Under normal circumstances, the voltage drop won't forward bias CR1 and CR2. However, when a severely excessive current flows, a larger voltage drop will occur and the diodes will become forward biased. When this happens the diodes shunt the meter, protecting it from harm (though probably destroying themselves).

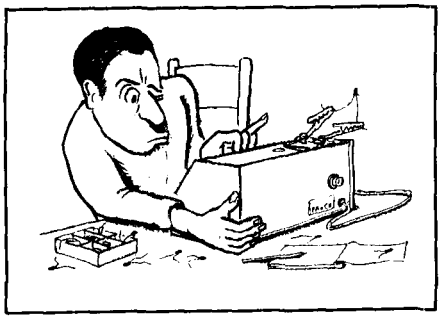
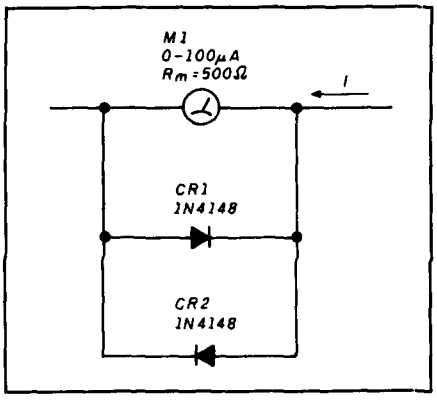
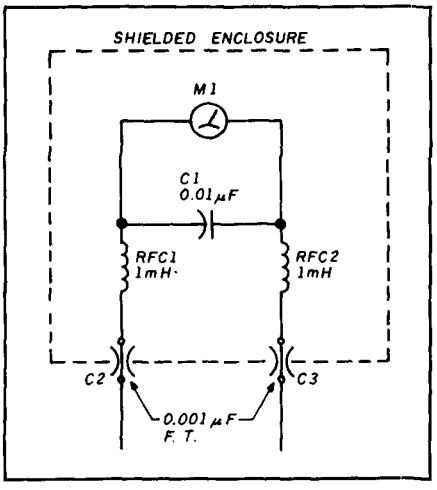


FIGURE 1




**Protection diodes shunted across the meter movement.**

FIGURE 2




**RF protection diodes.**

## RF protection

RF protection is a must for meters used in transmitters. **Figure 2** shows several of the basic methods used to protect analog meters in transmitters and linear amplifiers. It's always wise to shield the meter in the RF environment. Either use an aluminum box or fashion a shield of your own design from sheet metal. The meter movement is usually mounted directly to the front panel of the equipment, so the shield can also be mounted in that fashion. This simplifies the design and makes simple boxes a reasonable approach.

In most cases, you can get away with using a single bypass capacitor across the meter terminals. Use a disk ceramic capacitor (or other type that will work at the frequencies used in the equipment) between 0.01 $\mu$F and 0.5 $\mu$F. Make sure the capacitor is mounted directly on the body of the meter movement, stretched between the electrical connection terminals of the meter. The capacitor leads should be as short as possible. In some cases, especially with high powered amplifiers, the RF chokes (at least 1 mH) should also be added in series with the meter movement. Capacitors C1 and C2 are 0.001-$\mu$F feedthrough capacitors that are mounted directly to the shielded metal enclosure. If these aren't available or practical, use a 0.01-$\mu$F disk ceramic capacitor mounted directly on the input connections to the shielded housing. In any event, regardless of how many of these fixes are incorporated into your project, keep all leads inside the shielded

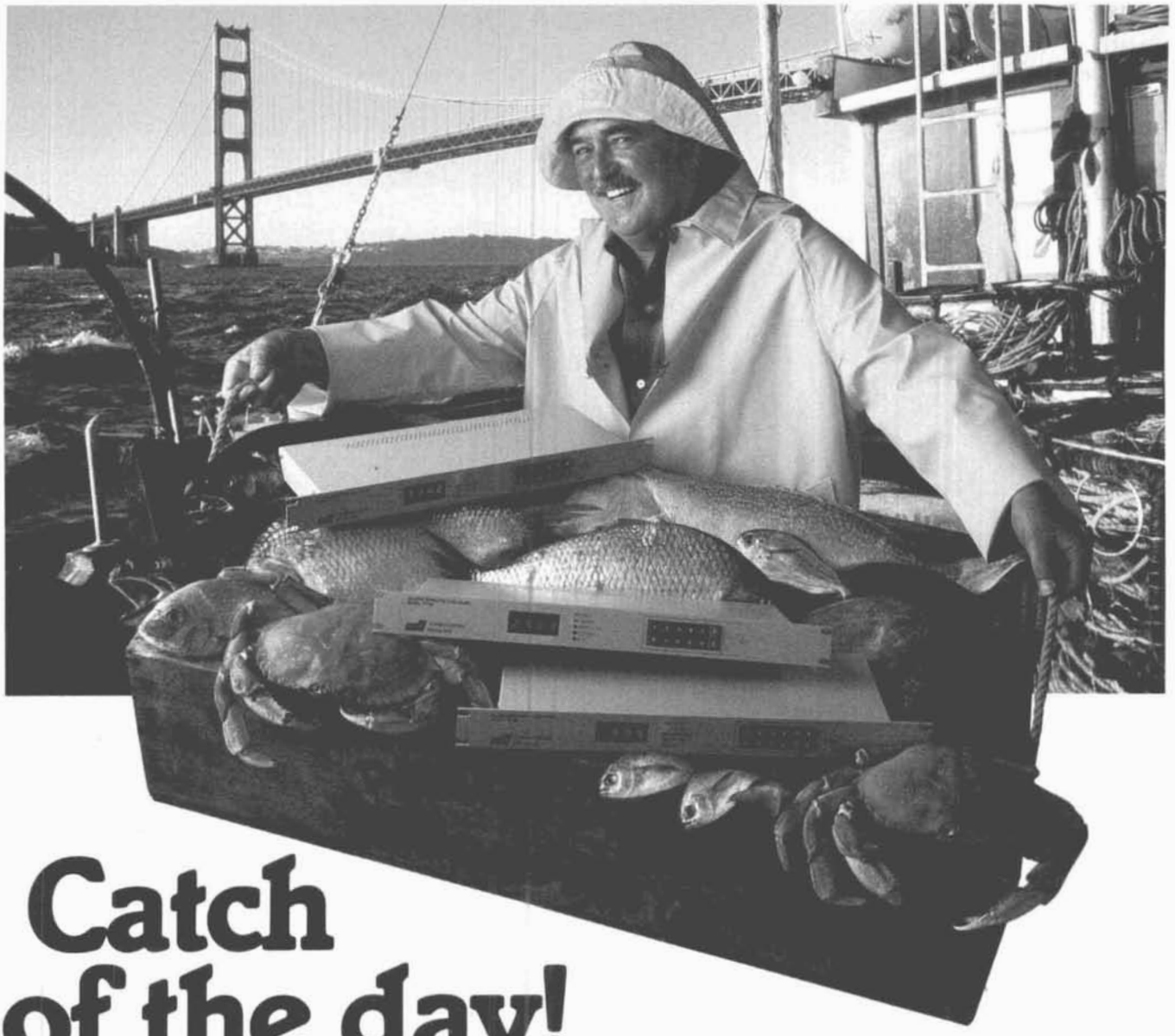


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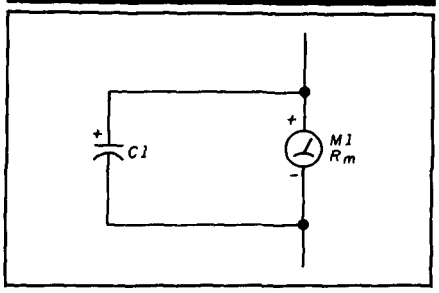
FIGURE 3




**Integrating or low-pass filtering the meter movement.**

enclosure as short as possible to minimize RF pickup.

## Damping

Some signals have either too much variation or too much noise present on them to be read effectively on an analog meter movement. In those cases it's possible to *integrate* (time average) the meter reading by connecting a capacitor across the meter terminals (see **Figure 3**). This is an old trick in the instrumentation trade. I've seen capacitors from 0.1 μF to 500 μF strapped across the meter terminals. The time constant of the circuit ($R_mC1$) tells you something about the amount of averaging in any given case.

## Amplified meters

The bare meter movement is sometimes boosted with an amplifier circuit to improve sensitivity or perform certain specific instrumentation jobs. One application (see **Figure 4A**) is to provide a higher input impedance to a microammeter or milliammeter (M1) that's connected as a voltmeter. The sensitivity of a 0 to 1-mA or even 0 to 100-μA meter movement might be too poor for practical use, especially in solid-state circuits. But if the meter is *buffered* with an operational amplifier (op amp), as in **Figure 4A**, then the circuit sees the high input impedance of the op amp (A1). In order to maintain that high impedance, be sure to use a device with a very high input impedance. The CA-3140 type is a good example of such a device, and it's both inexpensive and widely available from mail order and local parts sources.

In the circuit of **Figure 4A** the op amp (A1) is connected in the *unity gain noninverting follower* configuration. When a DC voltage ($V_{in}$) is applied to the +IN terminal of the op amp, exactly the same potential appears at the output terminal; that is, point A. The full-scale input voltage and the series resistors are used to custom tailor the meter movement for voltmeter service. The sum of R1 and R2, denoted R, is found from:

$$R = \frac{V_{in\,(max)} - I_{fs}R_m}{I_{fs}} \qquad (1)$$

where R is the sum R1 + R2, $V_{in(max)}$ is the maximum allowable full-scale input voltage, $I_{fs}$ is the full-scale current of meter M1, and $R_m$ is the internal resistance of M1.

If you want the meter to see an amplified voltage, then use the alternate *noninverting follower with gain* circuit[1] of **Figure 4B**. The DC voltage gain of this circuit will be:

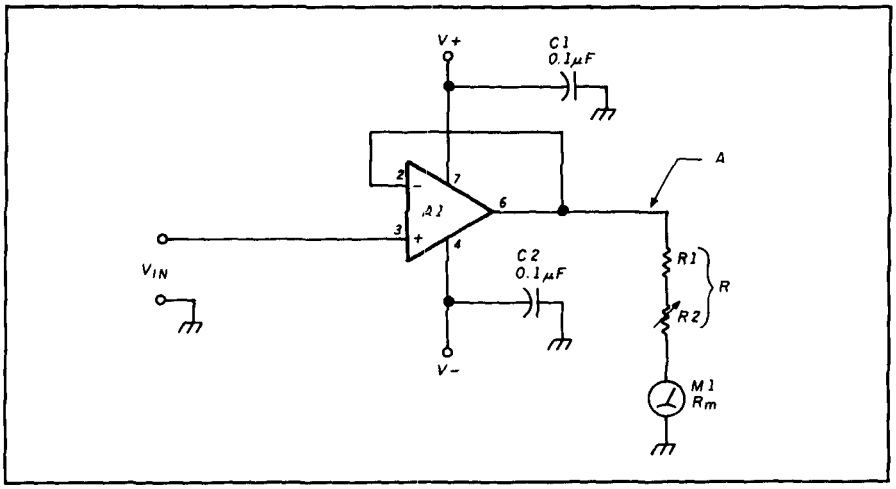
$$A_V = \frac{R4}{R3} + 1 \qquad (2)$$

The voltage at point A (**Figure 4A**) will be $A_V V_{in}$. In that case, the maximum voltage at point A will be used in place of $V_{in(max)}$ in **Equation 1** to find R.

The reason for making the meter multiplier resistor from two resistors (R1 and R2) instead of a single resistor is twofold. First, it allows a series combination of a fixed resistor and a potentiometer to replace a hard-to-find precision resistor. Second, it lets you adjust the meter to full scale easily. Apply the maximum value of input voltage to the noninverting input of the op amp and then adjust R2 for a full-scale deflection of M1.

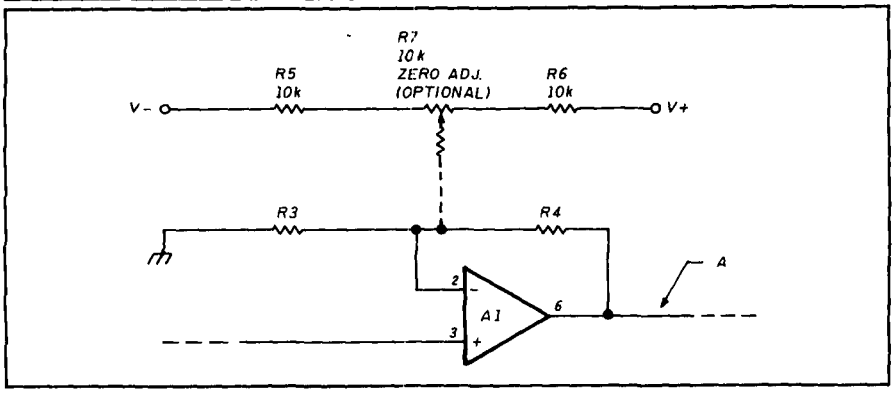
**Figure 4B** shows an optional **zero adjust** circuit. This circuit injects a counter current into the summing junction of the op amp (at the inverting input). This current is used to counteract any DC offset that's present in the op amp output. To adjust R7, short the input terminal (the noninverting input of A1) to ground, and mechanically zero the meter movement pointer

FIGURE 4A




**Unity gain voltmeter.**

FIGURE 4B




**Gain voltmeter.**

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|---|---|---|---|---|---|---|---|---|
| 0550G | 50-54 | 10 | 400 | .6 | 15 | 13.6 | 60 | UHF |
| 0552G | 50-54 | 25 | 400 | .6 | 15 | 13.6 | 55 | UHF |
| 1450G | 144-148 | 10 | 400 | .6 | 15 | 13.6 | 54 | UHF |
| 1452G | 144-148 | 25 | 400 | .6 | 15 | 13.6 | 50 | UHF |
| 2252G | 220-225 | 25 | 220 | .7 | 14 | 13.6 | 36 | UHF |
| 4450G | 420-450 | 10 | 175 | 1.1 | 12 | 13.6 | 34 | N |
| 4452G | 420-450 | 25 | 175 | 1.1 | 12 | 13.6 | 29 | N |

Models also available without GaAs FET preamp (delete G suffix on model #). All units cover full amateur band – specify 10 MHz bandwidth for 420-450 MHz amplifier. Continuous duty repeater amps also available.

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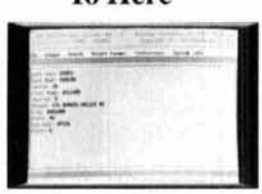

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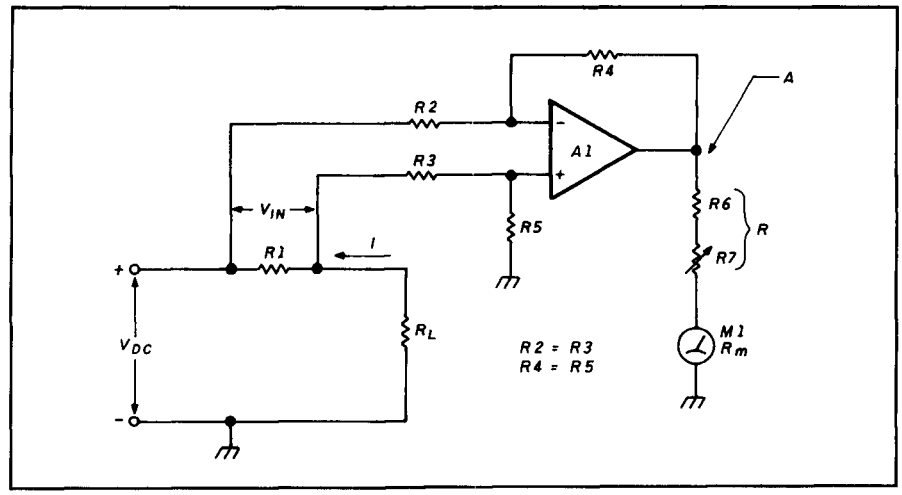
(there's usually a small screw adjustment on the front panel of the meter). Next, turn on the circuit and adjust R7 for exactly zero reading on the meter.

**Figure 5** shows how to make a milliammeter or microammeter into a higher current instrument with an amplifier. Resistor R1 is a small value, precision resistor that's placed in series with the load ($R_L$) of the circuit. The current flowing in the circuit without this resistor is $V_{dc}/R_L$, and with the resistor it reduces slightly to $V_{dc}/(R_L + R1)$. It's very important to make $R1 << R_L$ to prevent errors.

The op amp is connected in the *DC differential amplifier* configuration, and has a differential DC voltage gain of R4/R2. The input signal seen by the amplifier (A1) is the voltage drop across R1, which is I × R1. The voltage appearing across the meter circuit at point A is $V_{in}$R4/R2. It's important to keep R2 = R3 and R4 = R5. The values of R6 and R7 are found in the same manner as in the example of **Figure 4A**.

FIGURE 5




**Amplified current meter.**

## Precise AC voltmeters

The AC voltmeter is made from the DC meter movement plus a rectifier. Last month you saw a bridge rectifier that can be used to make the meter read AC values. Unfortunately, that circuit is limited when it comes to very low voltage circuits. The problem is caused by the fact that PN junctions are diodes which are used to rectify the AC before applying it to the meter movement, and aren't linear in the low range of forward bias voltage (see **Figure 6A**). The ideal diode I-versus-V curve is a straight line from zero to maximum current. It's said to be an *ohmic device* because it obeys Ohm's law. But actual diodes are ohmic only in the region above a certain critical voltage ($V_g$); this voltage is 0.2 to 0.3 volts in germanium diodes and 0.6 to 0.7 volts in silicon diodes. Below $V_g$ the diode is nonlinear, and therein lies the problem.

A "fix" for nonlinearity is to replace the diode with a *precise rectifier circuit* like **Figure 6B**. This circuit uses an op amp to "servo-out" the nonlinearities and make the diode act more like the ideal diode. That's why this circuit is also called the *ideal rectifier* or *ideal diode* circuit. Diodes CR1 and CR2 are ordinary silicon signal diodes like 1N914s and 1N4148s.

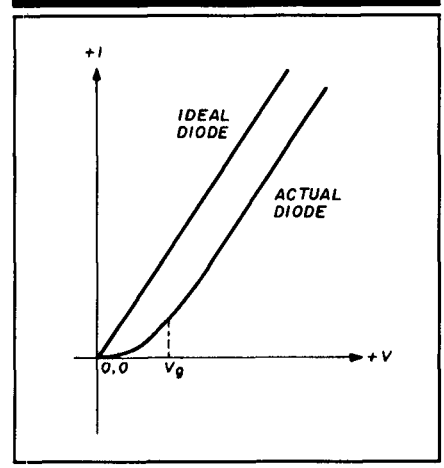
The diodes are shown in **Figure 6B** in the direction that will produce a positive $V_o$ with positive $V_{in}$; this circuit is called a *positive precise rectifier*. Reversing the diode direction (turning them around) will produce a positive output voltage with negative input voltage (*negative precise rectifier*).

Unfortunately, the circuit of **Figure 6B** is only a half wave and won't prove as useful in many applications. A *full-wave precise rectifier*, also called an *absolute value circuit*, is shown in **Figure 6C**. This circuit combines a positive precise rectifier and a negative precise rectifier into a single output. The output is obtained by amplifier A1, which is a gain-of-one summer. As long as all three resistors have the same value, the output voltage ($V_{odc}$) will be the sum of the two rectifier outputs ($V_{01}$ and $V_{02}$), and will be a full-wave rectified pulsating DC waveform.

## RMS AC voltage measurement

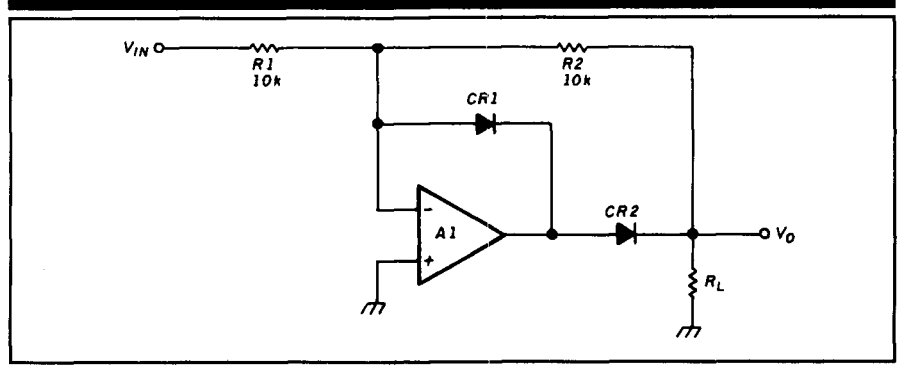
Simple rectified meters produce a reading that's related to the average voltage of the pulsating DC waveform produced by the rectifier. It can be interpreted to be the *root mean square* (rms) voltage only if the scale is so calibrated *and* the input waveform remains a sine wave. However, that condition isn't often met in practice. A

FIGURE 6A



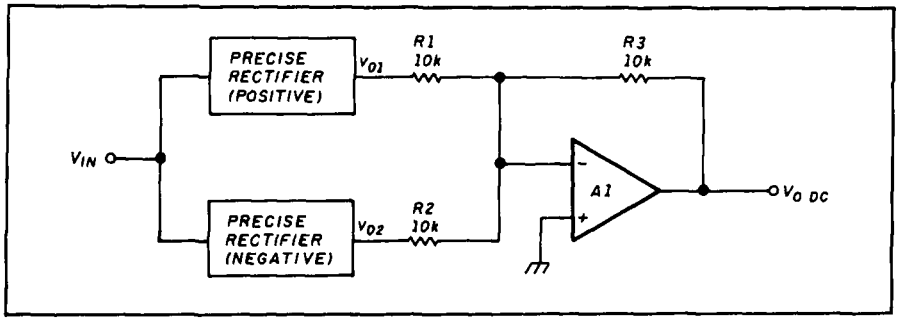

**Ideal and actual diode I-versus-V curves.**

FIGURE 6B



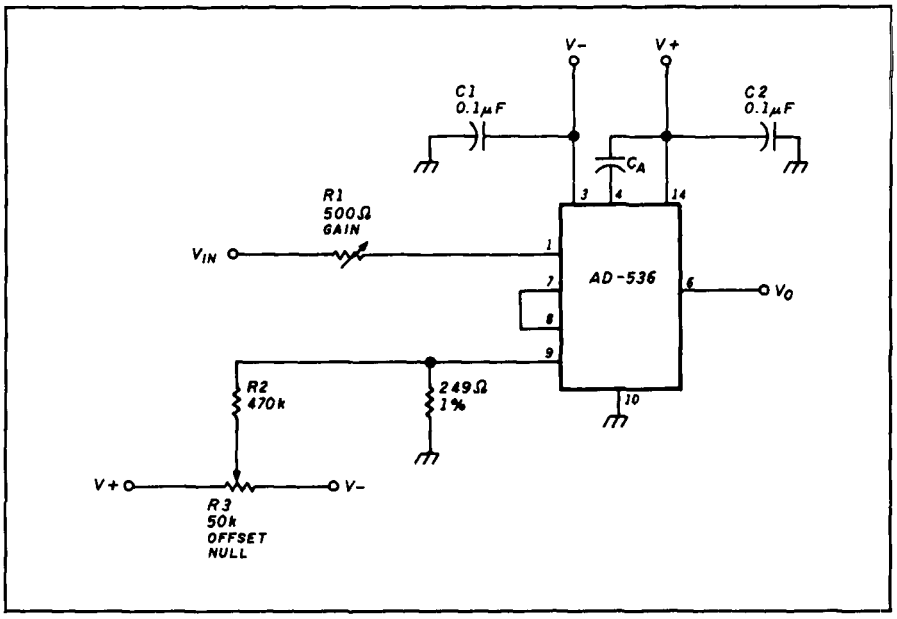

**Half-wave precise rectifiers.**

FIGURE 6C



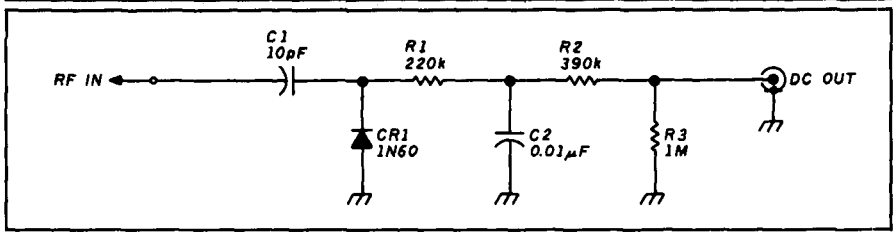

**Block diagram of a full-wave precise rectifier.**

FIGURE 7




**RMS-to-DC converter.**

FIGURE 8




**RF demodulator probe.**

true rms voltmeter requires an *rms-to-DC converter* circuit. **Figure 7** shows an rms converter based on the Analog Devices, Inc. AD-536 rms-to-DC converter. The output voltage will be a DC voltage that's proportional to the rms value of the applied AC voltage. The AD-536 will operate properly at frequencies to 300 kHz if the input rms voltage is >100 mV, and 2 MHz if the input rms voltage is >1 volt. This circuit offers *offset null* (zero) and *gain* controls.

## RF voltmeters

Many Amateur Radio applications require measurement of RF voltages. For example, the output power of a CW transmitter can be determined in many cases by measuring the voltage across a dummy load. In other cases, troubleshooting requires such measurements. **Figure 8** shows a simple, passive RF demodulator probe. This circuit must be built into a shielded enclosure, and consists of a rectifier (CR1) and an RC ripple filter that uses values consistent with the RF frequencies. The diode used should be a germanium diode of the 1N34, 1N60, or ECG-109 type (also NTE-109). This circuit works for voltages up to about 40 volts peak. Increased voltages are obtained by series connecting two or more rectifiers.

## Bridge amplifier: a construction project

There are many applications for a differential voltmeter, especially in circuits like the Wheatstone bridge used for measuring resistive impedances and other parameters. The circuit in **Figure 9** can be used for either single-ended or differential applications. **Photo A** shows the completed project.

The input or "front end" of the differential voltmeter is an instrumentation amplifier consisting of three operational amplifiers: A1, A2, and A3. The two input amplifiers, A1 and A2, should be RCA/GE CA-3140 devices or other high input-impedance equivalents. These devices are easily obtained on the mail order parts market. The gain of this section of the amplifier is found from:

$$A_v = \frac{2\ R2}{RI} + 1 \qquad \textbf{(3)}$$

For the value shown, the differential voltage gain of this section is ×20. The presumption is that R2 = R3.

The follow-on post amplifier (A3) has a gain of:

$$A_v = \frac{R6}{R4} \qquad \textbf{(4)}$$

assuming that R4 = R5, and R6 = R7+R8. With the values shown, the gain of the post amplifier is ×10, so the overall gain of the instrumentation amplifier section is the product of the two gains, or ×200.

The *scaling amplifier* (A4) has a switch selectable gain of ×1 or ×10, making the overall gain of the circuit either ×200 or ×2000. With this higher gain, the DC offsets of the amplifiers can be quite high, so a ZERO control (R13) is provided.

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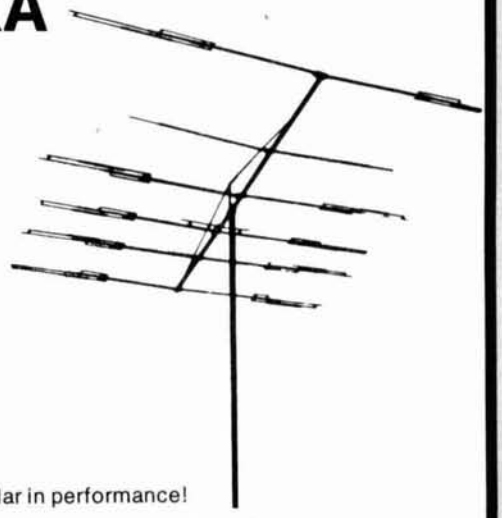

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  28-29.7 MHz
- Gain ..... 8.5-9 dB
  9-9.5 dB
  11-11.3 dB
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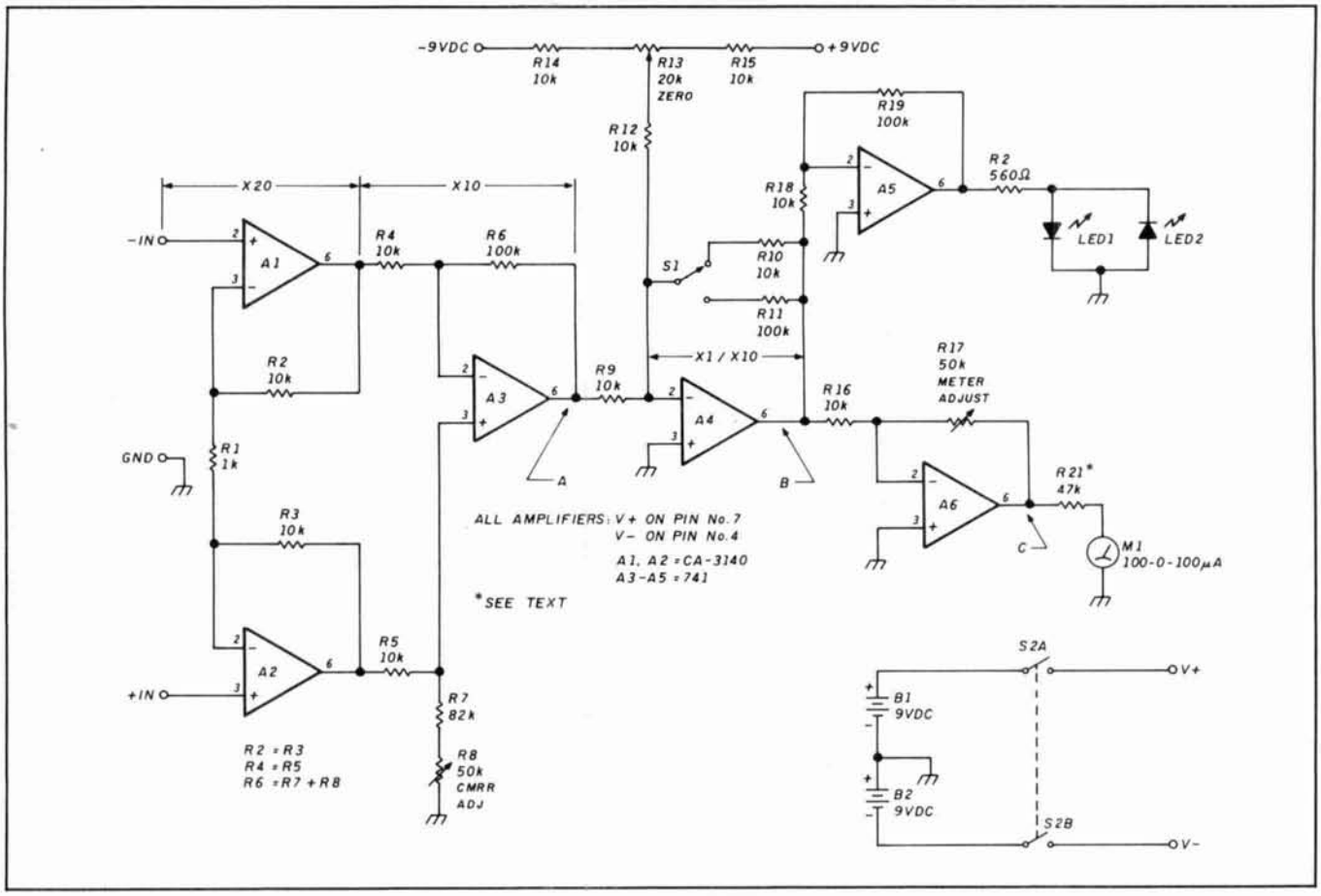
FIGURE 9




**Differential voltmeter circuit.**

differential voltmeter. The conventional output uses a 100 to 0 to 100 μA DC meter in the manner described previously. The second output stage is a null detector made from light emitting diodes (LED1 and LED2). The null condition is indicated when **both** LEDs are extinguished.

Initially, you should adjust the meter with the controls in the following positions:

−IN/+IN: shorted to ground
S1: in the ×10 position (to R11)
S2: off
R8, R13, R17: midrange

Turn on the meter power switch (S2) and let the instrument warm up for about ten minutes. If the meter movement is pegged, adjust R17 to bring it back to the proper range. Adjust CMRR ADJ (R8) for zero volts at point A. Next, adjust ZERO (R13) for zero volts at point B. Now, unground −IN/+IN and connect these two terminals together. Apply a 1.5-volts DC signal from a dry cell to the shorted inputs. Readjust R8 for zero volts at point A if any change has occurred. Recheck the voltage at point B and readjust R13 if necessary.

If a precision value of $V_{in}$ is to be read, then apply a 1-mV DC potential to −IN, and ground +IN. Adjust R17 for the proper deflection of M1. This mode isn't normally needed, however, because the purpose of the meter is to find a null.

The DC differential voltmeter is usable as is, or you can use the information contained in this article to design your own application. Don't try to obtain gains higher than ×2000, because the drift and other defects will conspire to cause problems. You need to use special compensation methods in those cases.

PHOTO A

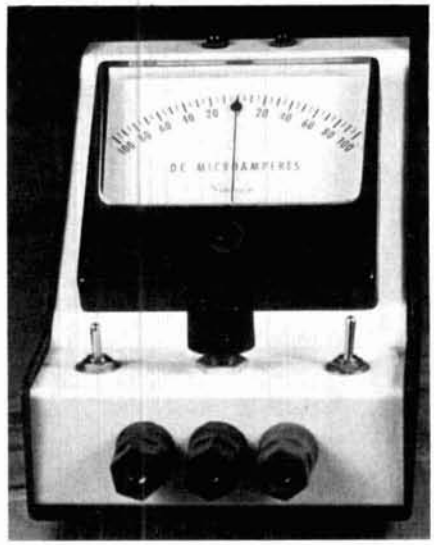


**Photo of *Figure 9* project.**

## Conclusion

The analog meter movement can be used in a variety of applications. These are some of the more popular applications of those meters used in conjunction with other circuitry. hr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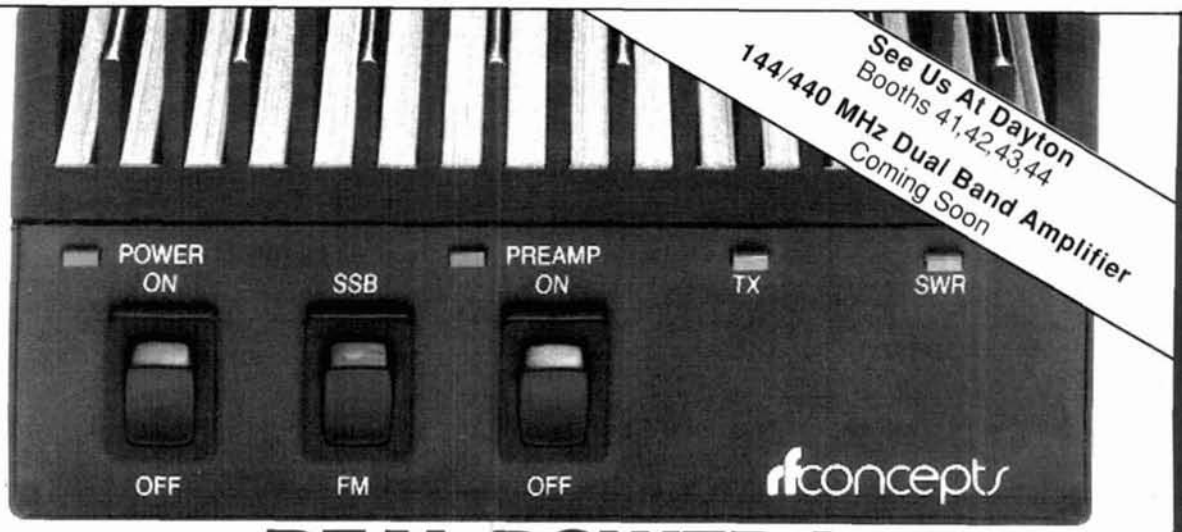
1. Joseph J. Carr, K4IPV, *IC User's Casebook*, Howard W. Sams & Company, Inc., Cat. no. 22488, $12.95.

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If you like to tinker and build, or if you just want to make routine equipment repairs, an assortment of electronic test equipment is undoubtedly on your "must own" list. If you're like me, you have a pretty strong attachment to your test equipment. After all, your equipment lets you put your knowledge to work.

Unfortunately, not all test equipment is created equal — even equipment with exactly the same specs. I'd like to talk about the pros and cons of various pieces of apparatus to help you select the right tools for your purposes, and get the most out of them once you own them. My topic for this article is oscilloscopes.

Years ago there was a question on the General class exam that asked which measurement tool was the most useful of all. The correct answer, to no one's amazement, was the oscilloscope. It's still a good answer. If you could own just one piece of equipment, you'd probably be wise to choose a good 'scope because it lets you look at both static and dynamic voltages.[1] Choosing a good 'scope isn't difficult if you take the time to learn a few indicators.

Note that I said a "good" 'scope. There are a lot of junky ones on the market, and your first task is to sort out the equipment that's part of the solution from the equipment that's part of the problem. A discussion of what constitutes a good 'scope can be broken down into sections: vertical amplifiers, timebases, trigger circuits, and extra or convenience features. Of these, the vertical amplifier portion of the oscilloscope deserves the most discussion.

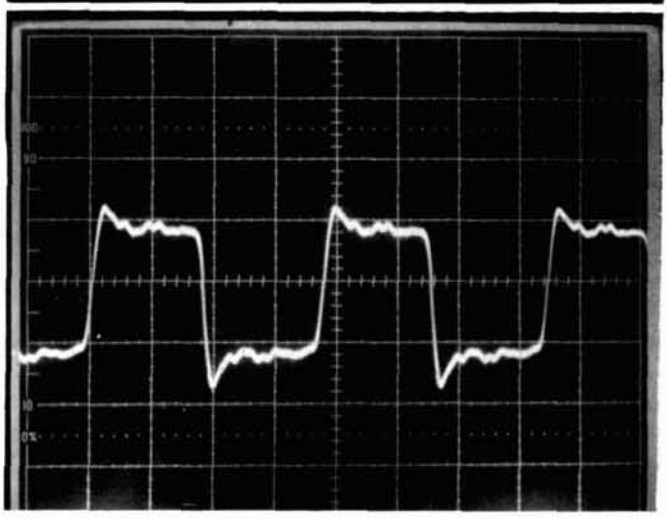
## Vertical amplifiers

Look at almost any oscilloscope ad and you'll find the bandwidth specification displayed prominently. In fact, for many buyers bandwidth has too often become the single figure of merit for oscilloscopes. That's too bad; buyers who consider only bandwidth when buying an instrument are courting disappointment. To further compound the problem, a lot of people misunderstand and are misguided by the meaning of bandwidth.

In a nutshell, you probably need a lot more bandwidth than you think you do. Looking at your 9-MHz IF? A 10-MHz 'scope ought to do the job with some room to spare, right? Well, no. In fact, not by a long shot.

**Photo A** shows the output of an oscillator designed to what I'd call minimal standards. The oscillator was made from a fast logic gate and crystal, breadboarded with little regard for layout. The result is a 13.3-MHz signal, which could charitably be described as "rich in harmonics." What you see in **Photo A** is a reasonably accurate representation of the output of the circuit taken from a 200-MHz oscilloscope.

**PHOTO A**



**A 13.3-MHz square wave, rich in harmonics, as seen on a 200-MHz oscilloscope.**

Now look at what happens when you try to use an instrument with inadequate bandwidth. **Photo B** shows the same signal, with the same sensitivity and sweep settings, on a 40-MHz oscilloscope. You'd think that a 40-MHz instrument would be ample for a 13.3-MHz signal, but in this case it's not. **Photo C** shows a worse case, from a 20-MHz oscilloscope. The bandwidth of the instrument is 50 percent higher than that of the signal, but the on-screen trace is nothing like the one from the 200-MHz 'scope. Why do you need so much bandwidth for a 13.3-MHz signal? The reason is easily understood.

Bandwidth is *not* the same as frequency range. When some instruments (like frequency counters) are specified, the term frequency range is used because counters work properly up to some limit and then cease counting. Oscilloscopes don't work up to a particular frequency and stop. They start rolling off slowly until they reach the point where they are 3 dB down. We call that the bandwidth. In short, a pure sine wave will appear at 70.7 percent of its true amplitude if its frequency is equal to the bandwidth of the 'scope.

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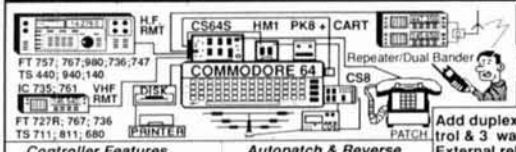


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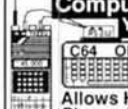

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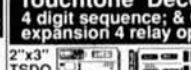


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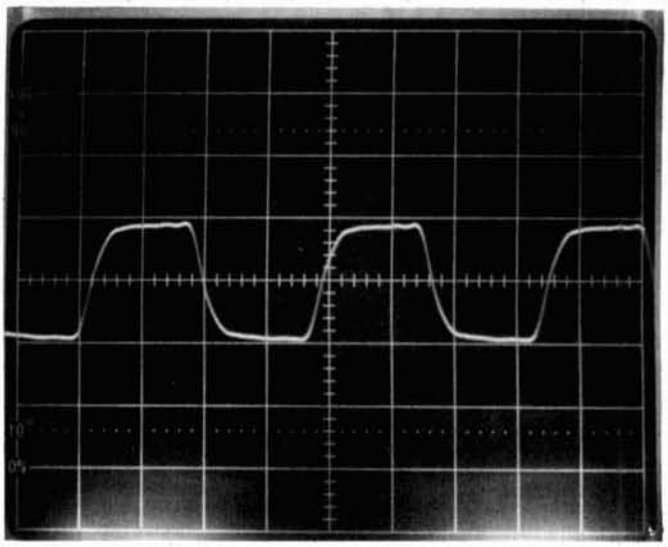
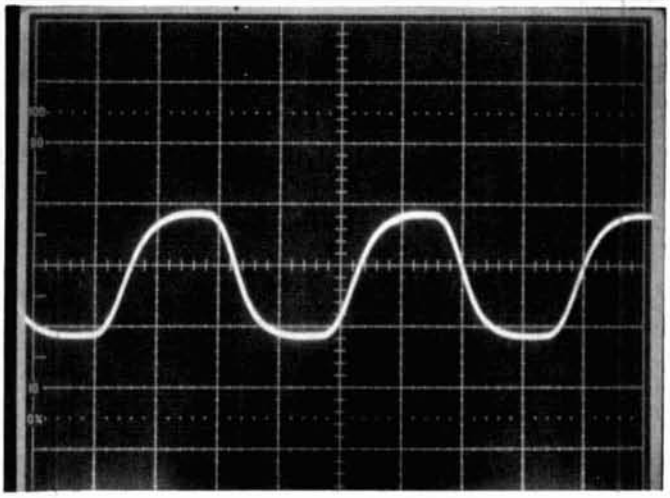
PHOTO B



**Same 13.3-MHz signal, as seen on a 40-MHz oscilloscope.**

PHOTO C



**Same signal again, but on a 20-mHz oscilloscope.**

If you know in advance that you're dealing with a pure sine wave, you might find such a measurement useful — but this is seldom the case. In fact, if you know that your signal is only a pure sine wave, you can get all the information you need from an RF voltmeter and a frequency meter. Normally circumstances don't justify the assumption that your signals are sine waves. Actually, in RF work it's often the deviations from sine waves that are interesting.

If you don't know in advance what your waveforms look like, you need a bandwidth *much* greater than your signal. An ideal square wave contains an infinite series of odd harmonics. To obtain a reasonable representation of a square wave, you probably ought to think about getting an instrument with a bandwidth five to seven times the frequency of the signals you want to look at. And next time someone tries to tell you that his 35-MHz oscilloscope is really good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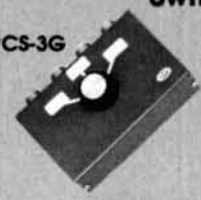

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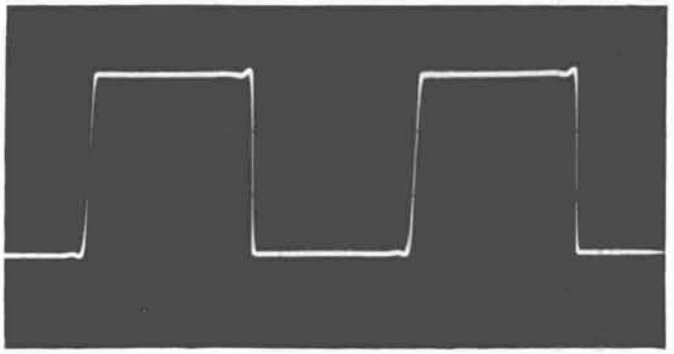
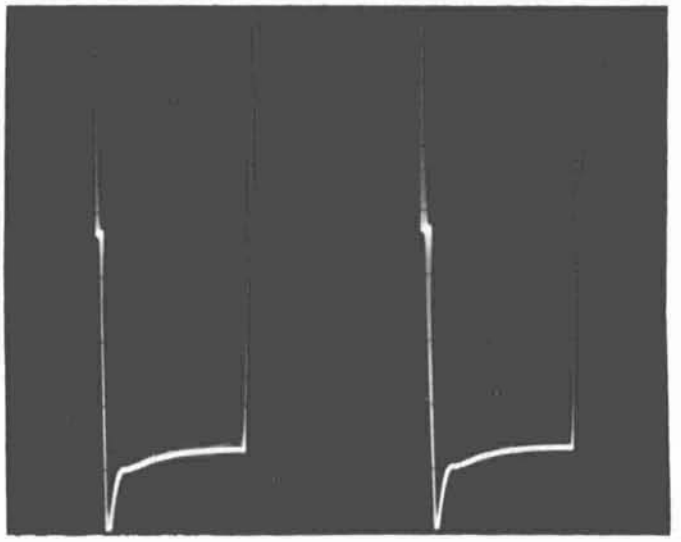
PHOTO D



**High quality square wave.**

PHOTO E



**Same square wave, same oscilloscope, but at twice the sensitivity.**

to 40 or 50 MHz, you'll know better! Even for sine waves, you'll want a bandwidth three to five times greater than your signal. Otherwise, your measurements will be in error by more than a reasonable factor. A 35 to 65-MHz instrument, rather than a 10-MHz instrument, is what you want for measurements on your 9-MHz IF.

Once you've decided on a bandwidth, you need to look at the pulse characteristics of your instrument. This is necessary *even if your're not looking at pulses.*

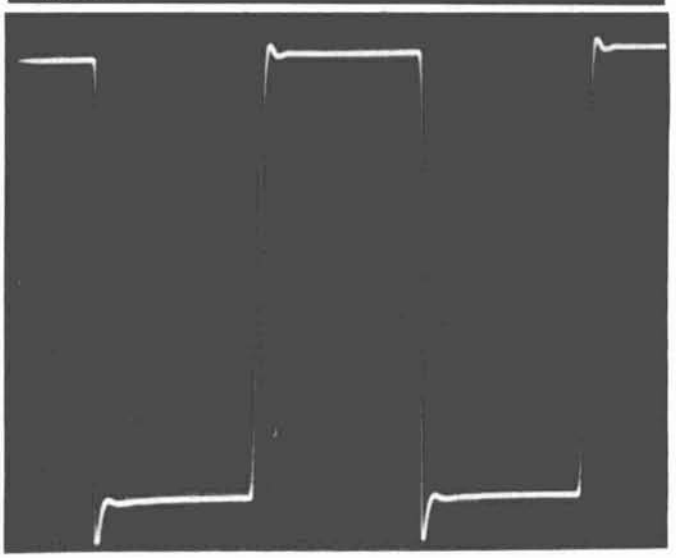
The impulse response of a system is the Fourier transform of its transfer function. This is simply a $64 way of saying that if you tell me exactly how something responds to a very fast pulse, I'll tell you exactly how it responds to *anything*. In other words, if you mess with pulse response you also mess with bandwidth. The consequence of accepting poor pulse response is that you also accept sensitivity that isn't flat and doesn't roll off in a neat, orderly manner. You'll find bumps and dips in the sensitivity of the instrument that are well within its useful range. So what do you look for, and how good is good?

The easiest spec of pulse fidelity is the overshoot and ringing spec. A small percentage of overshoot and ringing extends the bandwidth of the oscilloscope greatly. The usual practice is to allow 3 to 5 percent. Be very suspicious of any 'scope that doesn't have this spec. Some manufacturers ignore this parameter altogether. It lets them offer a very high bandwidth for a very low cost, but you won't be satisfied with the instrument for long.

It's important to look for a combination of a good bandwidth spec and good pulse response when you're checking out a new oscilloscope. However, there's one other parameter that's seldom mentioned, and it's one that's very important. I'm talking about dynamic range.

Photo D is taken from a 'scope that's a bit short on this parameter. It shows a square wave from a good quality signal generator. The representation is pretty good. There's a little negative tip on the negative-going slope that really isn't there, but it's small and well within reasonable specs. Now double the sensitivity and use almost the entire screen. You'll see that those innocuous little tips have grown to problem size in Photo E. Advance the gain a little more, as in Photo F, and they really get unruly. Position the trace upward, as shown in Photo G, and the picture changes dramatically once again.

PHOTO F



**Same square wave again, but at a sensitivity that puts half the signal offscreen.**

PHOTO G

**Same signal, same setttings, except that the vertical position control has been changed.**

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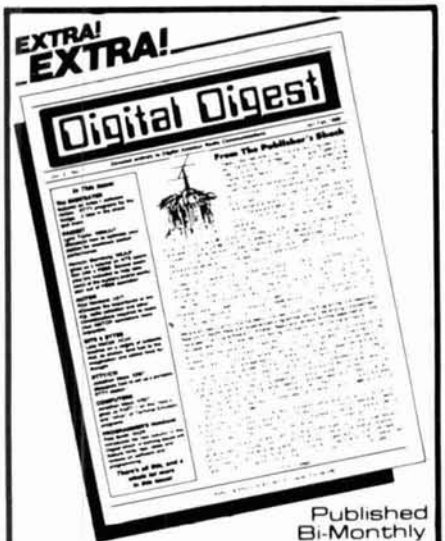


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You should expect *some* errors of this type; however, these are a bit large. Unfortunately, I've seen products far worse than this. Since dynamic range is seldom specified, it's an item you'll probably have to check for yourself using the method shown here.

Here's another very common trap. You plunk down $700 for a racy new 50-MHz oscilloscope, and then spend $50 on a pair of 50-MHz probes. Sounds reasonable — a nice, matched system. It's not. A 50-MHz probe on a 50-MHz 'scope will give you about 35 MHz at the probe tip! And the degradation is even more severe when you hook it to a point in your circuit with more than 50 ohms of source impedance!

It's better to spend a few more bucks for probes of 250 MHz (three to five times your 'scope bandwidth) or so. That will preserve almost all of your bandwidth at the probe tip, and give you more megahertz per dollar.

One more thing before I go on. Check out the cabinet material before you buy. You'd be surprised how many plastic 'scopes suddenly get inch-wide traces when you key your transmitter. A metal case doesn't guarantee immunity, but it does better the odds. I've found that plastic-cased oscilloscopes tend to be more EMI susceptible, and should be avoided unless they have good internal shielding.

There's a lot of information out there about vertical amplifier sections,[2] but what I've discussed here should provide you with a better than average survival kit. These simple questions and performance tests will keep you from making the most common errors in oscilloscope selection and use.

## Timebases

Old, low cost oscilloscopes used a horizontal oscillator with a synchronizing circuit. The idea was to synchronize the frequency of the timebase with the frequency of your signal, thus obtaining a usable display. That style of design has pretty much become a thing of the past.

The calibrated and triggered timebase has played a major part in making oscilloscopes as popular and useful as they are today. A timebase is fairly easy to check out; it's supposed to be linear. Just feed in any signal known to produce peaks or zero crossings at a rate of one or five times per division, and look for errors. If the largest error on any timebase setting is 2 to 3 percent, the 'scope is doing as well as analog models generally do. Photo H shows a typical example.

Usually, the oscilloscope has a magnifier which allows you to expand the trace five or ten times. This is quite handy. It's often the only way you can spread out a signal at the upper limit of your 'scope enough to really look at it. You'll pay a price in accuracy, but that's to be expected. Don't be surprised if the accuracy spec at the fastest sweep speeds isn't nearly as good as it is on the other ranges. That, also, is to be expected.

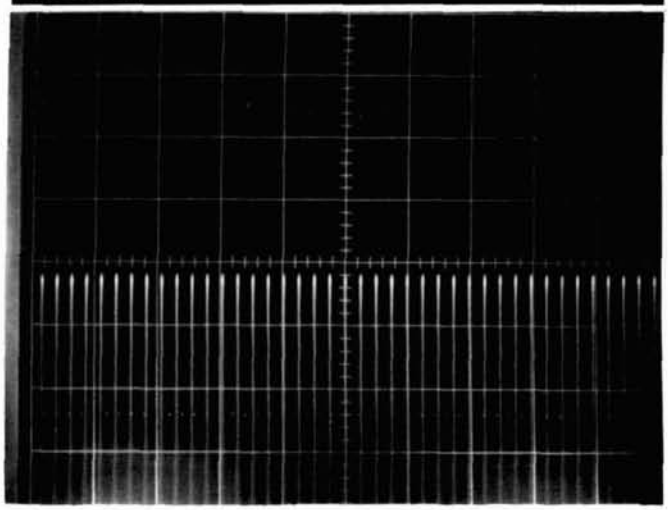
Delayed sweep is a commonly misunderstood feature. When I called on customers for a major instrument supplier, I was always amazed at the number of people who felt they positively, absolutely had to have delayed sweep — even though they didn't have the foggiest idea what it was or how to use it. If you really need it, nothing else will do. If you don't, save your money. Delayed sweep adds significantly to the cost of an oscilloscope.

Delayed sweep means that an instrument has two independent timebases. The second sweep runs after a delay of your choice, using the first timebase. The most useful implementation of this involves a precision, multiturn pot.

Why the precision pot? It lets you make a more precise time measurement than is possible on the screen. You don't have to worry about CRT nonlinearity, so you can get down into the 1/2-percent range by reading time off the pot rather than the screen. This can be important if, for example, you're trying to measure timing on a digital circuit. With the instruments available today, there are often better ways to make timing measurements without the expensive precision pot.

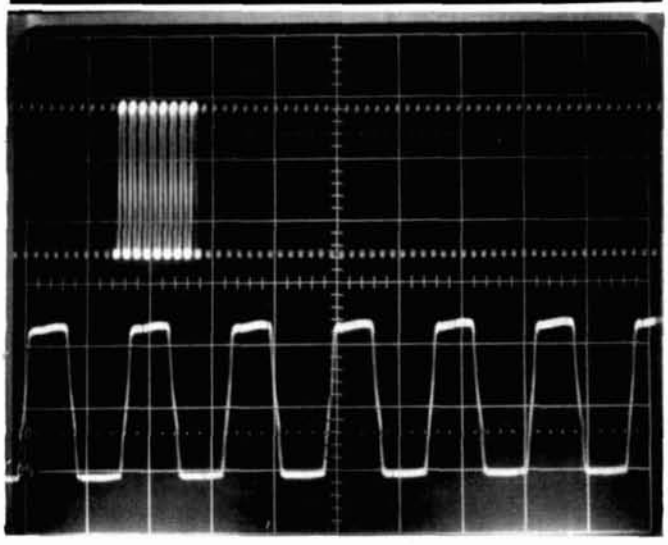
Another use for delayed sweep is found in digital work. Suppose you have some sort of index pulse, followed by a hundred or so data pulses. Let's also suppose that you want to look at each pulse in turn. A convenient way to do so is to use the delayed sweep. Let me illustrate.

PHOTO H



**RF generator and frequency counter used to check time base accuracy.**

PHOTO I



**Upper trace: main sweep with an intensified zone (A intensified by B). Lower trace: delayed sweep alone. This photo required a double exposure. Some oscilloscopes are capable of actually presenting both traces on screen at the same time.**

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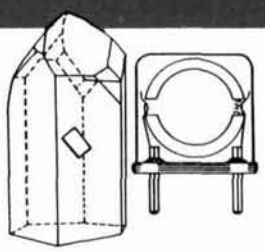

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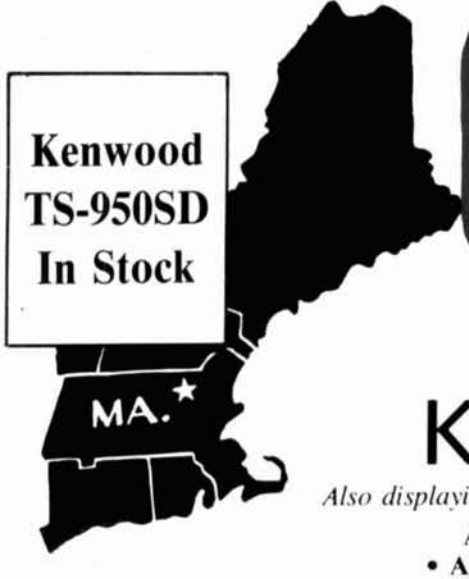


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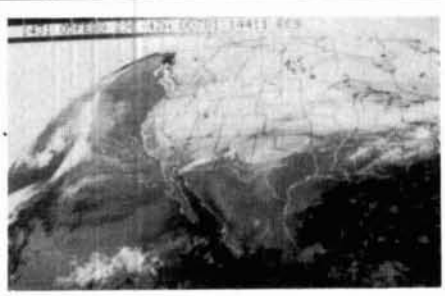

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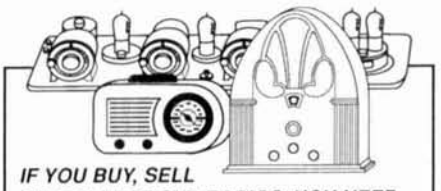

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Photo I shows an upper trace with an intensified zone. The whole trace is a product of the main timebase. The intensified zone is a product of the delayed (second) timebase. The starting point of the intensified zone depends on the pot setting, regardless of whether it's a precision pot or not. The length of the intensified zone is the length of the delayed sweep. Changing from "main" or "A" timebase to "delayed" or "B" produces the result shown in the lower trace — an expansion of the intensified zone.

If you want to examine a pulse train, set it up as shown in the upper trace and look at the greatly expanded pulses using the B sweep. A good example of an application of this kind is viewing data off hard disks.

Once again, I'd advise you not to spend the money for delayed sweep if you don't have a use for it. The sweep magnifier lets you do almost the same thing on a more limited scale, and for some applications it's just as satisfactory.

## Triggers

The most basic trigger circuit should have AC and DC coupled triggers, and a choice of positive slope or negative slope. This permits locking on the front or back side of most signals. These basic functions take care of a surprising number of situations. My own instrument spends practically all of its time in the AC+ position.

There are, however, some nice additions that don't necesssarily cost a lot. These are AC low frequency reject, AC high frequency reject, line, single sweep, TV horizontal, and TV vertical.

AC high frequency reject is useful if you need a stable display of an AM signal and want to look at the modulation envelope. With simple AC or DC coupling, your 'scope is going to have a hard time. Each RF cycle evokes a trigger, and you won't be able to stabilize the envelope. Switch in the high frequency reject, and the RF usually becomes a blur. But the envelope is stable, because that's all the trigger circuit can "see."

The reverse situation is also useful. If, for example, you want to look at an audio frequency signal that's carrying a bunch of RF trash, you can invoke AC high frequency reject and the 'scope will trigger on the AF only.

Line triggering locks you to the AC line. Want to know if a signal is derived from the AC line? Just switch to the line triggering, and look to see if your signal stands still on the screen. If it does, your signal is line related.

Single sweep doesn't rearm the trigger after the sweep. It lets you wait for a trigger, generate one sweep, and then waits for you to rearm the sweep manually. This gives you a convenient way to monitor for a single shot event.

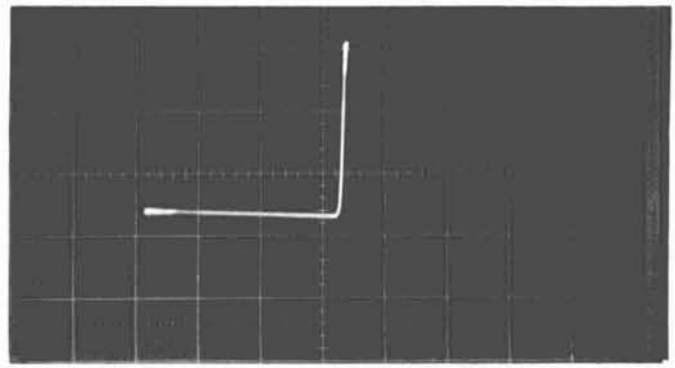
TV horizontal and vertical triggers are just that. They let you generate a stable trace that's locked to the horizontal or vertical synch pulses of a television signal.

## Extras or convenience features

The following features are ones that often make the difference between a tool that does the job and one that makes the tool a joy to use.

Consider a simple thing like graticule illumination. You'll notice that you can see the graticule in some of the photos, and in others you can't. The difference lies in the addition of a couple of small light bulbs and a pot on the front panel. This isn't a big deal from the expense standpoint, and is a very useful feature — even if you're not planning to take photos. Recently I finished up a little keyer for portable use, and had to trim the timing components for an exact 3:1 dash/dot ratio. I turned on my 'scope and put the timebase on one of its slower sweep speeds. My 'scope is a Heath SO-4251, which is not a storage model, and I had to turn down the lights to see the afterglow of the trace. When the lights are low, you can't see the graticule unless it's illuminated, so I was very glad to have that convenience.

PHOTO J



**Current versus voltage curve of a diode, using the component tester in an oscilloscope.**

A component tester is another feature that can be very handy. This is a simple gadget that allows you to put a low amplitude swept voltage on a component, even when it's in a circuit, and lets you see the component's current-versus-voltage curve. The component tester lets you make some intelligent estimates of component values and quality. You can see the trace of a good diode in Photo J. One word of warning. It's very tempting to use a component tester across the emitter base junction of a transistor. If you're not careful, this can have a very insidious effect. It will degrade the beta of the transistor you're testing,[3] and create problems you didn't have before. Emitter base breakdown apparently punches little holes in the junction; that's something you don't need.

You might also want to check out the tube's accelerating potential. The higher it is, the brighter the tube will be. On a 20-MHz instrument 2.5 to 3 kV is enough. This is often the voltage manufacturers use, because it provides satisfactory performance at a reasonable price. As bandwidth increases, so does your need for accelerating potential, since your timebase spreads your beam electrons thinner and thinner as you increase sweep speed. Seven to 9 kV are nice in a 40 to 50-MHz instrument; very high bandwidth units can have potentials up to 25 kV.

## Conclusion

Though this isn't a complete discussion of oscilloscopes, I hope you'll find it helpful as a purchasing guide. The information presented here should enable you to make an informed decision and have some sense of where the oscilloscope you choose can be properly applied. hr


REFERENCES

1. *The 1989 ARRL Handbook for the Radio Amateur*, ARRL, 1989, page 25-26.
2. Bob Orwiler, "Vertical Amplifier Circuits," *Tektronix*, 1969.
3. Wes Hayward and Doug DeMaw, *Solid-State Design for the Radio Amateur*, ARRL, 1986, page 24.

# More RIT/XIT for the Kenwood TS-530S

We all have our weak spots, and mine is CW DX. I don't rabidly chase DX, but when it's there I do my best to work it. So when I heard the XF4 expedition on 15 meters calling CQ and listening up 5 kHz, I found the frustration almost unbearable from not being able to tune that far from my transmit frequency.

## The problem

The TS-530S is a serviceable rig, but it has a range of only about ±1.5 kHz on its receiver incremental tuning (RIT) function. Because RIT is usually implemented with a varactor and pot for bias, I thought I'd try to find it on the schematic in the owner's manual and see if I could coax it into tuning farther. Stripped of a lot of extra lines on the drawing, the schematic is shown in **Figure 1**. Pot VR5 on the IF assembly sets the range midpoint, and the varactor in the VFO assembly is the actual component that changes.

The "cleanest" modification would be to change the varactor for one with a wider tuning range. This would be hard to do, though, as the VFO is a sealed assembly. A cheap and dirty alternative would be to increase the voltage swing to the varactor, and this is the method I used.

Not knowing anything about the particular diode used, I was concerned about increasing the diode voltage. A quick look at the schematic showed that this wasn't possible, because the full 9-volt supply is across the diode when the pot is at one end of its rotation. All I had to do was increase the voltage swing.

If I knew the varactor tuning curve, calculating the range would be easy. Lacking that, I clipped some test leads across the 6.8-k resistor to ground and paralleled some resistance until the tuning range was what I wanted. I settled on a 1-k resistor as the value to add in parallel to the existing part.

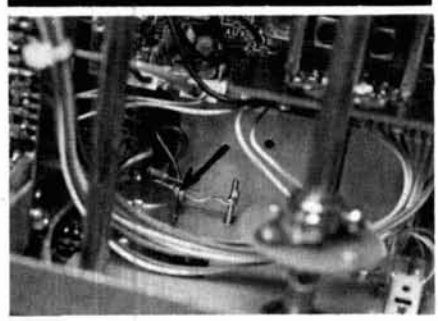
## Circuit changes

Installing the 1-k resistor (marked * in **Figure 1**) is the only thing tricky about this mod. First, I moved the existing resistor slightly out and down. Then I prepared the new part by forming a loop in one lead, tinning it lightly, and cutting the other lead to length. I looped the tinned end over the existing resistor, flowed the solder with my smallest iron, then wrapped the other end in place and soldered it. All of this inside work was made easier because I had a pair of hemostats* to do the fine positioning and holding. **Photo A** shows the completed modification.

Once the resistor is installed, you'll need to set VR5 so the frequency doesn't change when you turn on the RIT. Center the front panel pot and turn the RIT on and off. Note the frequencies with and without the RIT, and adjust VR5 so there's no change when you turn RIT on.

With the modification in place, the RIT tunes ±5.1 kHz. I was a little ner-

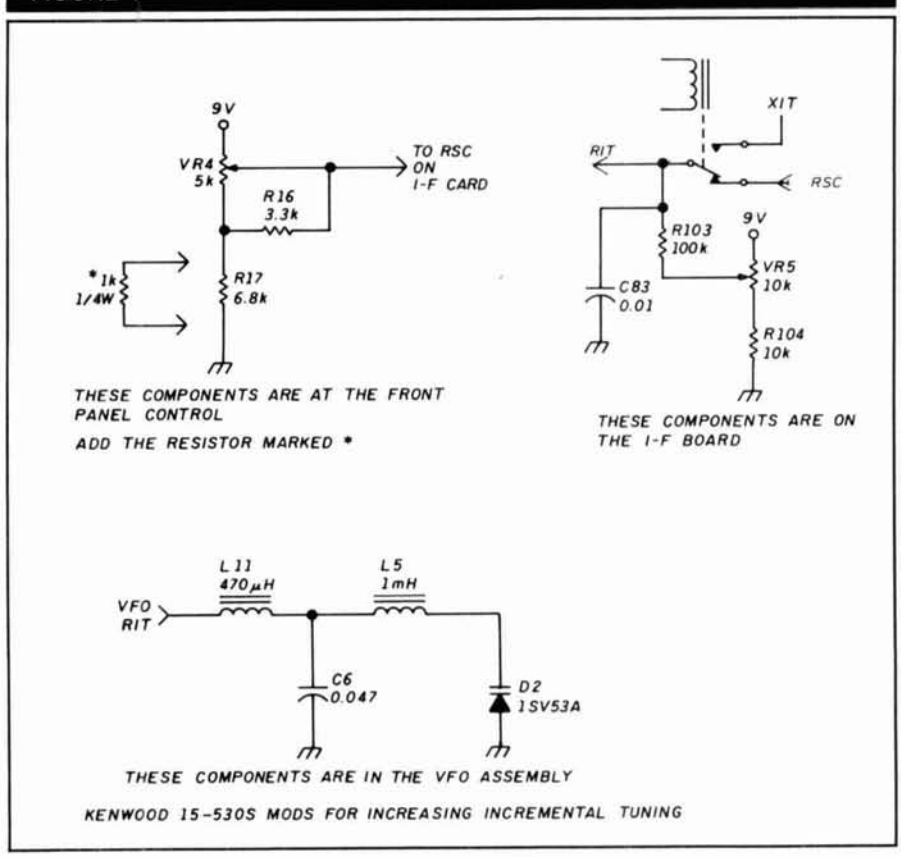
**PHOTO A**



**Changes to extend the Kenwood TS-530S RIT take about an hour.**

*A clamplike instrument used in surgery to reduce or prevent bleeding. Ed.

**FIGURE 1**




**Kenwood TS-530S mods for increasing incremental tuning.**

vous about trying for more range, because all but the most expensive varactors have a nonlinear tuning curve, and I wanted the resulting tuning range to be symmetrical on the RIT control.

This modification took about an hour from the time I decided to do it until the rig was buttoned up and operational. In case you were wondering, I worked the XF4 station a few hours later, both SSB and CW, with 5 kHz of RIT.

**Bob Lombardi, WB4EHS**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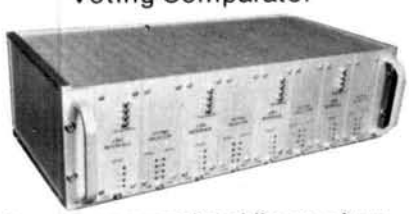

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| P50VD | 50-54 | <1.3 | 15 | 0 | DGFET | $29.95 |
| P50VDG | 50-54 | <0.5 | 24 | +12 | GaAsFET | $79.95 |
| P144VD | 144-148 | <1.5 | 15 | 0 | DGFET | $29.95 |
| P144VDA | 144-148 | <1.0 | 15 | 0 | DGFET | $37.95 |
| P144VDG | 144-148 | <0.5 | 24 | +12 | GaAsFET | $79.95 |
| P220VD | 220-225 | <1.8 | 15 | 0 | DGFET | $29.95 |
| P220VDA | 220-225 | <1.2 | 15 | 0 | DGFET | $37.95 |
| P220VDG | 220-225 | <0.5 | 20 | +12 | GaAsFET | $79.95 |
| P432VD | 420-450 | <1.8 | 15 | −20 | Bipolar | $32.95 |
| P432VDA | 420-450 | <1.1 | 17 | −20 | Bipolar | $49.95 |
| P432VDG | 420-450 | <0.5 | 16 | +12 | GaAsFET | $79.95 |
| **Inline (rf switched)** | | | | | | |
| SP28VD | 28-30 | <1.2 | 15 | 0 | DGFET | $59.95 |
| SP50VD | 50-54 | <1.4 | 15 | 0 | DGFET | $59.95 |
| SP50VDG | 50-54 | <0.55 | 24 | +12 | GaAsFET | $109.95 |
| SP144VD | 144-148 | <1.6 | 15 | 0 | DGFET | $59.95 |
| SP144VDA | 144-148 | <1.1 | 15 | 0 | DGFET | $67.95 |
| SP144VDG | 144-148 | <0.55 | 24 | +12 | GaAsFET | $109.95 |
| SP220VD | 220-225 | <1.9 | 15 | 0 | DGFET | $59.95 |
| SP220VDA | 220-225 | <1.3 | 15 | 0 | DGFET | $67.95 |
| SP220VDG | 220-225 | <0.55 | 20 | +12 | GaAsFET | $109.95 |
| SP432VD | 420-450 | <1.9 | 15 | −20 | Bipolar | $62.95 |
| SP432VDA | 420-450 | <1.2 | 17 | −20 | Bipolar | $79.95 |
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# SKYWAVE COMMUNICATIONS PART 2

## The atmosphere and the ionosphere

*By Cornell Drentea, WB3JZO*

According to Kenneth Davies,[1] Watson-Watt was the first to apply the term ionosphere to "that part of the atmosphere in which free ions exist in sufficient quantities to affect the propagation of radio waves." Other names associated with the ionosphere include Stuart (1878), Schuster (1889), Heaviside (1902), and Kennelly (1902). The discovery of how the ionosphere affects radio communications can also be attributed to the many Amateur Radio operators who have experimented with it during this century. The ionosphere can be a very hard thing to define due to its ever changing refractive and reflective characteristics. Although there is not complete agreement, it is generally accepted that the ionospheric region extends in the atmosphere from about 30 km (18.6 miles) above the earth's surface to almost 1,000 km (620 miles).

Assuming that its content remains constant with the altitude (a theory which is disputed), the general atmosphere is made up of the following gases:

| Gas | Volume (percent) | Mass (percent) |
|---|---|---|
| Nitrogen | 78.09 | 28.02 |
| Oxygen | 20.95 | 32.01 |
| Argon | 0.93 | 39.96 |
| Carbon dioxide | 0.33 | 44.02 |

Small amounts of water vapor, ozone, nitric oxide, and hydrogen also occur.

Within this concept, the atmosphere follows general meteorological behavior. Turbulent mixing exists up to 100 km (62 miles), with pressure and density variations behaving according to well-known barometric equations.

Some assume that above this altitude the atmosphere presents the same composition, but that its content is much more rarefied. Here, the ultraviolet rays and the x-rays coming from the sun generally produce the ionospheric reflective and refractive effects. During periods of disturbed sun, particle radiation — like high energy protons, alpha particles, and low energy protons and electrons — disturbs the ionosphere down to altitudes as low as 45 km (28 miles) above the surface. Under these conditions absorption of earth-generated waves takes place, so their energy will never reach the higher altitude reflective layers.

### Sunspots and flares

The mechanism of the sun causes the ionization of the rarefied gases when the radiation passes through the ionosphere. This gaseous globe of fire is 540,000 km (335,540 miles) in diameter. Its volume is one and a third million times the volume of the earth, while its density is only one-fourth that of the earth. The temperature at the center of the sun is believed to be approximately 7.2 million degrees Celsius (13 million degrees Fahrenheit). There are those who dispute this figure, believing that it is twice that much. At the surface of the sun, the temperature has been measured in the vicinity of 5,500 degrees Celsius (10,000 degrees Fahrenheit). The sun's hot gases are composed of dissociated atoms. Elements like hydrogen make up about 70 percent of the sun's mass, while helium makes up about 28 percent. The remaining 2 percent consists of all the heavier elements. Only a few hardy compounds hold together in the unspotted areas of the sun. The light and heat radiation received continuously on earth from the sun is the equivalent of about 1.5 kW per square meter. The sun has been outputting this energy for more than a billion years, and will likely continue to do so for several billion more.

The most important process on the sun impacting the ionosphere is the formation of sunspots. Sunspots are confined, eruptive masses of cooled gaseous plasma, which exhibit molecular (versus atomic) composition and move about the sun's surface in a swirling fashion. This happens to a greater extent every eleven years or so, creating what is known as the sunspot cycle. However, a certain number of spots exist on the surface of the sun at all times. To an observer, they appear in groups and move in a westerly direction across the sun's surface. Their speeds differ slightly, depending on their exact latitude. The sunspot rotation cycle is approximately 27.5 earth days (the sun's rotation period), making it easy to predict a rounded 28-day

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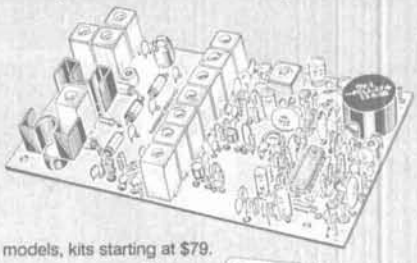

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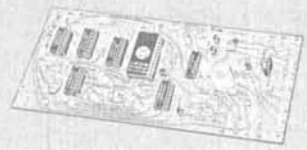

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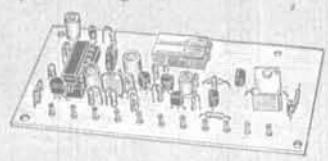

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propagation cycle. Sunspot diameters have been measured at up to 128,000 km (79,535 miles). The spots appear darker than the surface of the sun because of their lower temperature — 4,400 degrees Kelvin (4,126 degrees Celsius, or 7,460 Fahrenheit) compared with 6,100 degrees Kelvin (5,826 degrees Celsius, or 10,520 Fahrenheit). They are likely to appear in two rather narrow bands located equidistant from the sun's equator. Groups that are usually characterized by two larger spots accompanied by a number of smaller ones appear at latitudes of 5 to 30 degrees north and south of the sun's equator. Through selective spectrum analysis, the paired sunspots have been identified to contain reversed magnetic polarity from each other. For example, during a particular cycle the preceding spots in the Northern Hemisphere show their positive poles, while the following spots show their negative poles. The process is reversed in the Southern Hemisphere. An interesting phenomenon occurs when a new cycle begins; this magnetic mechanism appears to be reversed from the previous cycle.

Study of the molecular composition of sunspots has revealed that they contain about 18 compounds. These include titanium oxide and the hydrides (compounds of hydrogen with other elements) of calcium and magnesium. The sunspot's powerful magnetic fields have been measured at about 4500 gaus. They keep the colder matter confined within the spot's perimeters.

Explosive outbursts of matter and energy at the edges of the sunspots are called flares. These explosions are associated with sunspots, but not to be confused with them. Flares burst out in space from the edges of sunspots to an altitude of about 20,000 km (12,427 miles) at speeds of about 300 km per second (186 miles per second). The resulting radiation, composed of copious amounts of ultraviolet rays, x-rays, and cosmic particles, leaves the surface of the sun and travels out into space. Depending on its nature and, consequently, its speed, a certain amount of this energy arrives here on earth at various intervals. For instance, it takes approximately 8.3 minutes for the ultraviolet and x-rays to arrive in the ionosphere, 15 minutes for the high energy particles like alpha rays and protons, and 18 to 40 hours for the low energy protons and electrons. This latter type of radiation is also known as the solar wind.

## Ionospheric interaction

As flare radiation penetrates the earth's atmosphere, it interacts with the rarefied gases starting at about 600 km (373 miles) from earth. The interaction with the denser gaseous layers located closer to the earth's surface tapers off the flare's energy at about 30 km (18.6 miles) from the earth's surface where the ionosphere ceases to exist. According to a recent theory, the ionosphere can be thought of as a thick layer of disassociated atmospheric gases impacted primarily by ultraviolet radiation and x-rays. Because the ultraviolet radiation is not monochromatic, you can assume that certain gases, depending upon how they occur at different altitudes in the atmosphere, provide various degrees of absorption to certain wavelengths. This, in turn, creates several layers of ionization.[2]

Another current theory says that, contrary to appearances, the ionosphere is not made up of separate physical layers, but that the degree of ionization (the number of electrons per unit volume) is a continuous function of altitude, with maxima and areas of inflection. Therefore, we should talk about "ionized regions" rather than the more accepted layer terminology.[3] However, the word layer would be acceptable (because it has been used for many years) if defined as the altitude at which wave reflection takes place.

In any case, the ionization process consists of a breakup of neutral molecules and atoms into free electrons and positive ions. The free electrons recombine with the positive ions or attach to other neutral molecules, creating negative polarized ions. It is believed that the free electrons are responsible for the ionospheric reflections and refractions. If you accept the second theory, electron density is a function of the continuous equilibrium between the electron creation processes and the de-ionization processes. It depends on the recombination of the positive ions with the electrons, and the electron capture by other neutral atoms. Because of their heavy mass, it is believed that ionized atoms do not play a role in propagation.

The status of the ionosphere is always changing and depends on many things like sunspot activity, the magnetic field generated by the earth, the time of day, and even the winds in the upper atmosphere. Proof of some of these dependencies was demonstrated during ionosphere heating experiments at Arecibo Observatory when it was discovered that plasma waves were being generated in the ionosphere. This in turn accelerated the free electrons to high energies, making them travel from one hemisphere to the other. The experiments at Arecibo have led to speculation that certain new forms of electromagnetic waves can be generated through the modulation of these electric currents into the ionosphere, and that the process can be used for communication purposes.

To conclude, the ionosphere is comprised of several layers of dense ionization which are always changing and overlapping each other, but that can be defined distinctly enough to be studied individually. Now I'll look at what happens to HF radio signals arriving in the ionosphere.

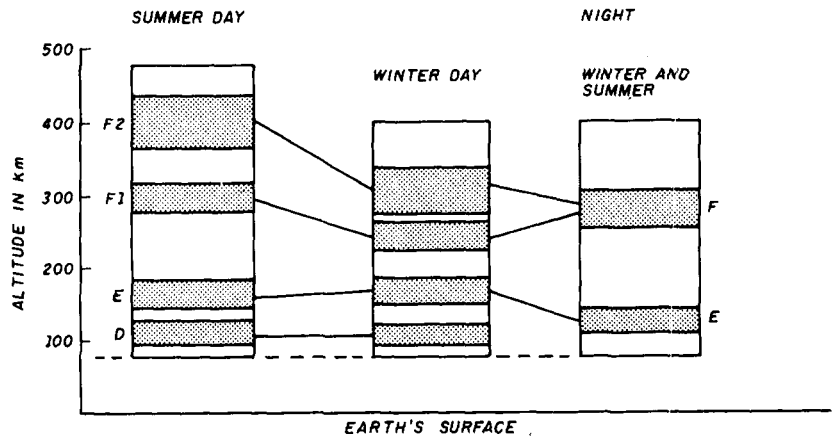
According to one theory, the electric field of an earth-generated HF wave arriving at an exited ionospheric layer causes the electrons to oscillate at the same frequency as the original signal. This oscillation produces a new electric field which is mathematically out of phase from the incident wave. The result is a perceived refraction or reflection of the wave at the ionosphere level. The transmitted signal can be received back on earth in the direction of the short or long paths at great distances. This is known as "skip." The regenerated (reflected) wave can bounce back and forth between the ionosphere and earth several times, making for audible and measurable delays over great distances (multiple hop propagation).

Some recent speculations suggest an even newer concept in long distance low power (QRP) communications. Some feel there is good reason to believe that the excited electrons in the ionosphere act not only as a reflector as is generally known, but as a plasma amplifier and a re-radiating antenna at the same time.* When conditions are right, a gain of anywhere from about 3 to 6 dB is believed to be obtained. The theory behind this mechanism is based on a lasing-like** reaction of the RF-excited mag-

* This theory was advanced by Albert Henderson, K6AJ, at RF Expo 1985 in Anaheim, California. It has not been validated by other laboratories, like the National Astronomy and Ionosphere Center, or by the Los Alamos National Laboratory.

** LASER stands for light amplification by stimulated emission of radiation; MASER stands for microwave amplification by stimulated emission of radiation.

FIGURE 1




**Altitude changes for D, E, F1, and F2 layers depending on season and time of day.**

netic fields, which in turn excites other electrons under the influence of the ultraviolet radiation in sort of a chain reaction.

## The ionospheric layers

I've discussed how the ionospheric layers form. I'll describe these layers and their properties only briefly, assuming that from reading other publications you are generally familiar with the position and properties of each layer.

Sir Edward Appleton was responsible for defining and classifying the ionospheric layers in 1924. According to Appleton, the most important layers for communications are the D, E, F1, and F2 layers. The first layer identified was named E for the word electric. The other nomenclature followed in place. **Figure 1** shows how these layers change altitude and combine, depending on seasonal and time of day parameters.

## The D layer and the lowest usable frequency (LUF)

The D layer is predominant between 55 to about 90 km (34 to 55 miles) and is a daytime event. At night the D layer disappears totally because the sun's radiation is stopped by the earth's shadow at these low altitudes. This layer is not very good at reflecting electromagnetic energy. Because gases are relatively dense at these altitudes, the ions produced in this layer encounter quick recombination processes (a lower number of free electrons) which produce energy losses rather than RF reflections and refractions.

In addition, the reflectance of this layer (if any) tends to be of a high angle because of its low altitude. Signal absorption occurs during the day at the lower HF frequencies. The frequency at which the D layer begins to allow HF waves to continue on to the higher reflective layers is known as the lowest usable frequency (LUF). This frequency varies with the sun's activity and can be shifted so high as to render a total blackout for all HF waves during periods of sunstorms.

## E layer and sporadic E propagation

The E layer extends from about 90 to 125 km (55 to 77 miles). With its lower air density and higher angle of reflectance, the E layer acts as a better reflector or refractor than the D layer for relatively short paths. Because of its higher altitude, the E layer tends to produce results from some time before sunrise to after sunset, with minimums around midnight. Therefore, the E layer is primarily a daytime layer. Sporadic ionized patches of up to 150 km (93 miles) in diameter sometimes tend to form at the bottom of the E layer (about 100 km or 62 miles), providing powerful reflective and refractive properties for waves up into the very high frequency (VHF) range. It is believed that this phenomenon is caused by the interaction of gaseous winds in the atmosphere with the earth's magnetic field. It is called sporadic E because it only lasts for a few hours and occurs randomly. When present, sporadic E is responsible for spectacular communications at distances of up to 1,000 km (621 miles) for HF and especially VHF frequencies. Sporadic E clouds appear to drift with the upper winds, generally in a westerly direction and at speeds of about 150 km per hour (90 miles/hr). This type of propagation occurs to a greater extent at lower earth latitudes, commonly 20 to 40 degrees. Up to 100 hours of sporadic E propagation has been reported yearly in the United States during the months of May through August.

## F layer, the maximum usable frequency (MUF), and the optimum frequency of transmission (FOT)

The F layer is used for long haul skywave communications. Its altitude is between 150 to 500 km (93 to 310 miles). Consquently, the F layer is almost always exposed to the sun's radiation. It is composed of two well-defined sublayers, F1 and F2. The F1 layer is the lower of the two, and extends from 150 to about 250 km (93 to 155 miles) during the day. The F1 layer behaves much like the E layer; that is, it achieves its maximum ionization at noon and disappears almost completely at night. Its altitude changes seasonally between 350 km (220 miles) during the day in the winter to about 500 km (310 miles) during the summer daylight hours. The nighttime altitudes are between 250 to 420 km (155 to 260 miles). Because of this, skip distance varies with seasons and time of day. Due to its height, the F2 layer is

always present but may be blocked to skywave communication by the absorption of the D and E layers. Efficient reflections and refractions can be obtained from the F2 layer year round.

The F2 layer is partly responsible for determining the optimum frequency of transmission. The art of predicting the FOT at any given time is dictated by complex calculations related to how the F2 layer impacts the maximum usable frequency (MUF) and how that relates to the lowest usable frequency (LUF) determined by the D layer. In practice, the FOT is taken as 85 percent of MUF. The MUF determination is also obtained from complex calculations based on standards set by the Comite Consultatif International des Radiocomunications (CCIR). Some of these standards, along with empirical data and exhaustive calculations, can be found in computer programs which readily predict propagation for the long term. They provide a means of planning radio communication systems and frequency schedules for months in advance while anticipating their variability under both quiet and disturbed conditions. Short term forecasting uses information derived from solar observations for day-to-day management of radio circuits. Long term forecasting takes into account the irregular behavior of the slow arriving particles known as the solar wind, and their variable interaction with the earth's magnetic field in the magnetosphere and the ionosphere. I'll discuss some of these programs in part 3 of this article.

## The grayline and auroral propagation

One of the more interesting propagation modes is the grayline. The grayline is a path created along a 500 km (310 miles) wide area separating day from night, usually called the terminator. Enhanced north-south propagation along this path is possible depending on the earth's diurnal tilt (arc varies up to 47 degrees relative to the poles in one year). The HF spectrum affected by this type of propagation is generally from 2 to 10 MHz. Simple computer programs (which I will not discuss here), that anticipate how this tilted line crosses different areas on earth, can predict with some certainty when and where communication can be established. In grayline propagation, the D and E layers are actively providing strong absorption to the waves being transmitted on the day side of the line. On the nightside these layers are nonexistent, with a MUF which is usually below the frequency of interest. The energy is then channeled up toward the reflecting layers by these two conditions, in much the same way that light would be radiated upward from inside a narrow canyon. The energy is finally reflected back to earth by the F layer on the short or long paths along this grayline. Grayline propagation is useful for HF communications.

Another interesting propagation phenomenon is based on the aurora. During a magnetic storm, the solar wind particles, and particularly the electrons arriving in the ionosphere, become trapped by the earth's magnetic field in the polar regions. Here is how the entire process happens. As the earth spins, its crust moves faster than its core. This is a motor-like mechanism responsible for creating the magnetic field which surrounds the earth. As discussed earlier, the magnetic field around the earth is known as the magnetosphere. The magnetosphere with its well-known Van Allen belts is impacted in turn by the solar wind. The interaction of the solar wind (for example, electrons) with the magnetosphere turns into electricity. It has been estimated that more than a million megawatts are being generated at any given time, making gases glow in the dark or become fluorescent. If you believe the earlier theory that different gases concentrate at different altitudes, then electrons spiraling down the magnetic lines of the magnetosphere in the oxygen region will glow red or green. At the lower levels, the nitrogen molecules will produce violets and blues. Other gases, like hydrogen (in lesser amounts) will produce oranges, and yet a different kind of red. This is the aurora. HF signals are generally absorbed by the aurora. A strange fading occurs making the signals flutter drastically at a high rate (arctic flutter). It is possible to bounce signals off the aurora curtains with occasional success. A new kind of propagation known as the Auroral E has been used recently at VHF frequencies. This is a mode which is similar to sporadic E, and has been used at frequencies of up to 144 MHz. hr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# Ham Radio Bulletin Board

**Dear Readers:**

Welcome to the *Ham Radio* Bulletin Board. The "Bulletin Board" is our new forum for sharing some of the more technical correspondence we receive here at **HR**, additional information on articles we have run, or any other pertinent bits of news. This new feature will appear in the magazine whenever we have messages for you, or you have messages for us. You'll find the "Bulletin Board" interesting, thought provoking, and informative.

**Terry Northup, KA1STC, Editor**

## Enhancing accuracy using insulating spacers

**Hear HR:**

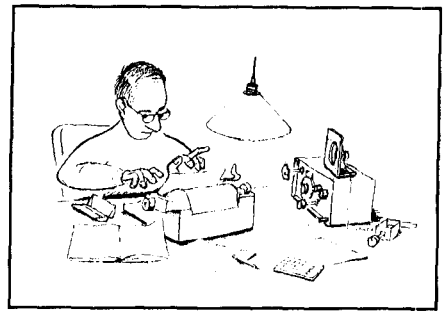
I really enjoyed the article in the January 1990 issue of *Ham Radio* on "Copper Pipe Transmission Lines." However, I would like to add a little information that will enhance the accuracy of the results.

The impedance values in **Table 1** or **Equation 1** are based only on air as the dielectric separating the inner and outer conductors. Many times it is necessary to use insulating spacers or beads in conjunction with air in order to maintain the concentric relationship between the conductors. When non-air or non-vacuum insulating material is introduced, this effectively raises the dielectric constant to some value greater than 1 (the value for air) and also effectively lowers the value of the characteristic impedance.

With this in mind, the following general rules should be used when insulating spacers are introduced with air dielectric:

1. Use the thinnest insulator spacer dimensions possible and the smallest number of spacers possible (the total spacer volume must be much less than the total air volume). Use low loss insulating material.
2. Always select a characteristic impedance value from **Table 1** that is slightly larger than the desired characteristic impedance.

**Hilton A. Turner, Jr., KB8LS, Kokomo, Indiana**

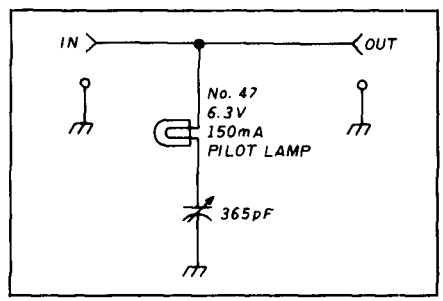


## Being creative at little expense

I found a great and pleasant bit of nostalgia in Charlie Tiemeyer, W3RMD, and his article in the December 1989 issue of *Ham Radio*. The article is on page 42 and is titled "The Five Band Junkbox Transmitter."

Many times during my Ham career I have whipped up a "rigg" of this type, *but* with one addition!

Taking tips from articles in *QST* (Hyde, April 1955, page 51, and McCoy, September 1956, page 22) I always added a simple output indicator. The sketch I've included is self-explanatory (see below). I have built it into the transmitter or separately in a coffee can.




To use, simply set to minimum capacity and tune the transmitter normally. Next, increase the capacity *slowly* until the lamp barely lights. Then *retune the transmitter for maximum* brilliance of the lamp. This represents maximum output even though the meter may not agree.

Be my guest to add this device to your readers' fund of knowledge.

Do issue more articles of this nature. The beginners to our craft need to know that a fortune is not necessary to enjoy Ham Radio!

**Gerald R. Skinner, K4LVZ, Winter Park, Florida**

## Theory and principle

**Dear HR:**

The article in the November issue ("A Low-Noise Design Primer," page 80) by Bob Lombardi, WB4EHS, on the subject of noise was very interesting but I would like to correct an historical reference.

Thermal noise in electrical circuits was discovered by J.B. Johnson in the course of research on telephone lines. The theoretical explanation of the noise (**Equation 1**, et sequence in Lombardi's article) was developed by H. Nyquist. Both Johnson and Nyquist published their work in the journal *Physical Review* in 1927 and 1928.

In physics it is common to refer to the noise, itself, as "Johnson noise" in honor of the discoverer. Nyquist's result is usually called the "Nyquist theorem" because he derived it from fundamental principles, specifically from the Second Law of Thermodynamics and the quantum theory of radiation.

It might interest your readers to know that the importance of Nyquist's result ranges far beyond the direct application of his theorem. His theory was generalized (in the early 1950s) to apply to such diverse phenomena as fluctuations in dielectrics and in magnetic materials, to scattering of electromagnetic radiation and neutron scattering. Even today, the Nyquist theorem and its generalizations are among the few exact theoretical connections between the science of equilibrium thermodynamics and electronic circuits (which never operate at thermal equilibrium).

**Walter A. Simmons, Ph.D, AH6HU, Honolulu, Hawaii**

## Article update

Here's an update to the article "Keeping an Eye on Your Sideband PEP," by John Fielden, GW5NAH (SK), that appeared in our September 1989 issue. The original article refers to balancing out op amp offsets but doesn't suggest how it might be done. For those of you not familiar with such techniques, the following method is suggested.

Apply a small DC voltage (about 2 to 4 mV) to the input of the module and

adjust RV3 to get the same voltage at the output. An accuracy of ±100 μV is quite sufficient. You can't make this adjustment with a zero input voltage as the output can't swing below about 1 mV. The settings of RV1 and RV2 don't matter for this adjustment, nor is the actual voltage used critical — as long as it doesn't exceed a few mV. Once the offset adjustment is completed, set RV1 and RV2 for the correct sensitivity as described in the article.

Probably the easiest way to get the small voltage needed is to connect the input of the module to the supply voltage via a resistor of 1 meg or so. You can use RV1 and RV2 to get a suitable voltage for the offset adjustment.

**Bob Wilson, WA1TKH,**
**Consulting Technical Editor**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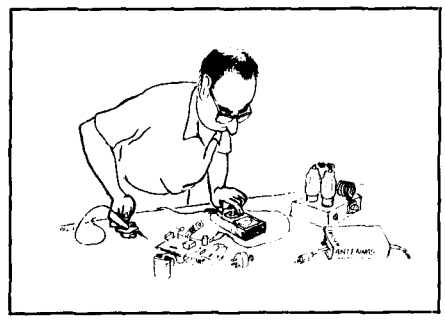


*Bill Orr, W6SAI*

## THE YAGI OPTIMIZER

The year was 1946. There was plenty of exotic DX on the 20-meter band: AC4YN, XZ2KN, W6VKV/I6, ET1JJ, AC3SS, YI5AL, and lots more. The problem was to work it. And more often than not, I was beaten out in a DX pileup by my good friend and aggressive competitor Paul, W6MJB, a local DXer with a keen sense for the jugular.

"The reason he beats you out is that he has a four-element beam and you only have a three-element beam," observed W6VFR.

"The reason I beat you out is that I'm a better operator than you," said W6MJB with an engaging smile.

Very frustrating. Ol' Paul had put the hex on me for sure. Did he have a better antenna? Was he a more experienced DX chaser than I? Or both? I never resolved that problem.

But that was part of the fun. When, on a rare occasion, I did manage to snooker W6MJB, it really made me feel good. Maybe my antenna was as effective as his, after all!

Much has happened since those good old days, and a lot of the Black Magic once invoked in designing and building beam antennas has now become a science. Here's a case in point. A few weeks ago I was looking through a dusty old log book and found the dimensions and data for W6MJB's mighty four-element DX antenna. I entered the pertinent information in my modern computer and compared the results with similar data about the old three-element beam I used when chasing Paul around the 20-meter band. The results? Both beams performed about the same as far as gain and front-to-back ratio were concerned! There was no discernible difference in overall performance between the two antennas. Maybe Paul was right — he was a better operator than I was!

### The Yagi Optimizer (YO)

Is a four-element beam a better performer than a three-element beam? What is the optimum boom length for a given number of elements? What is the effect of element taper? Is it better to lengthen the boom when adding an extra element, or to drop an extra element into an existing array in order to achieve a little more gain? What is the relationship between gain and front-to-back ratio? Is it necessary to sacrifice one to enhance the other? Do maximum gain and highest front-to-back ratio occur at the same frequency? These are good questions, ones that can finally be answered with the Yagi Optimizer program for the home computer.

The Yagi Optimizer Program* developed by Brian Beezley, K6STI, requires an IBM PC compatible computer with at least 512K of memory. You need a CGA, EGA, or HGC display. A math coprocessor is recommended, as is a hard disk, but neither is required. You should use DOS 3.00 or later to enable full access to all features of the YO program.

YO analyzes and optimizes a given Yagi antenna. It will model arrays of up to 50 elements. The model is examined in free space and the accuracy (compared with the MN program discussed last month) is typically within 0.1 dB for forward gain, a few dB for front-to-back, and within a couple of ohms for input impedance.

Optimization may be done at a spot frequency within a band, or at the low, middle, and high frequencies of the band of interest. You may choose the parameters, aiming for maximum gain, best front-to-back ratio, a given value of input impedance, or a combination of these parameters.

### Is there a maximum gain Yagi?

Before discussing program details, it's interesting to contemplate the possibility of achieving a maximum gain Yagi. A popular question among HF and VHF operators is: How much gain can be achieved for a given number of elements or a given boom length? YO will give interesting answers to such questions.

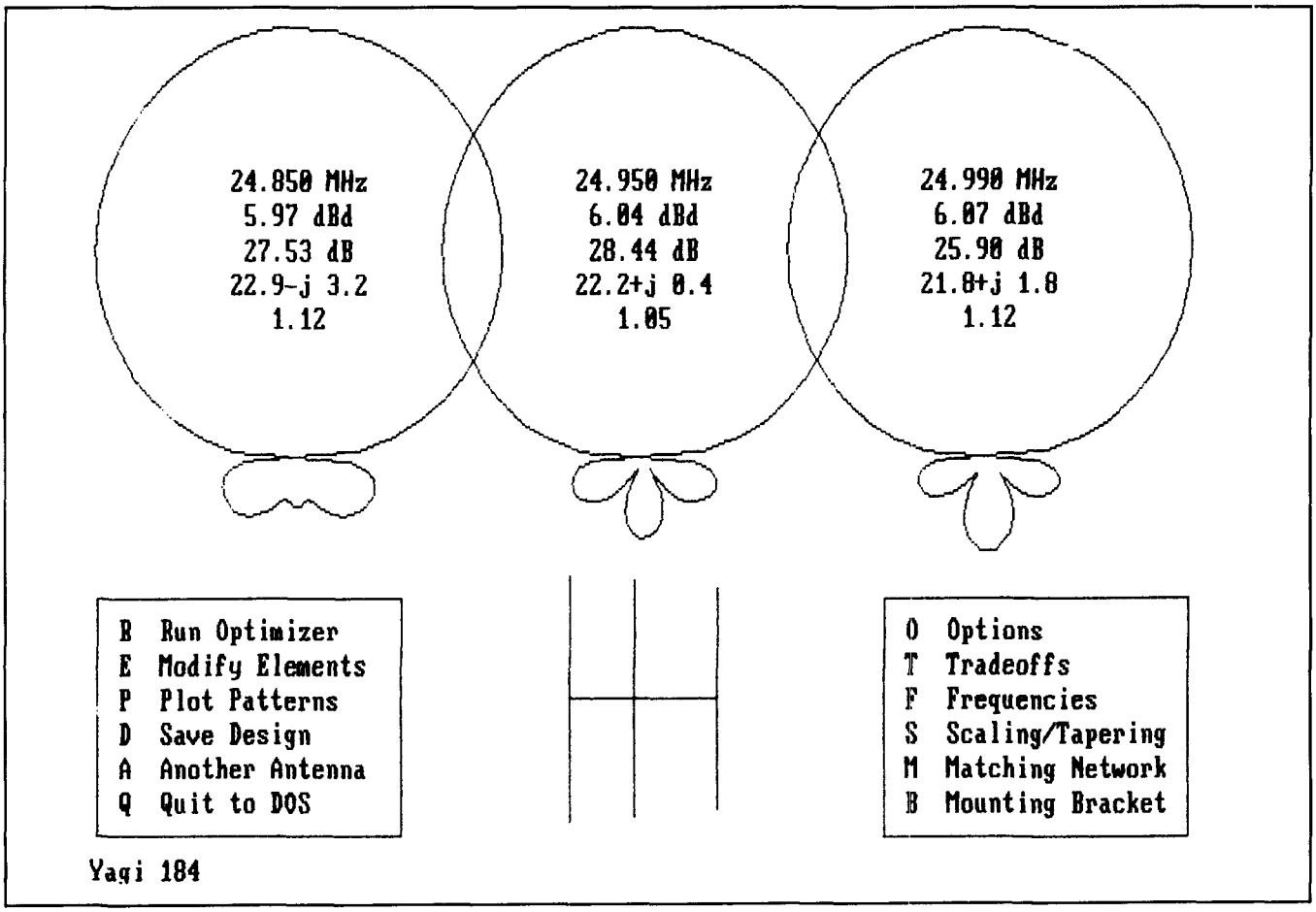
Yes, Virginia, there is a maximum gain Yagi. YO will find it for you by crunching through variations in element length and spacing that would be impossible to compute a decade ago. For example, I entered typical dimensions for a 24-MHz three-element Yagi built on a 12-foot boom into the computer. Starting with these basics, the program was directed to search for a maximum gain Yagi that retained a good front-to-back ratio while maintaining the same boom length. The results are shown in **Figure 1**. The program examined 184 length/spacing combinations to arrive at the optimized design. The array has 6.04-dBd gain at the design frequency of 24.95 MHz, and a front-to-back ratio of over 28 dB. Input impedance is about 22 ohms.

This is a good compromise design. Can the gain be increased by holding the boom length constant, but allowing element length and spacing to be varied? Yes. Setting YO in action again, I found that the program quickly examined 262 designs in sequence and arrived at the maximum gain design shown in **Figure 2**. This Yagi provides 7.82-dBd gain with a front-to-back ratio of only 7.55 dB. Input impedance is about 1.7 ohms. And look at the spacing!

A beam of this design is a theoretical concept. The radiation resistance is so low that substantial ohmic losses exist

* Brian Beezley, K6STI, 507-1/2 Taylor, Vista, California 92084.

FIGURE 1




**24-MHz Yagi optimized for good combination of gain and front-to-back ratio. The figures in the patterns indicate frequency, gain, front-to-back ratio, input impedance, and SWR. Beam is optimized at three frequencies. Below the patterns you see a representation of the Yagi. Monotaper dimensions are (length and position in inches): *reflector*, 118.25/-72.96; *driven element*, 112.75/-9.23; *director*, 105.8/72.96. Boom length is 12 feet 2 inches.**

in the array, and it's difficult to design a network that would match the low impedance load to a coax line. The real life gain of such an oddball design is questionable. Front-to-back ratio is poor and wind vibration would upset antenna parameters unless the driven element and reflector were suitably braced.

What's the upshot of this optimization? The design has gained 1.78 dB at the loss of nearly 21 dB of front-to-back ratio. At the same time, the feedpoint impedance has become impossibly low. Exotic results, but the beam is impractical. Optimization has been carried to an extreme.

## Practical results with the YO program

Running the YO program quickly shows the folly of being too eager to achieve forward gain at the expense of other important antenna parameters. The YO program emphasizes that maximizing forward gain results in poor front-to-back ratio, low input impedance, possible sidelobes, and small SWR bandwidth. But because of the tradeoff capability, the program lets you automatically optimize a combination of gain, front-to-back, and either input impedance or SWR. The combination of parameters to be optimized and their relative importance are determined with a Tradeoff Menu.

## The Tradeoff Menu

In most cases, maximizing forward gain at the expense of other parameters isn't practical. To obtain a good combination of parameters, set a ratio of forward gain to front-to-back of 9:1. This means that YO will weigh front-to-back by 10 percent and forward gain by 90 percent. (Other ratios may be chosen at your discretion.) This ratio, however, places 1 dB of gain equal to 9 dB of front-to-back ratio. If YO can change the design in such a way that the front-to-back ratio increases more than 9 times as much as the forward gain decreases, it will do so.

Experience has shown that this ratio yields a very practical design. Once this optimization has been completed, you can then optimize the design for input impedance and bandwidth tradeoffs. Your strategy is to obtain the most uniform antenna performance across the band. Obviously, the final design depends considerably on the width of the Amateur band; a 10-meter optimized design is quite different from a design optimized for the narrow 12-meter band.

## The optimization method

The Yagi Optimizer uses a modified "Method of Steepest Descent." Readers with a calculus background

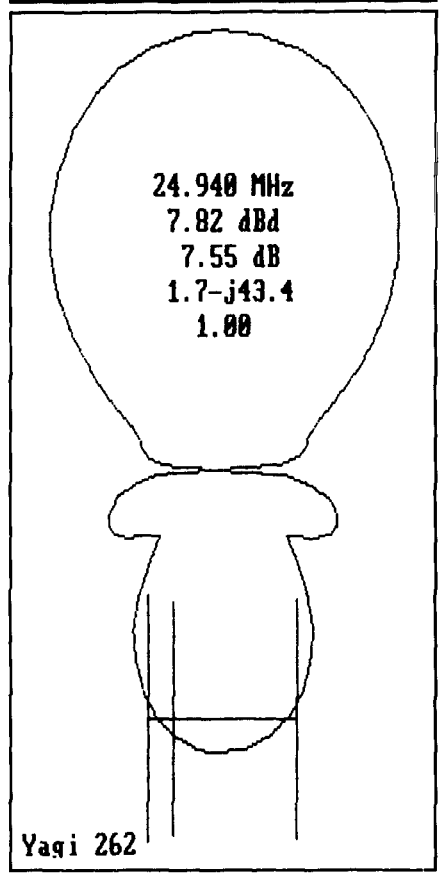
should recognize the technique immediately. Parameters of each element are changed by a small amount, while other antenna dimensions are held unchanged. The program calculates the sensitivity of the objective to each variable. The collection of the element sensitivities leads to an iteration where all the variables are updated, each in proportion to its respective sensitivity. The process is repetitive, and terminates either at user command or when no further improvement in the objective is possible.

This technique doesn't guarantee the very best set of antenna dimensions out of all possible dimensions. That nirvana can be accomplished only by an exhaustive and impractical search over all possible dimensions. The definition of "best" is user defined. As an analogy, imagine that you're climbing to the very top of a mountain range of multiple peaks, in a dense fog. You might arrive at a peak, but it might not be the highest one. Familiarization with the program and examination of various antennas in the program library will quickly guide you into a comfortable relationship with program tradeoffs. The program lets you change element lengths and positions by keyboard entry. This is a great way to get the "feel" for how each element affects the overall antenna performance. Common sense is a great help in Yagi optimization.

## YO matching networks and other stuff

YO contains models for several matching systems including the Gamma Match, the T-match, the "Hairpin" Match, and the Beta Match. The program allows you to determine the effects of element taper and mounting brackets. It's also possible to quickly scale any Yagi design to a new frequency, while maintaining essentially the same performance characteristics. Finally, the program provides pattern plots which may be printed using a dot matrix printer with Epson compatible graphics. There are other interesting "bells and whistles" in this program, but I won't cover them here.

FIGURE 2




**Same Yagi as in *Figure 1* optimized for maximum gain without regard to front-to-back ratio. Beam is optimized at only one frequency. Monotaper dimensions are (length and positon in inches): reflector, 115.251/0; driven element, 108.008/23.663; director, 111.006/142. Boom length is 11 feet 10 inches.**

TABLE 1

YO format for 4-element beam.

K7HYR's maximum gain Yagi
24.890 24.940 24.990 MHz
4 elements, inches

| | | | | |
|---|---|---|---|---|
| | 1.617 | 1.250 | 1.125 | 0.875 |
| 0.000 | 2.938 | 15.062 | 66.000 | 33.305 |
| 124.000 | 2.938 | 15.062 | 66.000 | 28.248 |
| 248.000 | 2.938 | 15.062 | 66.000 | 26.815 |
| 372.000 | 2.938 | 15.062 | 66.000 | 28.313 |

## The YO file format

The initial antenna configuration must be entered in a Yagi file in a specific format. The elements are specified in order, beginning with the reflector (see **Table 1**). One-half of each element, from the center of the boom to the element tip, is normally required. For monotaper elements, you may use the whole element length.

You place a title for the antenna on the first line. One or three analysis frequencies go on the second line. The program assumes you are using megahertz if you don't specify frequency units. On the third line enter the number of elements and the dimension units. In this example, inches are used.

The next line asks that you list the taper diameters, starting with those closest to the boom. The maximum number of taper sections you can use is seven. The first taper section in this example represents the element mounting plates.

On the remaining lines, list the element positions followed by the length of the taper sections, beginning with the section closest to the boom. (As an alternative, you may use element spacings instead.)

It's possible to accommodate different taper schedules for different elements. Various examples are given in the documentaion that accompanies the program.

*It's important to note that if you start out with a poor Yagi and optimize it, you'll get an optimized, poor Yagi. A modicum of common sense is vital in running a program of this type.*

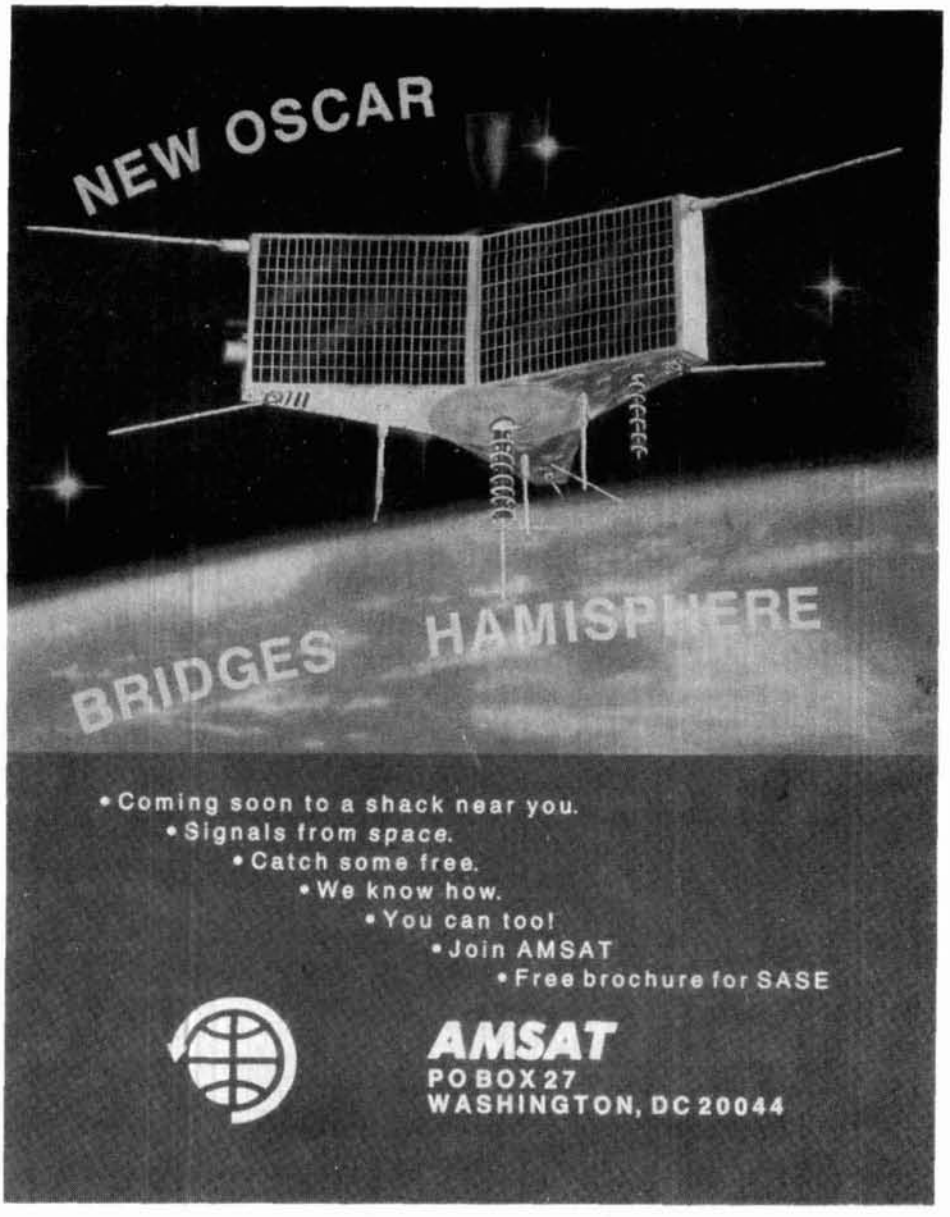
Used together, the MN and YO programs let you design and fine tune a Yagi array. You'll end up with a design that can be built, placed in final position, and used with a minimum amount of top-of-the-tower adjustment. I like that. As I've said before, I'm a coward when my feet are off the ground.

## A final word

Here are some notes of caution. These antenna analysis programs and other programs of this type require that you have a good working knowledge of MS-DOS or PC-DOS. The programs aren't tutorials for running the complex data operating systems used by the IBM-type computers. Most of my early cockpit problems with antenna programs were due to poor insight into DOS management. Once I overcame

this hurdle, it was smooth sailing from then on. I'd also like to note that IBM clones are equal, but some are more equal than others. You may find occasional pitfalls when using a particular clone with an idiosyncracy that doesn't quite match the IBM machine the author used to prepare the program. Luckily, these "glitches" can be worked out as time goes by. In my case, a suggested *print command on a different antenna* program refused to provide a readable copy. A friend who is a knowledgeable computer programmer provided me with the solution in a few moments.

I believe that antenna modeling is a powerful tool for Amateur Radio. From time to time I hope this column can offer antenna designs that have been refined by computer analysis for the best possible operating characteristics. *Antenna analysis is here to stay,* but don't throw away that SWR meter just yet! hr


**YAGI OPTIMIZER**

The YO program automatically optimizes Yagi dimensions for maximum forward gain, best pattern, and minimum SWR. YO updates radiation patterns at the central design frequency and band edges during optimization. A scale drawing of the Yagi changes shape as the design proceeds. YO is extremely fast, and can compute several trial designs per second. After optimization, high-resolution patterns may be displayed or printed in several formats. YO includes a powerful gain-F/B-SWR trade-off mechanism, minimization of all sidelobes, frequency scaling, and full EGA color. YO has models for gamma, T, hairpin, and beta matches, element tapering, and element mounting plates. A library of Yagi designs and extensive documentation are included. YO is intuitive, graphical, and highly interactive.

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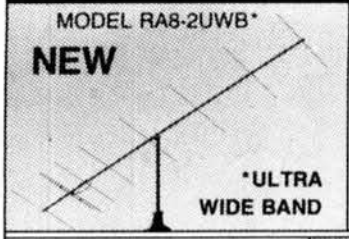

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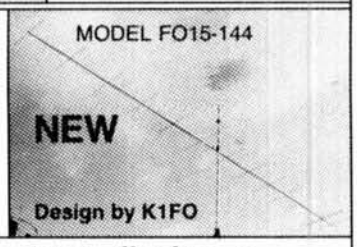

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# THE SUPER-DISCHARGER

## A simpler, easier to build test load for your NiCds

*By W.C. Cloninger, Jr., K3OF, 4409 Buckthorn Court, Rockville, Maryland 20853*

In a previous article[1] I discussed testing NiCds (nickel cadmium batteries) to determine their useful capacity. The process consists of charging the NiCds fully and then discharging them into a load at the mAh-times-1 rate. The multipurpose resistive load described[1] works well but requires close attention and several adjustments to maintain a reasonably constant load. Several full charge, full discharge cycles (1.0 volt per cell nominal minimum discharge voltage) can substantially improve the capacity of many NiCds.

During many NiCd discharge cycles, I've often wished that I had a constant current load that didn't vary with decreasing voltage. As a first attempt at a constant current load, I short circuited an existing current-limited power supply that uses a 723 regulator. It showed promise, but the input voltage versus load wasn't nearly as flat as I had hoped. This was probably due to the IC and the number of associated components. Then I discovered that it was best to keep things uncomplicated.

### Enter the Superdischarger

The circuit is simple, infinitely variable within its design range, smaller than the resistive load, doesn't require the special large resistor used in my earlier project,[1] and has a reasonably flat input voltage versus current load. The Superdischarger is shown in **Photo A**. It was specifically designed to discharge AA NiCds at 500 mA and sub-C, C, and D cells at 1.2 A (1200 mA). The load range of the Superdischarger extends from about 480 mA to just over 1.5 A.

### Theory of operation

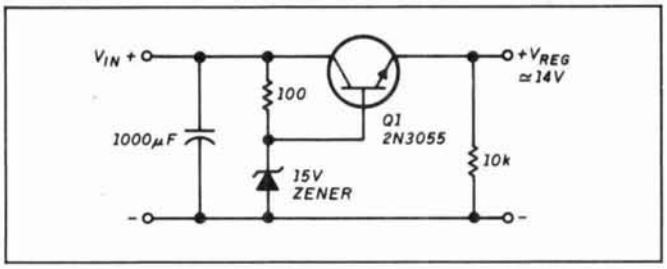
The Superdischarger is a current-limited power supply with an output load that will cause current limiting. A dead short of the output is the easiest "load" I could think of.

First, here's a quick lesson in one method of power supply regulation.[2] **Figure 1** shows a basic zener-regulated

**PHOTO A**

**The Superdischarger. The meter really indicates 0 to 2 A (but the price was right!). The two knobs are for coarse and fine-load adjustment.**

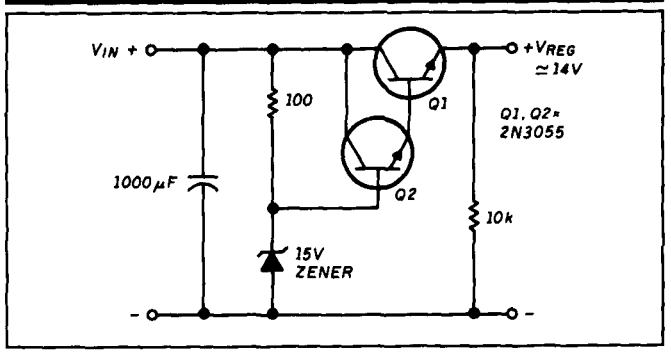
**FIGURE 1**




**A basic zener-regulated power-supply circuit. It is not good for current loads over about 2 A.**

power supply circuit using a single pass transistor. The maximum current output of this circuit is a couple of amperes, due to component losses and the gain factor of Q1. **Figure 2** is the same basic design; however, it uses a Darlington pair for the pass transistor. This circuit is good for maybe 5 A, but neither of these circuits has current limiting. I have seen both of these circuits used in commercial power supplies, usually less expensive ones.

**FIGURE 2**




**An improved zener-regulated circuit using a Darlington pair, which improves usable current output to about 5 A.**

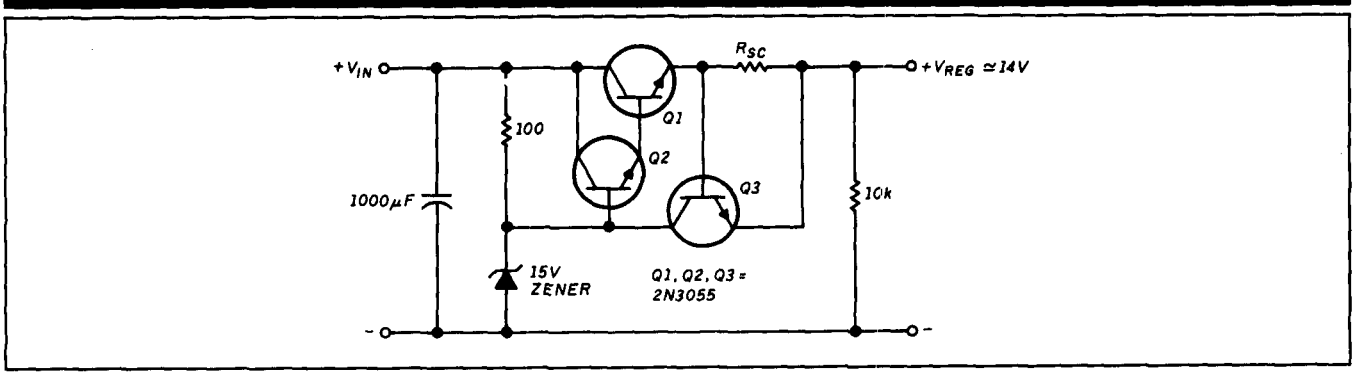
**Figure 3** shows how current limiting is added to either power supply circuit. When the voltage drop across Rsc reaches about 0.65 volt, Q3 begins to conduct and removes current from the base of Q2, reducing the output to a fixed level. This is the method of current limiting used in the Superdischarger.

These circuits have been published many times and are suitable for some voltage regulation applications. Better power supply circuits may use the LM723 voltage regulator[3] (Astron, Drake P7, and others). Don't forget the "three-legged" fixed or variable regulators if they suit your needs.

## Back to the Superdischarger

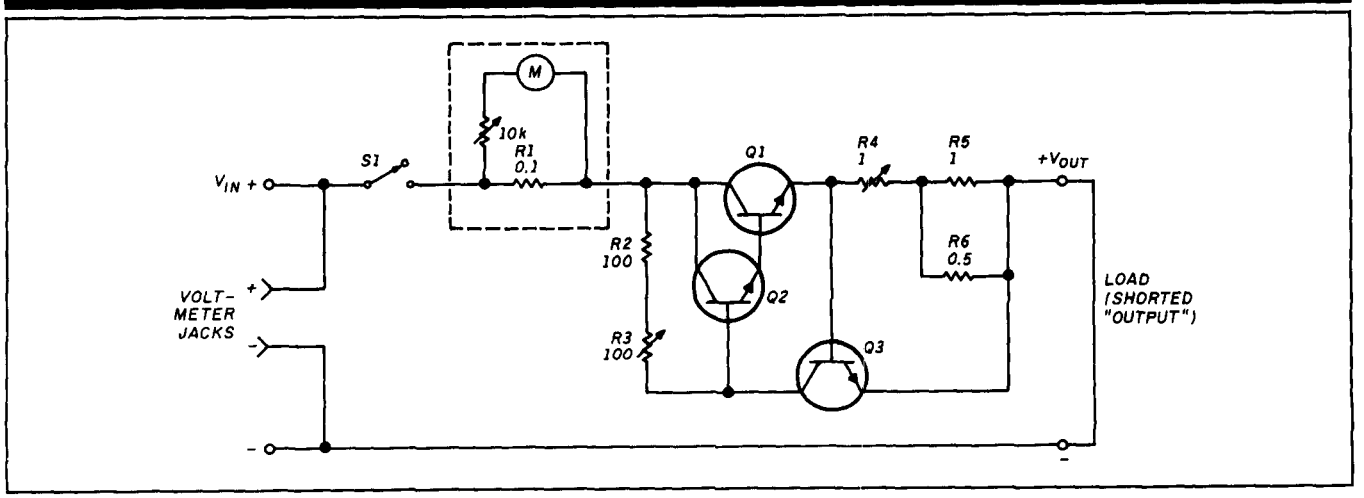
Because I needed only current limiting, I didn't use the zener diode. I now have an unregulated but current-limited "power supply" (see **Figure 4**). As the input voltage will vary with the voltage of the NiCd battery pack tested, voltage regulation isn't desired. My final design uses the Darlington pair because it will provide a 500-mA load down to 2.0 volts (2 AA cells). It won't maintain a 1.2-A load at 2.0 volts. You can use a single pass transistor, but the minimum voltage for a useful load is closer to 4 volts. There is one possible "advantage" to the single pass transistor version. You can

**FIGURE 3**




**Current limiting is added to the power supply circuit with Rsc and Q3. All transistors are 2N3055, but only Q1 requires a heat sink.**

**FIGURE 4**




**The Superdischarger circuit will provide a 500-mA to 1.5-A load over a range of 3 to 15 volts. Potentiometer R3 is optional, but permits fine load adjustment. A 0 to 2 A meter would eliminate the meter network (0.1-ohm and 10-k resistors) and the 200-μA meter. S1 turns the load on and off.**

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Sensitive superhet FM receiver tunes any 5 MHz segment of band. Listen to ham operations, high band police calls, weather or mobile phone calls! Easy to build receiver features varactor tuning, IC mixer stage, ceramic IF filters and dual conversion design with adjustable squelch. Less than 1 μv sensitivity, runs on 9 V battery, with 50 mw audio output. Optional ABS case with screened graphics and machined aluminum knobs provide a nice professional look.

| 2 MTR kit FR-2 | 10 MTR kit FR-10 | 220 MHz kit FR-20 | Receiver case kit CFR-2 |
|---|---|---|---|
| $29.95 | $29.95 | $29.95 | $12.95 |

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## BROADBAND PREAMP

A sensitive all purpose preamp, ideal for scanners, TV sets, VHF, UHF rigs, counters, etc. Features low noise, 4 db NF, 20 db gain, 100 KHz—1 GHz operation. Runs on 9—12 VDC, 50 ohms input.

Complete kit SA-7 **$14.95**

## LIGHT BEAM COMMUNICATORS

Transmits modulated infrared light up to 30 feet without lenses, up to 1/4 mile using lenses. Uses 30 KHz carrier for hum free operation. Transmits thru windows, etc. Ideal for "bugs" or listening to IR remote controls. Transmitter has sensitive mike input, receiver uses PIN detector and drives speaker output. Units operates on 9—12 VDC.

Transmitter kit LB-6 **$8.95**
Receiver kit LB-5 **$9.95**

## HIGH POWER FM WIRELESS MIKE

A high power unit that will transmit up to 1/2 mile to any FM broadcast radio. Sensitive input accepts any type of mike, will pick up normal voices 10 feet away using the available mini-electric mike cartridge. Operates on 9—12 VDC.

FM-4 kit **$12.95**
Sensitive microphone cartridge **$2.95**

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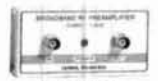

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wired includes AC adapter
PR-2 kit **$39.95**

The PR-2 is ideal for measuring weak signals from 10 to 1,000 MHz • flat 25 db gain • BNC connectors • great for sniffing RF • ideal receiver / TV preamp • 3 db NF.

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The PS-2 is handy for high resolution audio resolution measurements, multiples Up in frequency • great for PL tone measurements • multiplies by 10 or 100 • 0.01 Hz resolution & built-in signal preamp/conditioner.

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ML-1 Kit **$8.95**

**VOICE ACTIVATED SWITCH**
Voice activated switch kit provides switched output with current capability up to 100 mA. Can drive relays, lights, LED or even a tape recorder motor. Runs on 9 VDC.
VS-1 KIT **$6.95**

**VIDEO MODULATOR**
Converts any TV to video monitor. Super stable, tunable over ch 4-6. Runs on 5-15V, accepts std. video signal. Best unit on the market! Complete kit JM-7 **$12.95**

**LED BLINKY KIT**
Alternately flashes 2 jumbo LEDs. Use for name badges, buttons, warning panel lights. Runs on 3 to 15 volts.
BL-1 Kit **$3.95**

**MAD BLASTER**
Produces LOUD ear shattering and attention getting siren-like sound. Can supply up to 15 watts of obnoxious audio. Runs on 6-15 VDC.
MB-1 Kit **$4.95**

**UNIVERSAL TIMER**
Provides the basic parts and PC board required to provide a source of precision timing and pulse generation. Uses 555 timer IC and includes a range of parts for most timing needs.
UT-5 Kit **$5.95**

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Complete kit SS-7 **$29.95**
Case kit CSS-7 **12.95**

## CT-70 7 DIGIT 525 MHz

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## CT-90 9 DIGIT 600 MHz

**$169.95** WIRED INCLUDES AC ADAPTER

## CT-50 8 DIGIT 600 MHz

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## CT-125 9 DIGIT 1.2 GHz

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| MODEL | FREQ RANGE | SENSITIVITY | ACCURACY | DIGITS | RESOLUTION | PRICE |
|---|---|---|---|---|---|---|
| CT-70 | 20 Hz-550 MHz | < 50 mv To 150 MHz | 1 PPM | 7 | 1 Hz, 10Hz, 100Hz | 139.95 |
| CT-90 | 10 Hz-600 MHz | < 10mv To 150 MHz<br>< 150mv To 600 Mhz | 1 PPM | 9 | 0.1Hz, 10Hz, 100 Hz | 169.95 |
| CT-50 | 5 Hz-600 MHz | LESS THAN 25 mv | 1 PPM | 8 | 1Hz, 10Hz | 189.95 |
| CT-125 | 10 Hz-1.25 GHz | < 25mv @ 50 MHz<br>< 15mv @ 500 MHz<br>< 100mv @ 800 MHz | 1 PPM | 9 | 0.1Hz, 1Hz, 10Hz | 189.95 |
| CT-90 WITH OV-1 OPTION | 10 Hz-600 MHz | < 10mv To 150 MHz<br>< 150mv To 600 MHz | 0.1 PPM | 9 | 0.1Hz, 1Hz, 10Hz | 229.90 |

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|---|---|
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even short circuit Rsc and the whole circuit will draw only about 2.5 A at 14 volts. Remember what I said about Figure 1 having a limited regulating capacity. I'm not sure that a single pass transistor version could ever self-destruct if a reasonable amount of heat sinking is used. The power transistors used are so cheap, why not use the Darlington pair?

## Construction

Drill the project box and cut holes for the components. Pop rivet a terminal strip to the bottom of the box. You'll use this to support several of the components.

Mount pass transistor Q1 directly on the rear of the project box. I found that no additional heat sink was required (see Photo B). The Superdischarger does get warm, but not hot, for a 1.2-A load at 13.8 volts (about 16 watts of power dissipation). Solder Q2 directly to Q1; no heat sinking is required. Now solder Q3 to the terminal strip along with a couple of resistors and connecting wires as shown in Photo C.

The meter was a hamfest special. It senses the drop across R1 to give the proper current indication. R2 is 2 watts or larger; R3 and R4 are 5-watt wirewound potentiometers. They cost about $10 each, new. My meter was large, so I had limited room to mount the potentiometers on the front panel, and used the miniature Mallory VW series pots. I suggest using a 2 or 3-watt wirewound pot. Normal sized pots are about $5. I recently found several at a hamfest for $1.50 each. R3 is optional and provides the fine adjustment of the load. R4 provides all the load adjustment you really need, but I'm just a fanatic!

Provide input connectors to suit your needs. I added tip jacks to the back panel. I plug the leads from my voltmeter into these jacks to check the voltage during the discharge process. The project box from Radio Shack has a nice steel top, but the aluminum chassis is quite thin. The Cinch-Jones connector I chose requires considerable effort to use, so I pop riveted a piece of aluminum angle stock to the top and attached the rear panel to it with sheet metal screws (see Figure 5). The top now acts as part of the heat sink.

## Use and operation

Connect your NiCd battery to the Superdischarger, monitor the voltage, switch on the load, and adjust R4 to

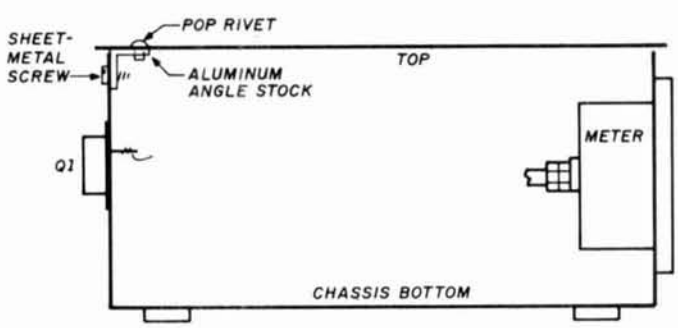
**PHOTO B**

**Back panel showing input jack and meter jacks. Q1 is mounted directly to the chassis (insulated if you wish) and requires no additional heat sink.**

**PARTS LIST**

| | Radio Shack part no. | Price |
|---|---|---|
| *S1—SPST switch* | | |
| *R1—0.1 ohm (optional depending on meter)* | | |
| *R2—100 ohm, 10 watt* | *271-135* | *$ 2/0.99* |
| *R3—100 ohm wirewound pot (optional)* | | |
| *Mallory VW-100 (5 watts)* | | *about 10.00* |
| *Clarostat 58C1 (100 ohms, 3 watts)* | | *about 5.00* |
| *R4—1 ohm wirewound pot* | | |
| *Mallory VW-1 (5 watts)* | | *about 10.00* |
| *Clarostat 58C1 (1 ohm, 3 watts)* | | *about 5.00* |
| *R5—1 ohm, 1 watt* | *271-131* | *2/0.99* |
| *R6—0.5 ohm, 1 watt* | *271-139 (0.47 ohm)* | *0.49* |
| *Q1—2N3055 NPN, TO-3 case* | *276-2041* | *1.99* |
| *Q2, Q3—TIP3055 NPN, TO-220 case* | *276-2020* | *1.59* |
| *Metal cabinet* | *270-253* | *5.69* |
| *Terminal strip* | *274-688* | *4/0.89* |
| *1-inch knobs* | *274-416* | *4/1.79* |
| *Meter and hardware to suit your needs* | | |

**FIGURE 5**




**Sectioned view of the project box shows the aluminum angle stock and sheet-metal screws used to secure the top to the back panel for added strength and heat sink.**

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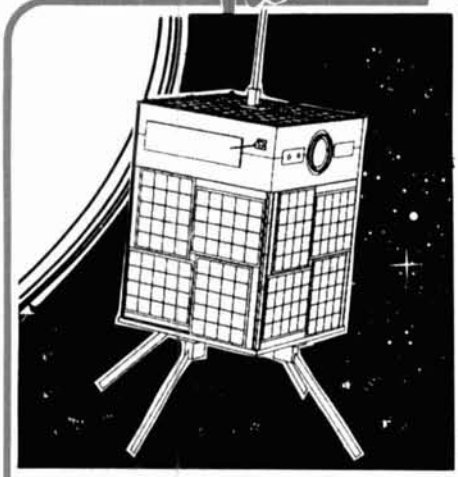

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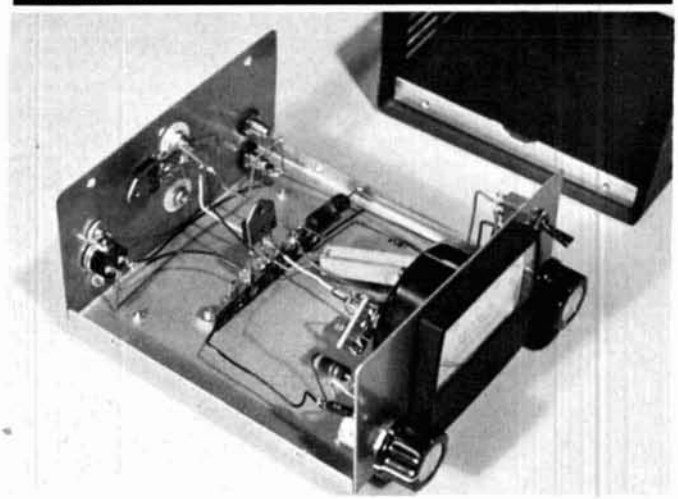
**PHOTO C**



**Simple point-to-point construction is used. Q2 is mounted directly to the terminals of Q1. Q3 is mounted on the terminal strip.**

500 mA or 1.2 A as needed. Make slight adjustments to keep the load constant as the battery voltage drops. Calculate the capacity of your battery by multiplying the time (hours) by discharge rate. If your AA cells give 500 mA for 1 hour, their capacity is 500 mAh or probably 100 percent of what they're rated (I won't discuss the 1-hour versus 10-hour ratings here).

I find that I'm easily distracted during the discharge cycle and tend to forget about the battery, so I use a kitchen timer and set it to go off every 5 minutes. I can then adjust the load and log the voltage and time, so I can plot a discharge curve if I wish. I also check and log individual cell voltages if I'm not discharging a sealed NiCd pack.

Now if I had a Superduper Discharger, which would automatically turn off the load at a preset voltage, log the time, graph the results... hr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3. W.C. Cloninger, Jr., K3OF, "723 Voltage Regulators," *Ham Radio*, March 1989, page 42.

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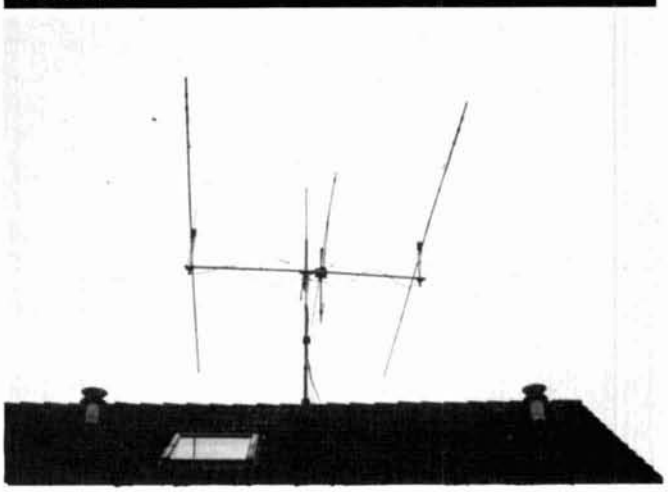
# OUTSTANDING 144-MHZ EME ACHIEVEMENT

## Over 100 initial EME contacts with a two-Yagi station

*By Bill Takacs, KB8ZW, 16724 Snyder Road, Chagrin Falls, Ohio 44022*

PHOTO A



**Twin 16-element Yagis perform very well for PAØJMV's EME experiments.**

On a recent trip to Holland I had the pleasure of visiting PAØJMV's remarkable 2-meter, two-Yagi earth-moon-earth (EME) station. What Joe Mutter has done with two Yagis is almost unbelievable. Looking over his QSL cards and logs made me a believer. Joe has managed over 100 initial EME contacts with his two-Yagi station. He has worked 67 countries, 28 states, 55 four-Yagi stations and 2 two-Yagi stations. **Photos A** through **D** show this remarkable EME station.

### Antennas

Joe has kept to basics, and the numbers on VK3UM's EME program will demonstrate why this is possible. His antennas are modified homemade 16-element arrays, patterned after KLM's, stacked horizontally at the peak of his roof. Nineteen and one-half feet of coax from each antenna goes to a quarter-wave power divider and to a preamp coax relay. Another 20 feet of coax goes from relay to hamshack, giving 0.7-dB loss into the transmit coax. The preamp is part of a coax relay, with the connector on the relay incorporated as the input circuit of the preamp. He also uses a 70-dB isolation relay. The preamp measures 0.2 noise figure with a 3SK129.

### Azimuth and elevation control

As demonstrated by his azimuth and elevation system, Joe believes in keeping things simple. Azimuth is obtained by turning a big steering wheel which rotates a mast that extends to the roof. The brake is a simple wing clamp affair that tightens the mast. A 12-inch, 360-degree compass rose on the floor gives azimuth readings. I questioned the elevation and Joe said, "Wait, let me get my elevation shoes." Puzzled at this comment, I waited while Joe disappeared downstairs. When he came back up with his elevation

PHOTO B

**Antenna azimuth control.**

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| CA-2X4FX | 146 MHz<br>446 MHz | 4.5dB<br>7.2dB | 200W | 5'11" | BASE/REPEATER |
| CA-2X4M | 140-155 MHz<br>440-460 MHz | 4.5dB<br>7.0dB | 150W | 5' | MOBILE |
| CA-2X4SR | 146 MHz<br>446 MHz | 3.8dB<br>6.2dB | 150W | 3'4" | MOBILE |
| CX-901 | 146 MHz<br>446 MHz<br>1.2 GHz | 3.0dB<br>6.0dB<br>8.4dB | 150W | 3'6" | BASE/REPEATER |
| CX-801 | 146 MHz<br>446 MHz<br>1.2 GHz | 3.0dB<br>6.8dB<br>9.6dB | 100W | 3'3" | MOBILE |
| CA-630TN | 146 MHz<br>446 MHz<br>1.2 GHz | 2.15dB<br>2.15dB<br>5.5dB | 150W<br><br>50W | 1'5" | MOBILE |

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**PHOTO C**

**EME enthusiast Joe Mutter, PA0UMV.**

**PHOTO D**

**PA0JMV adjusts antenna for elevation control.**

shoes, he demonstrated his system. To adjust the elevation, Joe opens the skylight, climbs onto the tile roof (hence the shoes), and tightens or loosens a tent guy line arrangement until proper elevation is achieved. The system is simple but effective. (I'd sure hate to do this at night during one of our northern Ohio winters.)

## Electronics

Other equipment consists of a homebrew transverter with a GaAsFET 3SK124 to a Schottky mixer to 28 MHz. The power amp is made up of two 4CX250Rs driven by a BLY88. The power after a low pass filter is just over the kilowatt level. There are two HF transceivers; one is a JRC JST135 and the other a Yaesu FT102 plus FV102DM. Audio is processed through a homemade audio filter that is adjustable to 20 Hz.

## Background data

Joe's interest in EME was sparked in 1973 when the United States Naval Research Lab (W3KE) started experiments with their big dish. He called them and was heard 459 plus sideband using a pair of F9FTs. They didn't transmit, so he made no two-way contacts. The day after, Joe's father looked at what he had built and asked him to "remove those dangerous looking things." In 1975, Stanford Research Institute organized an EME test weekend. This station was equipped for two-way contacts, but Joe decided that, with its 150-foot dish, it wasn't really an Amateur station. He had called W6PO the day before and asked him for a try. This was Joe's first 2-meter EME QSO. The next day he worked WA6LET without any problems.

Shortly after this, Joe's radio activity went to low-key operation. But in 1986, after some encouragement from SM2CEW, Joe again became interested in EME. A contact with W5UN using a ten-element CueDee and a kilowatt infected Joe with the EME bug, and brought him to where he is today. hr


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# COMING EVENTS

## Activities — "Places to go . . ."

**SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS:** PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC, ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

**MARCH 31-APRIL 1: MARYLAND:** The Baltimore ARC's 1990 Greater Baltimore Hamboree and Computerfest, Maryland State Fairgrounds Exhibition Complex, Timonium. Gates open 8 AM. Admission $5 for both days. Children under 12 free. For information and space reservations contact GBH&C, PO Box 95, Timonium, MD 21093-0095 or 1-800-HAM-FEST 24 hours a day. In MD dial 301-HAM-FEST on TT phone.

**APRIL 7: MINNESOTA:** The Lake Region ARC's 3rd annual Hamfest, Otter Tail County Fairgrounds Hockey Arena, Hwy 59 So., Fergus Falls. 8 AM to 3 PM. Registration $3/advance; $4/door. 6' tables $4. For information call (218) 826-6274 or write Keith McKay, N0FKF, Rt 1, Box 46, Battle Lake, MN 56515.

**APRIL 7: MICHIGAN:** The Blossomland ARA's Hamfest, Fair Plain Jr. High School, Benton Harbor. 8-12. Admission $3. For information Paul Reissmann, WD8MWT, B.A.R.A., PO Box 175, St. Joseph, MI 49085 or call (616) 429-6230.

**APRIL 7: WASHINGTON:** The 12th annual Inland Empire and Eastern Washington Section Hamfest, Spokane Youth Sports Bingo Hall, E. 2230 Sprague Ave, Spokane.

**APRIL 7: PENNSYLVANIA:** AARG Hamfest and Computer Show, Lebanon Fairgrounds, Lebanon. Admission $4. Tailgating $3/space. Handi accessible. For information contact Homer, WA3YMU, (717) 345-3780, Ron, WB3HNX (717) 345-8667, Paul, WB3HEC, (717) 566-2606.

**APRIL 8: NORTH CAROLINA:** The Raleigh ARS is sponsoring its 18th annual Hamfest and NCS ARRL Convention, Jim Graham Building, NC State Fairgrounds, Hillsboro St, Raleigh. Admission $4 to April 2. For information contact Rollin Ransom, NF4P (919) 269-4406.

**APRIL 14-15: TEXAS:** The Key City ARC' Swapfest and ARRL Convention, Abilene Civic Center, Pine St.8 AM to 5 PM and 9 AM to 4 PM Sunday. Admission $5 advance; $6/door. Tables $2. Send pre-registration to KCARC, PO Box 2722, Abilene, TX 79604 or contact Bill Jones, N5DOX (915) 698-4606.

**APRIL 15: MASSACHUSETTS:** Tailgate electronics, computer and Amateur Radio flea market, sponsored by the MIT Radio Society and the MIT Electronics Research Society, Albany and Main St, Cambridge. Admission $1.50. Free off street parking for 1000 buyers. For space reservations or info call (617) 253-3776. Mail advance reservations before April 5 to Richard Brezina, 3 Ames St, Cambridge, MA 02139.

**APRIL 21: OKLAHOMA:** The Lawton-Fort Sill ARC's 43rd annual Hamfest, County Fairgrounds, Lawton. 8 AM to 5 PM. For information contact Claude R. Matchette, 3411 NW Atlanta Ave, Lawton, OK 73505. (405) 357-5870.

**APRIL 21: ARKANSAS:** ARRL All Arkansas Ham Convention, North Little Rock Community Center. Sponsored by the C.A.R.E.N. Radio Club. Admission $3. For information contact Bob Hancock, KB5IDB (501) 771-2617 or Leon Schlosser, N5GYW (501) 835-4004 evenings.

**APRIL 22: MASSACHUSETTS:** Wellesley ARS Hamfest, Wellesley Senior High School parking lot, 50 State Street, Wellesley. 9 AM to 2 PM. Admission $2. Contact Gerry Driscoll, NV1T (617) 444-2686.

**APRIL 22: CONNECTICUT:** The 7th annual Southington ARA's Flea Market, Southington HS Cafeteria, 720 Pleasant Street, Southington. 9 AM to 1 PM. Admission $3. For information Chet, KA1ILH, 628-9346, 5 to 9 PM.

**APRIL 22: ILLINOIS:** The Moultrie ARK's 27th annual Hamfest, Cadwell Road, Sullivan. Tickets $4. For information, tables, space call or write Ralph Zancha, WC9V, 502 E. State St. Lovington, IL 61937. (217) 873-5287 evenings.

**APRIL 27: DAYTON, OHIO:** The Dayton-Cincinnati Chapter of the Quarter Century Wireless Association announces its annual QCWA Banquet, Neil's Heritage House. COD bar 6:30. Banquet 7:30. Tickets $14, reservations required. For tickets or information contact Bob Dingle, KA4LAU, 657 Dell Ridge Drive, Dayton, OH 45429. (513) 299-7114.

**APRIL 27: DAYTON, OHIO:** The Southwest Ohio DX Association will again host the DX dinner at the Stouffer Center Plaza Hotel. Advance tickets only $23.00 per person. SASE to with check payable in US funds to SWODXA to Scott Lehman, N9AG, PO Box 803, Greenville, OH 45331.

**APRIL 27: DAYTON, OHIO:** The 21st annual B*A*S*H*, Conference Center (Madison Room) of the HARA Arena. Starts 7 PM. No admission charge. Hot food available. Free entertainment. Also a hot-air balloon launch Saturday, April 28 at 6:45 AM (weather permitting) conducting aeronautical QSO's. Frequencies announced at B*A*S*H*. Special event cards will be issued. Sponsored by the Miami Valley FM Association, PO Box 263, Dayton, OH 45401.

**APRIL 29: NEW HAMPSHIRE:** "Dayton East", Contoocook (near Concord). Starts 7 AM. Buyers $1. Sellers $5. For information K1OPQ (603) 746-5090 or PKT BBS ⁹ WA1WOK-2.

**MAY 4-6: ARIZONA:** The Cochise ARA's annual Hamfest, club training facility, Moson Road, Sierra Vista. Free tailgating. Handi facilities. For information contact N7INK (602) 378-3155 after 6 PM or write CARA, PO Box 1855, Sierra Vista, AZ 85636.

**MAY 5: WISCONSIN:** The Ozaukee Radio Club's 12th annual Cedarburg Swapfest, Circle B Recreation Center. 8AM to 1 PM. Admission $2/advance; $3/door. For tickets, table reservations, maps, information send business SASE to ORC Swap, N5415 Crystal Springs Ct, Fredonia, WI 53021. (414) 692-2329.

**MAY 20: ILLINOIS:** The annual Hamfest sponsored by the Kankakee Area Radio Society, Will County Fairgrounds, Peotone. 8 AM to 2 PM. Indoor flea market and exhibitor tables (limited). Large outdoor flea market. ARRL booth. free parking. Food and drink available. Admission $2.50 advance, $3.00 at the door. Setup kMay 20, 6-8AM. Talkin on 146.34/94. More information from KARS, c/o Frank DalCanton, KA9PWW, RR1, Box 361, Chebanse, IL 60922. Tel (815) 932-6703 after 4 PM CST or (815) 937-2452 before 4 PM CST.

# OPERATING EVENTS

## "Things to do . . ."

**The Tipton ARS** will operate special event station KC7YL from Fort Pillow State Park in Lauderdale County, Tennessee. 1600Z April 7 to 1900Z april 8. Voice: 28.450, 21.350, 14.325. CW: 28.200 and 21.150. For an attractive certificate send QSL and large SASE to WB4DPL, PO Box 402, Covington, TN 38019.

**The Cornish Radio Amateur Club of England** is sponsoring International Marconi Day to celebrate Marconi's birthday. 0000Z -2400Z April 21 in the General bands. Work 10 stations and send log to CRAC, PO Box 100, Truro, TX1 1RX, Cornwall, UK.

**The Warminster ARC's** 2nd annual DXpedition to Delaware, April 8, operating WA3DFU/3. Freq: 7.275, 14.275, 28.375. CW on request. QSL with SASE to Warminster ARC, Box 113, Warminster, PA 18974.

**The Great River ARC,** Dubuque, IA, will operate special event station WK90 (N9GBY) April 28 at the 36th annual U.S. Boy Scout Pilgrimage, 10-4 local time, lower General band and Novice 10m band, SSB. No SAE, QSL required for certificate.

**Clairemont Repeater Association** will operate "Sam's Day" in honor of the 199th birthday of Samuel F.B. Morse. Control op W6FZZ, Samuel Morse 3d. 1800 UTC to 2400 UTC Saturday April 28. 28.350 and 21.300 phone and 14.050 CW. QSL to CLARA, PO Box 7675, Huntington Beach, CA 92615.

**THROUGHOUT 1990** the Major Armstrong Memorial Amateur Radio Club (MAMARC) will sponsor events commemorating Major Edwin Howard Armstrong's achievements in the field of radio broadcasting. The club is seeking other Amateur operators around the world who are willing to research Major Armstrong's accomplishments and become official MAMARC special events stations. Major Armstrong was a pioneer responsible for the creation of Wideband FM and the inventor of the superhetrody

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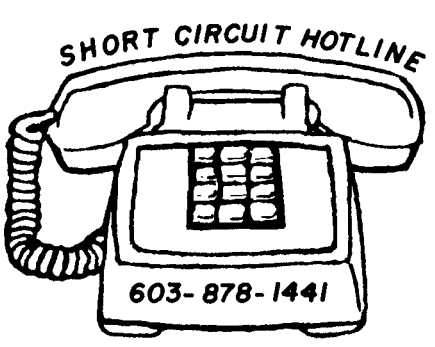


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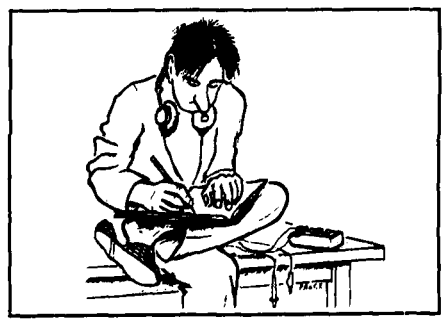


*By Tom McMullen, W1SL*

## ELEMENTARY ELECTRONICS: INDUCTORS AND AC

In past issues I've looked at resistors, capacitors, and magnetic fields around wires and simple coils. I explored them using direct current (DC) as the energy source. Life in the DC world is somewhat simple. Current either flows or it doesn't, and when it flows, it goes from one power source terminal to the other in a straightforward manner. Along the way, it can be restricted by resistors, smoothed by capacitors, divided, combined, and so on. It's all just "plain vanilla" electronic theory. When you enter the world of alternating current (AC), things get a bit more complex. Most of the simple rules apply, but you have to keep one basic rule in mind — alternating current flows from one power source terminal to the other for a short time, *and then turns around and goes back!* Secondary to this is the fact that the voltage isn't constant, but varies from zero volts to some nonzero value (either positive or negative) and then returns to zero twice during each cycle.

### Waveforms

The figures you see on an oscilloscope are generally voltage waveforms, so it's a bit strange to call them AC waveforms. However, if you remember that the voltage you're looking at was probably derived from alternating current flowing through a resistance, it makes sense that the oscilloscope shows a representation of an AC waveform.

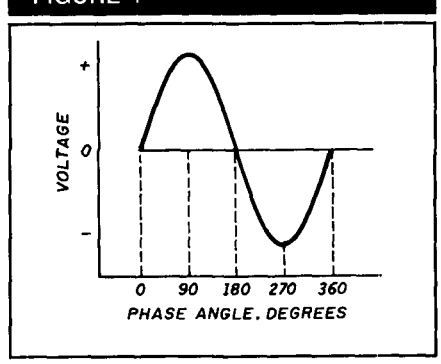
A basic AC waveform, like the one provided by most household electric service, is shown in **Figure 1**. Household AC is generated by mechanical means — a generator that forces a massive winding of wire through a magnetic field. Waveforms generated by electronic circuits (oscillators) can be made with the same shape and cycle period, or with an almost unlimited variety of shapes and cycles.

### Current and voltage relationships

It doesn't matter if the waveform you're examining is made by a mechanical generator or is created by electronic oscillators; the relationship between the voltage and current is what's important.

At the start of each cycle, there's no voltage and no current flow; this is the zero (0) point. As the voltage builds up, current increases. In a circuit with only pure resistance, the current and voltage stay in unison throughout the cycle; when the voltage reaches its peak, so does the current. The complete cycle is said to take place in 360 degrees, and is related directly to the mechanical generator concept where a coil (armature) rotates about its axis while in a magnetic field. Portions of a cycle can be described conveniently by referring to the "degrees" (of rotation) along the waveform. This degree concept also comes in handy when determining how the current is doing compared with the voltage (more on this later).

FIGURE 1




**The waveform of common household AC power is a sine wave with equal positive and negative peaks during a cycle. The degree references are explained in the text.**

The waveform demonstrated in **Figure 1** is called a sine wave. You can generate one on graph paper by plotting the sine of the angle, as shown in **Figure 2**. Not all AC waveforms are as neat as this, however. They can be distorted by amplifiers, capacitors, and inductors, or they may not have been sine waves to start with. Square, triangle, and sawtooth waves, in addition to random audio (speech) waveforms, are also types of AC. But the easiest way to understand how AC behaves in a circuit is to stick with sine waves.

### Phase shift

In simple resistive circuits, the current flow stays exactly in time with the voltage buildup or decay. When the voltage is at maximum, so is the current flow. As voltage changes, so does the current — in exact proportion. Under these conditions, the voltage and current are said to be "in phase" with each other. However, AC really spends most of its life in a world that's full of reactance (I'll talk about reactance shortly). Both inductors (coils) and capacitors have reactance, but I'll just look at inductors this time and save capacitors for a future issue.

In a past discussion of DC flowing in an inductor, I noted that the current flow created a magnetic field which not only used and stored energy, but created a counter force (called counter electromagnetic force or counter emf) that opposed the current flow during the buildup. When the current flow stopped, the magnetic field collapsed and the energy was released — returned to the circuit in the form of electron flow in the same direction as the original current flow!

The same thing happens when AC flows through a coil, but because of the changing nature of AC, the effects are more noticeable. As the voltage builds up through the 0 to 90-degree portion of the waveform, the current gets behind. Remember, it's current (electron movement) that does the work in a circuit. So while the voltage gets from one end of a coil to the other quickly, the current is busy building up a magnetic field which further slows the current buildup because it creates a

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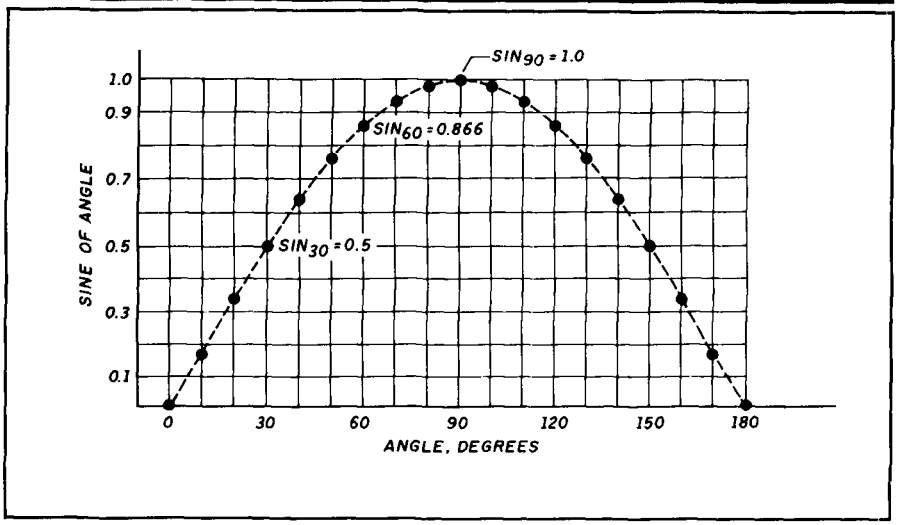
FIGURE 2



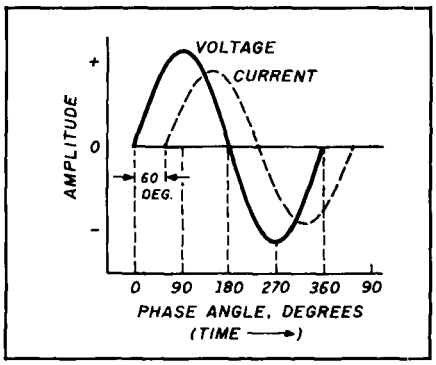

**A sine wave can be drawn on quadrille paper by plotting the sine of the angle.**

counter emf that's pushing back.* As a result, current lags behind voltage, as shown in **Figure 3**. The voltage and current are now "out of phase" with each other. Remember when looking at **Figure 3** that time progresses to the right. Although the current is delayed in time with reference to the voltage, it's far easier to express the delay in degrees. This is because the time for one cycle varies, depending on the frequency. For household power, a cycle occurs 60 times per second, or one cycle equals 0.016666 second. At a frequency in the Amateur 80-meter band, a cycle might occur 3.6 million times per second, or approximately 0.0000002 second for one cycle. If you use degrees, you don't have to worry about how small the fraction of a second is.

The number of degrees the current lags the voltage depends on the inductance value and the frequency of the AC. If the inductance is very small, say ten or 12 turns of wire on a plastic form, and the AC frequency is very low, like 60 cycles per second (60 Hz), the current lag is undetectable. However, with those same ten turns of wire and a frequency of 500 million cycles per second (500 MHz), the inductance has a very great effect on the current — so great that no alternating current can flow through the coil. (This is the principle behind an RF choke.)

Aha, you caught me trying to sneak a term past you! Inductance. Let's

FIGURE 3




**An inductor (coil) causes the current flow to lag behind the voltage, 60 degrees in this example. The voltage and current amplitudes are not drawn to scale.**

define it before continuing. Inductance is the tendency of a wire or coil to oppose any change of current flow through it. It does this by using the magnetic field to first oppose a current increase, and then release the energy stored to oppose the current decrease. The unit of inductance is the henry, abreviated H; smaller units are the millihenry (mH) and microhenry ($\mu$H). The units do not take on the "ies" suffix for plural; it's millihenrys, not millihenries.

Inductance is inductance, and it doesn't matter whether you're trying to make AC or DC flow through it — it opposes both. So you need a way to relate its opposition to the frequency involved, because you've seen that the inductance of a ten-turn coil doesn't bother 60 Hz, but completely shuts out 500 MHz. The property that relates inductance and frequency is called inductive reactance, $X_L$. Because inductive reactance tends to limit current flow in a circuit, it is expressed in ohms.

$X_L$ can be calculated at the frequency of interest. For example, a coil with an inductance of 1 mH (0.001 H) at a frequency of 4 MHz (4,000,000 Hz) will have an inductive reactance ($X_L$) of:

$$X_L = 2\pi fL$$

where f = frequency in Hz
L = inductance in henrys (H)

thus:

$$X_L = 6.28 \times 4{,}000{,}000 \times 0.001$$
$$= 25{,}120 \text{ ohms}$$

Many circuits have resistance as well as inductive reactance, so it's necessary to find some term that combines the two for ease of calculation. This term is impedance. The symbol for impedance is Z.

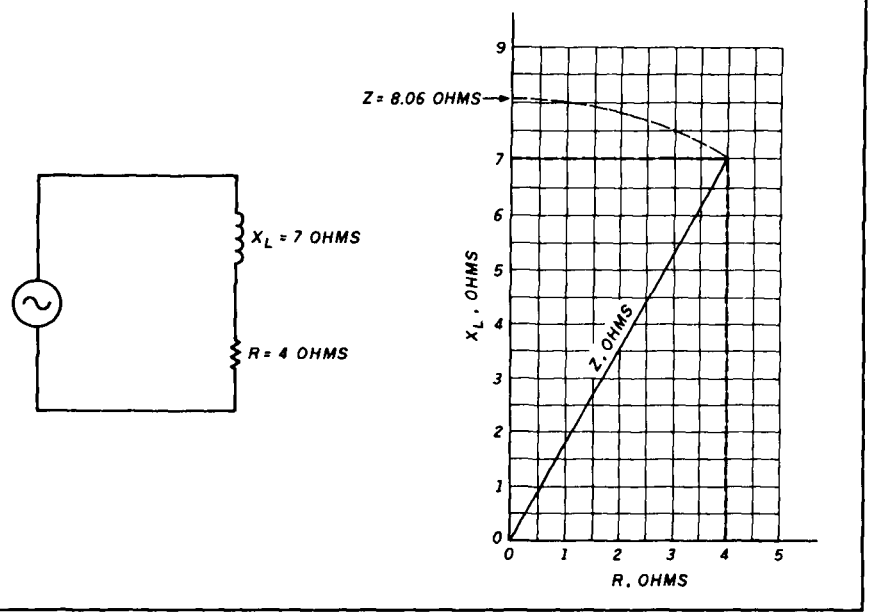
Here's where some of you will get cold chills, blurred vision, and sweaty palms — but hang in there, what follows isn't life threatening (unless you're in the middle of final exams). Impedance is the vector sum of the inductive reactance and the resistance in a circuit.

## Oh, no, not vectors!

I know, I hated vectors too. But like much of life, once you understand them, they're not half bad. To simplify the vector concept a bit, look at a vector as the result of two forces working on the same object. Imagine that you're trying to move a ping-pong ball from your end of the table to the other by blowing on it. At the same time, another person is trying to move the ball from one side to the other in the same way. As a result, the ball takes a path at an angle to both the end and the side of the table. The exact path depends on how hard each of you blows. The amazing thing about vectors is that they add the forces involved. If the other person is at the opposite end of the table instead of the side, the ball might follow a straight path but wouldn't get as far. The opposing negative force counteracts your positive force, and the result is the sum of the two. Thus, if two people were at the other end of the table, the vector could be even more negative, and the ball would go off the table at your end. This

* Don't misunderstand what's being delayed here. The counter emf opposes the number of electrons flowing, but doesn't change the speed of those electrons.

FIGURE 4




**A vector diagram can illustrate the impedance of a circuit with both reactance and resistance.**

ability to sum positive and negative forces is important; you'll see more of it in a future column.

Back to the electronic world, before I run out of space. **Figure 4** shows an example of a vector used to determine the impedance (Z) of a circuit that has both inductive reactance and resistance. The inductive reactance ($X_L$) is plotted vertically, and resistance (R) is plotted horizontally. You can do this on ordinary quadrille paper. The hypotenuse represents the impedance, and for this example is approximately 8 ohms.

Impedance can also be calculated by rearranging the formula for finding the hypotenuse in the Pythagorean theorem: "The square of the hypotenuse value of a right triangle is equal to the sum of the squares of the values of the other two sides." Or,

$$Z^2 = R^2 + X_L{}^2$$

Rearranged:

$$Z = \sqrt{R^2 + X_L{}^2}$$

Using the values from **Figure 4**:

$$Z = \sqrt{4^2 + 7^2} = \sqrt{16 + 49} = \sqrt{65} = 8.06 \text{ ohms}$$

The vectors can also be used to determine how much the current lags the voltage in a circuit. This is only one of the intriguing things yet to be explored. Stay tuned!

## Conclusion

You've seen how inductance affects alternating current, causing current flow to lag behind the voltage. This is *important to engineers* when designing power, audio, and RF circuitry. It's also important to you for understanding how a properly designed circuit can provide maximum power transfer between stages of a transmitter, or between your transmitter and an antenna. You'll need another term to help do this — capacitance — which we'll look at in the future. hr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| PL-259/ST | UHF Male Silver Teflon, USA | 1.50 |
| UG-21D/U | N Male RG-8, 213, 214, Amphenol | 3.25 |
| UG-21B/U | N Male RG-8, 213, 214, Kings | 4.00 |
| 9913/PIN | N Male Pin for 9913, 9086, 8214 fits UG-21D/U & UG-21B/U N's | 1.50 |
| UG-21D/9913 | N Male for RG-8 with 9913 Pin | 3.95 |
| UG-21B/9913 | N Male for RG-8 with 9913 Pin | 5.75 |
| UG-146A/U | N Male to SO-239, Teflon USA | 6.00 |
| UG-83A/U | N Female to PL-259, Teflon USA | 6.00 |

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# DX Forecaster

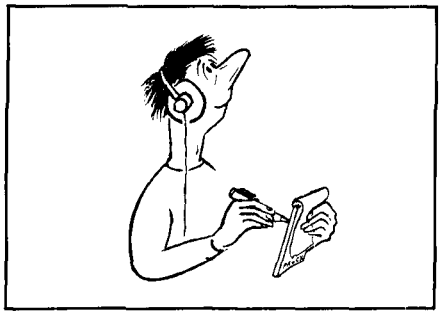
*By Garth Stonehocker, KØRYW*

## DX PROPAGATION SOFTWARE

More and more ham shacks sport computers these days. The popular PCs with up to 640K of on-board computing storage give power that a few years ago was available only in institutions, businesses, and government laboratories. Now Amateurs have new help in logging, operating, and performing computer-aided design of antennas and equipment. Much of the software to accomplish these tasks has been published or advertised for sale in Amateur Radio magazines over the last few years. Today, with the different software packages available, PCs can even be used as aids in propagation prediction.

The early use of lowest and highest usable frequencies (LUF/MUF) predictions was once limited to those who received Amateur Radio magazines or worked where these propagation programs were part of their business activities. (The Institute for Telecommunciations Science, with ITS-78 and IONCAP, is one such place.) In December 1982 Navy researchers developed a propagation program called MINIMUF-3. Written in BASIC, this program was made available to Radio Amateurs in the December 1982 issue of *QST*.[1] It calculated the MUF for each hour of the day for any month and the sunspot number for station-to-station paths, by latitude and longitude. The appearance of MINIMUF-3 led to a proliferation of modifications of the MINIMUF program. In most cases the output was changed to do a specific task or produce a more interesting display. *Ham Radio* published such a display modification in February 1987.[2] It showed the MUF as pseudo-concentric contours around a transmitter.

Advertised propagation prediction programs usually cost between $25 and $50 and support Commodore, Macintosh, and IBM compatibles. Several of these commercial offerings display the MUF/LUFs on a map of the world. Most have grayline propagation, sunrise/sunset times, and antenna bearings. Typical of these are The DX



Helper by W7HR, The DX Edge,* and BANDAID/MUFMAP by Heath. Some programs, like MINIPROP by W6EL, include signal level information. The ultimate software for signal and noise levels is W1FN's IONOSOUND. The program determines the signal to noise and reliability of each propagating mode (E, F, or mixed) over the path, at frequencies you select as being most useful for DX. The display simulates an oblique chirp sounder showing the modes. Combining a MUF program with IONOSOUND gives you capabilities similar to those of the ITS-78 and IONCAP programs — with the exception of world noise and antenna patterns. So you can bring the power of the mainframes of a few years ago to your ham shack PC and use it to help find DX openings.

How do you know how "good" these programs are? Play the beacon game. Pick a beacon path close to your DXing path of interest, if possible. If just one beacon frequency is transmitted, the whole diurnal MUF curve for that path will be hard to define. In any case, monitor the frequency until the signal disappears in the afternoon, defining a time point on the curve. If the signal disappears at a later hour the next day, the MUF was probably higher for that time period. Monitor the signal for several days to build up the statistical average. Use the same technique in the morning as the signal arrives. The morning points shouldn't show as much scatter in time as the afternoon's MUF downward curve.

### Last minute forecast

The second and third weeks are the days of high MUFs (high solar flux). They may decrease signal strengths but lengthen the nearly daily openings on the higher HF bands, 10 to 20 meters. This month's evening one long hop transequatorial openings to the south could come on the 11th and 19th. Other ionospheric disturbed conditions may exist near the 1st and 28th. The low band east-west paths may have weak signals and QSB caused by the two latter disturbances. The low bands will be best at times of low disturbance.

The perigee of the moon's orbit (for moonbounce DX) is on the 26th, with the moon at full phase on the 10th. There will be a short meteor shower, the Lyrid, on April 20th to 22nd with a rate of five per hour — hardly much help for meteor-scatter DX. But a bigger shower, the Aquarid, starts before the end of April, peaks on the 5th of May, and ends in mid-May. Its rate is 10 to 30 per hour.

### Band-by-band summary

Ten, 12, 15, 17, and 20 meters will be open from morning until late evening most days to most areas of the world. The higher bands open closer to local noon and mainly to southern countries. One long hop openings will probably occur in late evening during high solar flux periods with a mildly disturbed geomagnetic field. The lower of these bands work mainly to the north and south in the evenings.

Thirty, 40, 80, and 160 meters will be open mainly at nighttime. The highest band, 30, is open near dawn and dusk on northern paths. The other band's signals are so absorbed during the day at this high sunspot number (SSN) that DXing is relatively rare. Good nighttime signals are present this time of year with low thunderstorm activity. So, except for an occasional weather frontal passage by your QTH, these conditions should make DXing a joy. Disturbed conditions from the high SSN solar activity may bring weak and fluttery signals from interesting places at night on these bands. hr

* Available in C-64 and MS-DOS versions from the HAM RADIO Bookstore for $34.95 plus $3.75 shipping and handling.

REFERENCE

1. Robert B. Rose, K6GKU, "MINIMUF: A Simplified MUF-Prediction Program for Microcomputers," *QST*, December 1982, page 36.
2. Henry G. Elwell, Jr., N4UH, "360-Degree MINIMUF Propagation Prediction," *Ham Radio*, February 1987, page 25.

# APRIL

## WESTERN USA

| GMT | PST | N | NE | E | SE | S | SW | W | NW |
|---|---|---|---|---|---|---|---|---|---|
| 0000 | 4:00 | 12 | 20 | 12 | 10 | 10 | 10 | 10* | 10 |
| 0100 | 5:00 | 12 | 20 | 15 | 10 | 10 | 10 | 10* | 10 |
| 0200 | 6:00 | 12 | 30 | 15 | 10 | 10 | 10* | 10* | 10 |
| 0300 | 7:00 | 12 | 30 | 15 | 10 | 12 | 10* | 10* | 12 |
| 0400 | 8:00 | 12 | 17 | 17 | 12* | 12 | 10 | 10 | 12 |
| 0500 | 9:00 | 15 | 17 | 15 | 12 | 17 | 10 | 10 | 12 |
| 0600 | 10:00 | 20 | 20 | 12 | 15 | 17 | 10 | 10 | 15 |
| 0700 | 11:00 | 20 | 20 | 12 | 15 | 20 | 10 | 10 | 17 |
| 0800 | 12:00 | 20 | 20 | 15 | 15 | 20 | 10 | 10 | 17 |
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| 1000 | 2:00 | 20 | 30* | 15 | 17 | 20 | 12 | 12 | 20 |
| 1100 | 3:00 | 30 | 17 | 15 | 15 | 20 | 15 | 15 | 20 |
| 1200 | 4:00 | 17 | 15 | 10 | 15 | 20 | 15 | 15 | 20 |
| 1300 | 5:00 | 15 | 12 | 10 | 12 | 20 | 15 | 15 | 20 |
| 1400 | 6:00 | 15 | 12 | 10 | 10 | 20 | 17 | 17 | 20 |
| 1500 | 7:00 | 20 | 10 | 10 | 10 | 20 | 17* | 15 | 20 |
| 1600 | 8:00 | 20 | 10 | 10 | 10 | 15 | 17 | 15 | 20 |
| 1700 | 9:00 | 20 | 12 | 10* | 10 | 12 | 15 | 17 | 20 |
| 1800 | 10:00 | 20 | 12 | 10 | 10 | 10 | 12 | 17 | 17 |
| 1900 | 11:00 | 20 | 12 | 10 | 10 | 10 | 10 | 12 | 15 |
| 2000 | 12:00 | 30 | 12 | 10 | 10 | 10 | 10 | 10 | 12 |
| 2100 | 1:00 | 17 | 15 | 10 | 10* | 10 | 10 | 10 | 12 |
| 2200 | 2:00 | 15 | 15 | 10 | 10* | 10 | 10 | 10 | 12* |
| 2300 | 3:00 | 12 | 17 | 10 | 10 | 10 | 10 | 10 | 10 |
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## MID USA

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| 8:00 | 15 | 30 | 15 | 10 | 12 | 10* | 10 | 12 |
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| 10:00 | 20 | 15 | 15 | 12 | 17 | 10 | 10 | 15 |
| 11:00 | 20 | 20 | 12 | 15 | 17 | 10 | 10 | 17 |
| 12:00 | 20 | 20 | 15 | 15 | 20 | 10 | 10 | 20 |
| 1:00 | 20 | 30* | 15 | 15 | 20 | 12 | 12 | 20 |
| 2:00 | 30 | 30 | 15 | 17 | 20 | 12 | 12 | 20 |
| 3:00 | 30 | 17 | 15 | 15 | 20 | 15 | 12 | 20 |
| 4:00 | 17 | 15 | 12 | 15 | 20 | 15 | 15 | 30 |
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## EASTERN USA

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| 1:00 | 30 | 30* | 15 | 15 | 20 | 12 | 12 | 20 |
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| 6:00 | 15 | 20* | 12 | 10 | 10 | 10 | 10 | 12 |
| | ASIA FAR EAST | EUROPE | S. AFRICA | CARIBBEAN S. AMERICA | ANTARCTICA | NEW ZEALAND | OCEANIA AUSTRALIA | JAPAN |

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**TS-950SD**

TRANSMIT THE ULTIMATE SIGNAL

- Digital Signal Processing
- Dual Frequency Receive
- Digital AF Filter • 100 Memories

**CALL FOR DETAILS AND ORDER TODAY!**

## YAESU

**FT-1000**

THE BEST OF THE BEST

- 200 Watts Output
- All Amateur Bands
- Dual Receive
- DDS-Direct Digital Synthesis

**CALL FOR ALL THE DETAILS!**

## ICOM

NEW!

**IC-765** NEW HF TRANSCEIVER

- Built-in Automatic Antenna Tuner and Power Supply
- 99 Memories • 100 W Output
- 160-10M/General Coverage Receiver
- Band Stacking Registers

## ALINCO

NEW!

**DR-570T**

VHF/UHF TWIN BANDER

- 45W on 2M/35W on 70cm
- Receive on both Bands at Same Time
- Extended Receiver Range
- More Features for the Money Than Anyone Else

**CALL TODAY!**

## KENWOOD

**TS-140S** AFFORDABLE DX-ing!

- HF Transceiver With General Coverage Receiver
- All HF Amateur Bands
- 100 W Output
- Compact, Lots of Features

## YAESU

**FT-736R** VHF-UHF BASE STATION

- SSB, CW, FM on 2 Meters and 70 cm
- Optional 50 MHz, 220 MHz or 1.2 GHz
- 25 Watts Output on 2 Meters, 220 and 70 cm
- 10 Watts Output on 6 Meters and 1.2 GHz • 100 Memories

## ICOM

**IC-725** NEW ULTRA-COMPACT HF TRANSCEIVER

NEW!

- USB/LSB/CW, AM Receive Optional Module for AM Transmit and FM TX/RX
- 160-10M Operation • 100 W Output
- Receive 30 kHz to 33 MHz
- 26 Memories with Band Stacking Registers

## AMERITRON

**AL-80A AMPLIFIER**

- Full Kilowatt Output
- 160-15 Meters
- 3-500 Z Tube for Maximum Life
- Precise and Easy Tuning
- Step-Start Inrush Protection™

**SPECIAL SALE!**

## KENWOOD

**TH-225A**

HIGHER POWER 2 METER H.T

- Now 5 Watts Output
- Odd Off Sets
- Wideband Frequency Coverage
- Same Accessories as TH-215

**CALL TODAY!**

## YAESU

NEW!

**FT-470**

COMPACT DUAL BAND FM HANDHELD (2M/70CM)

21 Memories for Each Band
Dual VFO's for Each Band
Up to 5 Watts Power
Built-in CTCSS
Built-in 10-Memory DTMF Autodialer

## ICOM

**IC-24AT**

COMPACT DUAL BAND, FM

NEW!

- 140-150 MHz; 440-450 MHz
- 5W Output
- Crossband Full Duplex
- 40 Double-Spaced Memories
- 4 DTMF Code Memories

**CHECK OUT ALL THE FEATURES!**

## ASTRON

| | | | |
|---|---|---|---|
| • RS7A | $49 | • RS35M | $159 |
| • RS12A | $70 | • VS35M | $174 |
| • RS20A | $89 | • RS50A | $199 |
| • RS20M | $109 | • RS 50M | $219 |
| • VS20M | $124 | • RM50M | $259 |
| • RS35A | $139 | • VS50M | $232 |

## uniden

**HOT WINTER SPECIAL**

**HR-2600 10 Meter Mobile**

**$229.95 Delivered**

*Great Deals On Scanners Too! CALL*

## ARRA

**PK-232**

MULTI-MODE TNC

- AMTOR, ASCII, Baudot, CW, FAX, NAVTEX, Packet
- PAKMAIL™ Mailbox With Third Party Traffic
- Two Radio Ports

**THE ORIGINAL MULTI-MODE TNC**

## ICOM

**2 Meter Sale**

**Your Favorite Mobiles At Tremendous Savings!**

IC-228A, 25 watts
IC-228H, 45 watts
IC-28A, 25 watts
IC-28H, 45 watts

**DON'T WAIT—CALL TODAY**
**SAVE SAVE SAVE**

## MFJ SALE MFJ

**LARGEST STOCK OF ALL YOUR MFJ FAVORITE ACCESSORIES CALL TODAY FOR BEST PRICE**

**MFJ-1278**

Multi-Mode Data Controller

**CALL FOR EXTRA SAVINGS**

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# Now Look What Hand Held Frequency Counters Can Do!

## New from OPTOELECTRONICS

Introducing two advanced hand held frequency counters incorporating many unique features usually found only on laboratory bench counters.

These counters are designed for virtually every measurement application from DC through Microwave including measuring RF transmission frequencies at the maximum possible distance.

The UTC 3000 does all of this and is also the world's first Hand Held universal counter timer with Period, Time Interval, and Ratio measurement capability.

- 10 Digit LCD Display with Gate, Function, and Input Annunciators.
- Direct count (1 Hz resolution in 1 Sec) to over 150 MHz.
- 16 Segment Bargraph displays Input Signal Level. Ensures reliable counting, proven effective in locating concealed transmitters.
- High Accuracy, 1 ppm 10 MHz Crystal Time Base is standard with optional 0.2 ppm TCXO available.
- More usable Sensitivity than in any other counter for efficient antenna pick up measurements.
- Four push button selectable Gate times.
- Ni-Cad battery pack and AC adapter-charger included.

**In addition, The Model UTC3000 features:**

- In addition to Frequency, additional Functions Include: Period, Ratio, and Time Interval and Average.
- Single Shot Time Interval 100 ns, .1 ns averaged.
- Two input channels with High Impedance and 50 ohm input

Also Available from Optoelectronics, 8 Digit LED Hand Held Frequency Counters:
Model 2210 10 Hz - 2.2 GHz General Purpose Audio to Microwave $219
Model 1300H/A 1 MHz - 1.3 GHz RF Counter $169
Model CCB Relative RF Signal Strength Bar Graph Meter With 10 Segment LED Display $99

| | |
|---|---|
| Model 2600H | $325 |
| Model UTC 3000 | $375 |
| Model TCXO-30 | $ 80 |
| Model TA100S Telescoping Whip Antenna | $ 12 |

**OPTOELECTRONICS INC.**

5821 N.E. 14th Avenue, Fort Lauderdale, Florida 33334

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# YAESU

# Power...and More!

## FT-212RH

### Frequency Synthesized VHF/UHF FM Transceiver

The compact, versatile FT-212RH is a 45 watt, 2 meter mobile that boasts a lot more than just high power. Inside its sturdy compact frame hides an impressive array of performance features plus high reliability...like 18 general purpose memories; one-touch call channel memory; two scanning range memories; CTCSS on any of the 37 standard tone frequencies may be programmed into any memory channel. Choice of standard, or optional, high performance tone encoding microphones. The FT-212RH and its 35 watt UHF counterpart, the FT-712RH are packed with state-of-the-art refinements... power and more!

- **Frequency Range:** 140-174 MHz on receive (144-148 MHz TX—Modifiable for MARS and CAP). Specifications guaranteed on amateur bands only.
- **Power Output:** 45 watts output with selectable 5 watt low power.
- **CTCSS:** Access any of the 37 standard CTCSS tone frequencies, plus 97.4 Hz can be displayed, selected and programmed into any memory for transmission.
- **19 Memories:** Each memory stores either programmable repeater shift or independent TX and RX frequencies.
- **Automatic Repeater Shift (ARS):** Enables selection of repeater transmitter offset automatically when tuned to a standard repeater subband.
- **Programmable Scanning:** Scans band, band segment or memories. Scan auto-resume with carrier drop or after 5-second pause.
- **Tuning Steps:** Operator selectable steps in 5, 10, 12.5, 20 and 25 KHz increments.
- **CAT System Control:** Provides for external control of VFO frequency, mode and memory functions from operator's personal computer.
- **Amber Backlit LCD Display:** Automatically controls the brightness of the display backlighting and pilot lamps.
- **Tone Encoding Microphone:** Choice of standard, or optional high performance DTMF tone encoding microphones.
- **Digital Voice System (DVS-1):** Optional sytem which allows local and remote digital voice recording and playback.

## FT-4700RH

### Dual Band VHF/UHF Trunk Mountable FM Transceiver

• **Frequency Range:** 140-174 MHz on 2m (modifiable for MARS and CAP); 430-450 MHz on 70cm. • **Power Output:** 50 watts on 2m; 40 watts on 70cm. Selectable 5 watts low power on both bands. • **Full Duplex Cross Band Operation:** • **Dual Receive:** • **CTCSS Encode/Decode:** • **Remote Control Kit Included:** • **Amber Backlit LCD Display** and controls with dimmer switch. • **20 Memories:** • **Dual Antenna Ports:** • **Programmable Scanning:** • **MH-15C8** Mic standard: **MH-15D8** Mic optional.

For information on these and Yaesu's full line of products, call our literature desk toll-free at **1 (800) 999-2070.**

**YAESU**

*Performance without compromise.*

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# KENWOOD

## TS-950SD

"DX-clusive" HF Transceiver

**The new TS-950SD is the first Amateur Radio transceiver to utilize Digital Signal Processing (DSP), a high voltage final amplifier, dual fluorescent tube digital display and digital meter with a peak-hold function.**

# The Ultimate Signal.

• **Dual Frequency Receive Function.** The TS-950SD can receive two frequencies simultaneously.

• **New! Digital AF filter.** Synchronized with SSB IF slope tuning, the digital AF filter provides sharp characteristics for optimum filter response.

• **New high voltage final amplifier.** 50 V power transistors in the 150-watt final section, resulting in minimum distortion and higher efficiency. Full-power key-down time exceeds one hour.

• **New! Built-in microprocessor controlled automatic antenna tuner.**

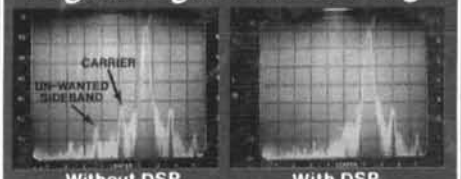
• **Outstanding general coverage receiver performance and sensitivity.** Kenwood's Dyna-Mix™ high sensitivity direct mixing system provides incredible performance from 100 kHz to 30 MHz. The Intermodulation dynamic range is 105 dB.

• **Famous Kenwood interference reduction circuits.** SSB Slope Tuning, CW VBT (Variable Bandwidth Tuning), CW AF tune, IF notch filter, dual-mode noise blanker with level control, 4-step RF attenuator (10, 20, or 30 dB), switchable AGC circuit, and all-mode squelch.

*Complete service manuals are available for all Kenwood transceivers and most accessories.*
*Specifications, features and prices subject to change without notice or obligation.*

DSP

*Digital Signal Processing*




Without DSP | With DSP

• **Digital Signal Processor.** DSP is a state-of-the-art technique that maximizes your transmitted RF energy.

• **High performance IF filters built-in†** Select various filter combinations from the front panel. For CW, 250 and 500 Hz, 2.4 kHz for SSB, and 6 kHz for AM. Filter selections can be stored in memory!

• ***Multi-Drive Band Pass Filter* (BPF) circuitry.** Fifteen band pass filters are available in the front end to enhance performance.

• **Built-in TCXO for the highest stability.†**

• **Built-in electronic keyer circuit.**

• **100 memory channels.** Store independent transmit and receive frequencies, mode, filter data, auto-tuner data and CTCSS frequency.

• **Digital bar meter.**

**Additional Features:** • Built-in interface for computer control • Programmable tone encoder • Built-in heavy duty AC power supply and speaker • Adjustable VFO tuning torque • Multiple scanning functions • MC-43S hand microphone supplied

**Optional Accessories**

- DSP-10 Digital Signal Processor *
- SO-2 TCXO* • VS-2 Voice synthesizer
- YK-88C-1 500 Hz CW filter for 8.83 MHz IF*
- YG-455C-1 500 Hz CW filter for 455 kHz IF*
- YK-88CN-1 270 Hz CW filter for 8.83 MHz IF
- YG-455CN-1 250 Hz CW filter for 455 kHz IF*
- YK-88SN-1 1.8 kHz SSB filter for 8.83 MHz IF
- YG-455S-1 2.4 kHz SSB filter for 455 kHz IF*
- SP-950 External speaker w/AF filter
- SM-230 Station monitor w/pan display
- SW-2100 SWR/power meter
- TL-922A Linear amplifier (not for QSK)

* Built-in for the TS-950SD
† Optional for the TS-950S

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