

FEBRUARY 1990 / \$2.95

HR's New Column—
Microwaves, by KA7GT

NEW **HAM RADIO**



Understanding Over-the-Horizon Radar



THE BEST OF BOTH WORLDS.

The pacesetter IC-R9000 truly reflects ICOM's long-term commitment to excellence. This single-cabinet receiver covers both local area VHF/UHF and worldwide MF/HF bands. It's a natural first choice for elaborate communications centers, professional service facilities and serious home setups alike. Test-tune ICOM's IC-R9000 and experience a totally new dimension in top-of-the-line receiver performance!

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In 26 countries around the world, tens of thousands of amateurs know that Kantronics is the leader in bringing tomorrow's technology to their stations today. They also know they will always be among the first to incorporate just-introduced features and modes with Kantronics software and firmware updates.

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In this age of telco LANS, E-mail and FAX,



PBBS is just one of the firsts Kantronics delivered.

you will know you have mail in your **Personal Packet Mailbox™** when your KAM "STA" LED is blinking. New firmware level 2.85 has also added a handy automatic mailbox user-

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and Tomorrow...

Will the Real Dual-Port Please Stand Up?

Read our lips. **The KAM™ is the only true dual-port when it comes to packet.** Your Personal Packet Mailbox™ is accessible from both HF and VHF! Version 2.85 has dual-port compatibility with RLI/MBL boards and KISS mode for both ports. You can monitor HF and VHF packet operations at the same time. Users can even gateway from HF to VHF (or in reverse) through your KAM.

Kantronics All-Mode™ (KAM) has Packet, WEFAX, ARQ, FEC, RTTY and CW reception. But we have five models to suit your particular taste. **Ask your dealer for the best choice today...and tomorrow.**

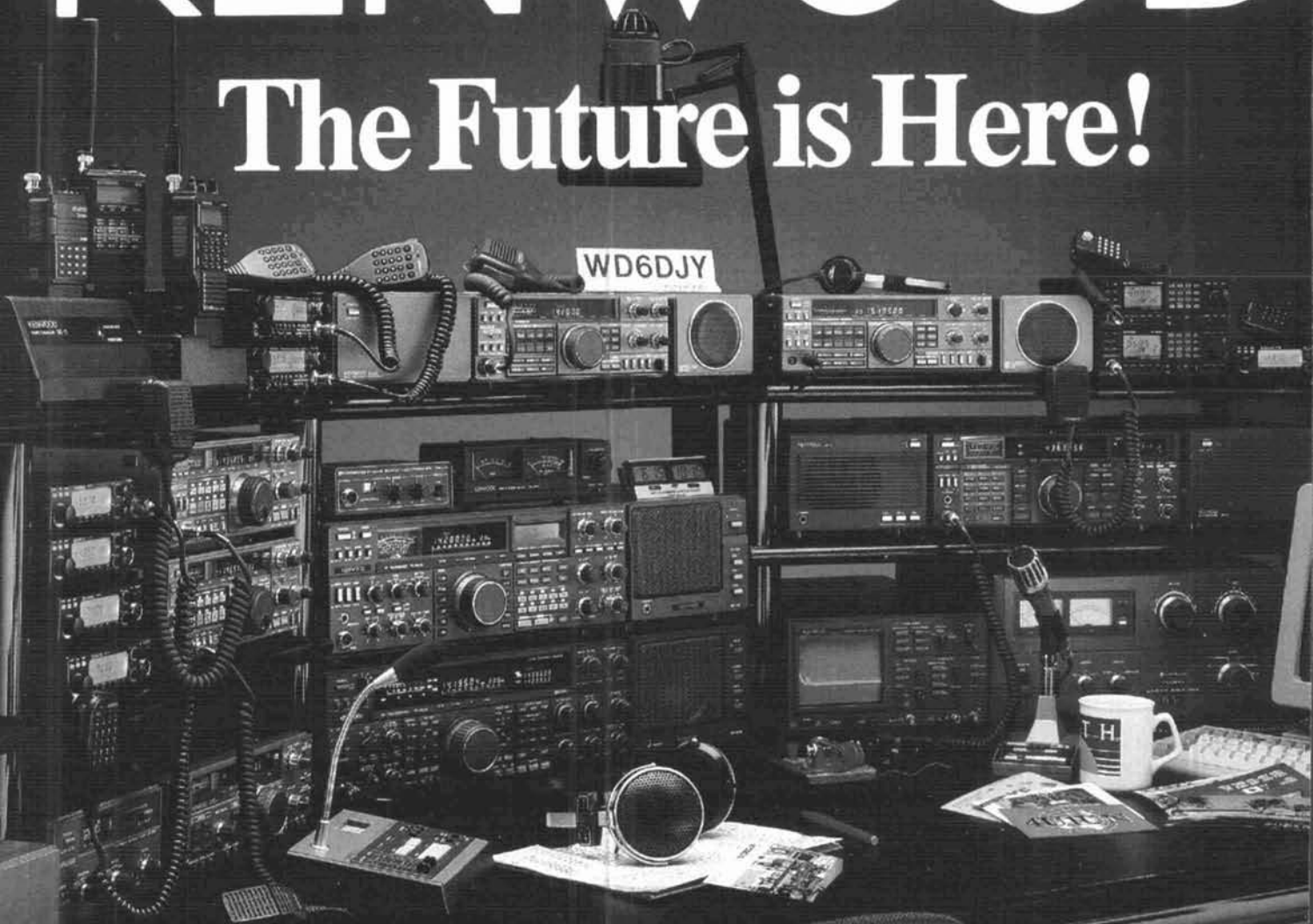
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The Future is Here!



Kenwood — producing the finest Amateur Radio products for over three decades — introduces you to our world of affordable, high-quality, high performance products for today's active Radio Amateur. From HF to VHF, from base to mobile to HT, there's surely a Kenwood radio that will fit your needs and budget.

Being Number One means that we are committed to offer you the finest line of Amateur Radio products in the world. Take a look at the station equipment most winning contest and DXCC Honor Roll operators use. Read the Product Reviews. Compare our rigs against the rest. You'll see that we really are the best.

As the "pacesetter" in Amateur Radio, Kenwood continues to incorporate tomorrow's techniques and innovations into practical products today. Digital Signal Processing (DSP) in the TS-950SD is only one example. SSB Slope Tuning, the original "Dual Bander" concept, built-in antenna tuners for HF rigs, and many other techniques were all developed by Kenwood, and imitated by others. Leading edge technology, and superior field-proven performance — That's the Kenwood Experience!



1) TH-26AT: Compact HT. 2) BC-11: Rapid Charger for TH-26AT. 3) TH-315A: 220 Deluxe HT. 4) TH-75A: Dual Band HT. 5) BC-10: Compact Charger. 6) TM-631A: 144/220 MHz Dual Bander (shown w/supplied MC-44DM). 7) PS-50: Heavy Duty Power Supply. 8) TS-440S: HF Transceiver w/AT-440 (shown w/supplied MC-43S). 9) SP-430: External Speaker. 10) R-5000: High Performance

Receiver. 12) HS-6: Small-size Headphones. 13) SP-430. 14) TM-2530A: Deluxe 2m, 25 W FM Transceiver (w/optional MC-48B mic). 15) TM-3530A: Deluxe 220 MHz, 25 W FM Transceiver. 16) TM-701A: 144/450 MHz Compact Dual Bander Mobile Transceiver. 17) TM-231A/331A/431A/531A: 144, 220, 450, 1200 MHz Compact Mobile Transceivers. 18) TS-811A: 70 cm, 25 W and 19) 2m, 25 W, All Mode Base Station Transceivers. 20) PC-1A: Phone Patch (FCC part accepted). 21) SW-2100: SWR/Power Meter. 22) TS-940S: Deluxe HF Transceiver w/AT-940 installed. 23) SP-940: Matching External Speaker. 24) SP-31: External Speaker. 25) TS-790A: All Mode Tri-band Satellite Transceiver. 26) PS-31: Matching Power Supply. 27) PS-430. 28) AT-250: External Automatic Antenna Tuner. 29) TS-680S: 160-6 m Multi-Bander. 30) TS-950SD: HF Transceiver w/DSP. 31) SP-950: Matching External Speaker. 32) SM-230: Deluxe Station Monitor. 33) TL-922A: HF Linear Amplifier. 34) MC-85: Multi-function Desk Mic. w/3 outputs and tone controls. 35) HS-5: Deluxe Headphones. 36) MC-60A: Base Station Mic.

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...pacesetter in Amateur Radio

HAM RADIO

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Cover photo: Portions of the transmit antenna arrays for the West Coast Over-the-Horizon Radar System. The antenna sections pictured are part of a nearly 4,000-foot long array located near Christmas Valley in southern Oregon. Photo courtesy of GE Aerospace, Syracuse, New York. Special thanks to the Air Force Public Affairs Office at Hanscom Air Force Base, Massachusetts for their help in obtaining the photo.



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Backscatter



Somewhere Over the Horizon (radar, that is)

Though it has long been one of the most insidious of the co-inhabitants of the Amateur bands, over-the-horizon (OTH) radar is also one of the most fascinating technologies in the field of electronics. There are few Amateurs, anywhere, who haven't had at least one run-in with the infamous Russian Woodpecker. But did you know that the United States, Australia, and Canada have working systems and such systems may be in the developmental stages in other countries?

Bryan Bergeron, NU1N, has written a well-documented piece on OTH radar. It appears as our feature article this month. It's interesting to note that this technology has a number of different uses. From peaceful applications like determining wind and wave conditions over the oceans to searching the skies and seas for potential military threats, OTH radars are very powerful tools.

Bergeron points out the significant potential for interference created by OTH systems. However, those in the United States have been carefully designed to minimize or eliminate this problem. And, overall, there have been few if any complaints generated by the U.S. systems in use today.

The computers that control U.S. OTH radars are preprogrammed to operate over a spectrum selected to reduce harmful interference to Amateur, broadcast, and fixed service communications. These radars can also listen in on specific frequency ranges before transmitting, to determine if there is a potential for interference. And if the frequency is in use, the system searches for another usable spot.

Many of the advances in digital signal processing have come as a direct result of OTH radar development. Computers are the single most critical ingredients of a functioning OTH radar system. The complexity of the return signal can't be processed in the same manner as a more conventional line-of-sight radar. In fact, World War II radar operators would marvel at the differences between today's OTH radar and their old PPI (cathode ray tube) displays!*

Additional advances have been made in the scientific study of our ionosphere and radio wave propagation. In order to maximize OTH efficiency, transmissions must be made near the maximum usable frequency (MUF). Taking the input from various sources, the OTH radar's transmitter frequency varies continuously depending on the best propagation path.

Finally, the United States has extensive and stringent environmental impact study regulations for radar that must be met before an operational license can be granted.

Defense has been the prime mover behind developmental work in OTH technology and hams have benefited from much of the work that has been done. I'm sure you'll find Bergeron's article fascinating reading. Perhaps you'll want to look at many of the references he's listed. Most should be available from your local library, through inter-library loans, or at the libraries of local colleges or universities.

de NX1G

* *A Race on the Edge of Time* by David E. Fisher, Ph.D. is a fascinating look at the development of radar. It's interesting to compare the OTH technology of the eighties to the "Chain Home" British radar that operated between 25 and 30 Mcs at the beginning of World War II. It's mind boggling when you contrast what was done then with what we are doing now! (If you'd like to read this book, it's available from the HAM RADIO Bookstore for \$19.95 plus \$3.75 shipping and handling.

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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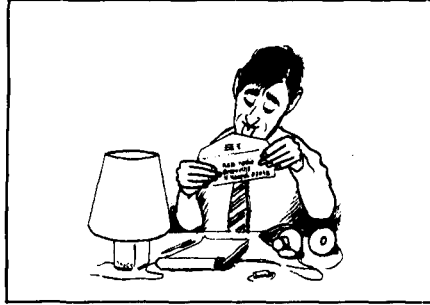
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"Dynamic Duals"



Comments



AM is alive and well

Dear HR

I would like to take umbrage at the remark made by Bill Orr in the December issue, "Too bad the days of amplitude modulation are past —" They are not! If Bill would listen in his own back yard (California) he would find that AM operation is rampant. The SPAM (Society for the promotion of Amplitude Modulation) is headed by WB6TRQ in California. The NCS for the regular Wednesday night activity is W6RNC on 3870. Saturday evenings AM is on 7160. Almost every day that the band is open on 10 meters AM congregates around 29,000.

Besides the usual Johnson Rangers, Vikings, and Valiants, there are quite a few Collins KW-1 transmitters on the air: W6HDU, W7ISX, W7GCO, W8AHC, and WA3PUN.

QST has made similar snide remarks about AM. I hope you will not permit your writers to do the same again.

Byron H. Kretzman, W2JTP/7, Woodinville, Washington
W2JTP/7 is the author of *The New RTTY Handbook* (CQ Publishing Group, out of print), the definitive book on "green key" RTTY operation. Ed.

First choice

Dear HR

This is my first year with *Ham Radio* magazine: what a blessing, what a change! I've seen early printings, which were fine reading. This "new" magazine is by far the best. I subscribe to others. This is the pick of the litter. Please keep up the excellent work. All your staff have shown us a super magazine, and I am sure of more great things in the future.

Congratulations to all.

V. L. Henzel, WB7QBE, Richland, Washington

Must be a duck

Dear HR

We live in a real world. So quack, quack, quack. How quickly we forget.

OMB Director Darman stated to Congress: "If it looks like a duck, walks and talks like a duck, then it's a duck."

Take any time-worn political doublespeak like "revenue enhancement" and it still quacks like a tax duck. A tax by any other name like "user fee" is still a tax. Why try to convince any Congressional tax writing staff member of the extraordinary contributions of Amateur Radio over the past 70 years? When push comes to shove, the bottom line is the paranoid frenzy to get bucks — any way, any form, any pretense. We shall have user fees. So: "Read my lips, no new taxes" and be happy you are only paying user fees for licenses; of course this is *not* a tax!

As was stated a very long time ago in our history, the only time that person and property are safe is when Congress is not in session.

Gene Shapiro, W0DLQ, Prairie Village, Kansas

Pet peeve

Dear HR

The "new" *Ham Radio* is great. Best wishes for continued success.

I'd like to air a pet peeve: Generals and above on the 28.3 to 28.5 MHz band...

I, and other Novices and Technicians, get run off the band every day by these operators. Yesterday was a perfect example: an Extra class running 500

watts, splattering over 5 kHz each way for three hours. You Generals, Advanced, and Extras can run 1500 watts and talk all over the bands; us "little guys" are limited to just 28.3 to 28.5 MHz and 200 watts PEP.

I think it's rude, inconsiderate, and not in the ham spirit for you big strappers to operate on the only 10-meter voice spectrum available to us. Yes, you are qualified to work there — that isn't the point. We'd love to work 29 MHz FM — that's out of band for us. How about taking your linears and privileges somewhere else?

Mike Gleeson, N6SWA, San Diego, California

Advice needed!

Dear HR

I am about to attempt an amateur R/C airplane project, and could certainly use some expert advice from any hams out there who have experience with homebrew radio control or TNC projects. I would be happy to hear from anyone with related interests.

Dave Ewen, N0KET, 4818 New Jersey #203, Wichita, Kansas 67210

Summer conference planned

Dear HR

The Central States VHF Society conference will be held in Wichita, Kansas during July of 1990. We are currently seeking technical programs to be presented at the conference. Programs can range from 15 minutes for a simple hint to 1 hour for a complete project. Persons wishing to present material at the conference should contact Jon Jones, N00Y, 1116 Gatewood Court, Wichita, Kansas 67206 or John Lock, KF0M, 1427 S. St. Clair, Wichita, Kansas 67213.

John Lock, KF0M, Wichita, Kansas

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All-mode
tri-bander!

Warp Drive!



TS-790A Satellite Transceiver

The new Kenwood TS-790A VHF/UHF all-mode tri-band transceiver is designed for the VHF/UHF and satellite "power user." The new TS-790A is an all-mode 144/450/1200 MHz transceiver with many special enhancements such as automatic uplink/downlink tracking. Other features include dual receive, automatic mode selection, automatic repeater offset selection for FM repeater use, VFO or quick step channel tuning, direct keyboard frequency entry, 59 memory channels (10 channels for separate receive and transmit frequency storage), multiple scanning and multiple scan stop modes. The Automatic Lock Tuning (ALT) on 1200 MHz eliminates frequency drift. Power output is 45 watts on 144 MHz, 40 watts on 450 MHz, and 10 watts on 1200 MHz. (The 1200 MHz section is an optional module.)

- **High stability VFO.** The dual digital VFOs feature rock-stable TCXO (temperature compensated crystal oscillator) circuitry, with frequency stability of ± 3 ppm.
- **Operates on 13.8 VDC.** Perfect for mountain-top DXpeditions!
- **The mode switches confirm USB, LSB, CW, or FM selection with Morse Code.**
- **Dual Watch** allows reception of two bands at the same time.
- **Automatic mode and automatic repeater offset selection.**
- **Direct keyboard frequency entry.**
- **59 multi-function memory channels.** Store frequency, mode, tone information, offset, and quick step function. Ten memory channels for "odd split."
- **CTCSS encoder built-in.** Optional TSU-5 enables sub-tone decode.
- **Memory scroll function.** This feature allows you to check memory contents without changing the VFO frequency.

- **Multiple scanning functions.** Memory channel lock-out is also provided.
- **ALT—Automatic Lock Tuning—on 1200 MHz eliminates drift!**
- **500 Hz CW filter built-in.**
- **Packet radio connector.**
- **Interference reduction controls:** 10 dB RF attenuator on 2m, noise blanker, IF shift, selectable AGC, all mode squelch.
- **Other useful controls:** RF power output control, speech processor, dual muting, frequency lock switch, RIT.
- **Voice synthesizer option.**
- **Computer control option.**

Optional Accessories:

- **PS-31** Power supply • **SP-31** External speaker
- **UT-10** 1200 MHz module • **VS-2** Voice synthesizer unit • **TSU-5** Programmable CTCSS decoder
- **IF-232C** Computer interface • **MC-60A/MC-80/MC-85** Desk mics • **HS-5/HS-6** Headphones
- **MC-43S** Hand mic • **PG-2S** Extra DC cable

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



BEST OF MFJ

MFJ, Bencher and Curtis team up to bring you America's most popular keyer in a compact package for smooth easy CW



MFJ-422B

\$129.95

The best of all CW world's -- a deluxe MFJ Keyer using a Curtis 8044ABM chip in a compact package that fits right on the Bencher iambic paddle!

This MFJ Keyer is small in size but big in features. you get iambic keying, adjustable weight and tone and front panel volume and speed controls (8-50 WPM), dot-dash memories, speaker, sidetone and push button selection of automatic or semi-automatic/ tune modes. It's also totally RF proof and has ultra-reliable solid state outputs that key both tube and solid state rigs. Use 9 volt battery or 110 VAC with MFJ-1305, \$12.95.

The keyer mounts on a Bencher paddle to form a small (4-1/8 x 2-5/8 x 5 1/2 inches) attractive combination that is a pleasure to look at and use.

The Bencher paddle has adjustable gold plated silver contacts, lucite paddles, chrome plated brass and a heavy steel base with non-skid feet.

You can buy just the keyer assembly, MFJ-422BX, for only \$79.95 to mount on your Bencher paddle.

Deluxe 300 W Tuner



MFJ-949D

\$149.95 MFJ-949D is the world's most popular 300 watt PEP tuner. It covers 1.8-30 MHz, gives you a new peak and average reading Cross-Needle SWR/Wattmeter, built-in dummy load, 6 position antenna switch and 4:1 balun -- in a compact 10 x 3 x 7 inch cabinet. Meter lamp uses 12 VDC or 110 VAC with MFJ-1312, \$12.95.

Antenna Bridge

MFJ-204B

\$79.95

Now you can quickly optimize your antenna for peak performance with this portable, totally self-contained antenna bridge.

No other equipment needed -- take it to your antenna site. Determine if your antenna is too long or too short, measure its resonate frequency and antenna resistance to 500 ohms. It's the easiest, most convenient way to determine antenna performance. Built in resistance bridge, null meter, tunable oscillator-driver (1.8-30 MHz). Use 9 V battery or 110 VAC with AC adapter, \$12.95.

Super Active Antenna

"World Radio TV Handbook" says MFJ-1024 is a "first rate easy-to-operate active antenna ... quiet ... excellent dynamic range ... good gain ... very low noise ... broad frequency coverage ... excellent choice."

Mount it outdoors away from electrical noise for maximum signal, minimum noise. Covers 50 KHz to 30 MHz.

Receives strong, clear signals from all over the world. 20 dB attenuator, gain control, ON LED. Switch two receivers and aux. or active antenna 6x3x5 in. Remote unit has 54 inch whip, 50 ft. coax and connector. 3x2x4 in. Use 12 VDC or 110 VAC with MFJ-1312, \$12.95.

VHF SWR/Wattmeter

MFJ-812B

\$29.95

Covers 2 Meters and 220 MHz. 30 or 300 Watt scales. Also reads relative field strength 1-170 MHz and SWR above 14 MHz. 4 1/2 x 2 1/4 x 3 in.

MFJ Coax Antenna Switches



\$34.95

MFJ-1701



\$21.95

MFJ-1702B



\$59.95

MFJ-1704

Select any of several antennas from your operating desk with these MFJ Coax Switches. They feature mounting holes and automatic grounding of unused terminals. They come with MFJ's one year unconditional guarantee.

MFJ-1701, \$34.95. Six position antenna switch. SO-239 connectors. 50-75 ohm loads. 2 KW PEP, 1 KW CW. Black alum. cabinet. 10x3x1 1/2 inches.

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UNDERSTANDING OVER-THE-HORIZON RADAR

A glimpse at how the other half lives

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

Have you ever wondered what the "Russian Woodpecker" is really up to? Did you know that the United States has several woodpeckers of its own, and that several more are scheduled to begin operation in the near future on frequencies that could include the Amateur bands? I'd like to examine the source of one of the most bothersome signals on the HF bands: over-the-horizon (OTH) radar.

Some radar fundamentals

Before I get into the specifics of OTH radar, a brief review of some radar basics is in order. Radar, a term derived from the phrase **RA**dio **D**etection **ANd** **R**anging, is based on the principle that RF pulses generated by a carefully timed transmitter are reflected by the objects they encounter. The time it takes for the reflected beam to return to the transmission site can be used to determine the distance or range of the object. Radio waves travel at the speed of light, about 186,000 miles per second. For a target at a distance of 186 miles, the round trip (2×186 , or 372 miles) would take about 2,000 μ s:

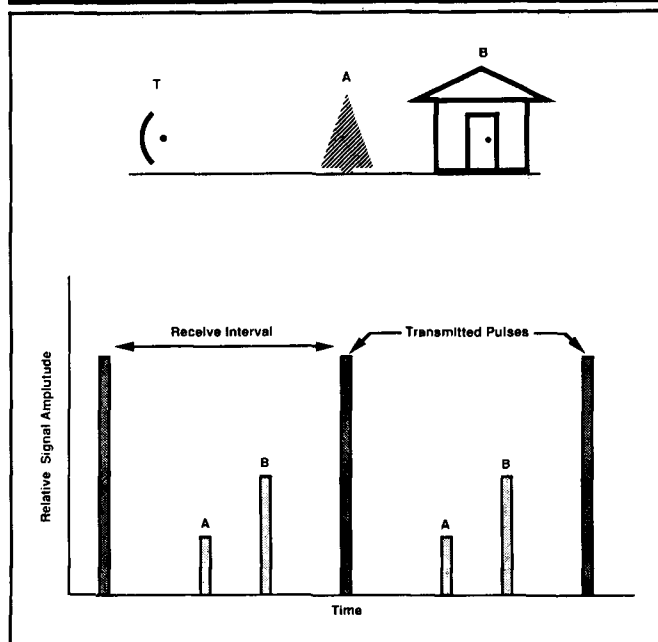
$$\frac{186 \text{ miles} \times 2}{186,000 \text{ miles/sec}} = 2/1,000 \text{ sec} = 2,000 \mu\text{sec} \quad (1)$$

Using this relationship, you can compute relative distances accurately for any given elapsed time.

Because the frequency of the reflected signal is the same as that of the transmitted signal (assuming that there is no relative movement between the transmitter, receiver, and the reflecting object), the transmitter must be turned off for the receiver to detect the much weaker reflected signal. It follows, then, that most radar is pulsed. The transmitter sends out short intense pulses of energy with relatively long intervals between pulses, during which time the receiver is active (see Figure 1). When sufficient time has elapsed to permit

reception of echoes from the most distant area under inspection, the transmitter sends out another pulse. For example, a radar with an interval of 2,000 μ s between transmitted pulses would have a maximum range of 186 miles (Equation 1). Radar pulses vary from a few microseconds to several milliseconds in duration, with rates ranging from 1 to a few hundred hertz.

FIGURE 1



A pulsed radar system with targets (top) and the associated relative amplitudes and time intervals (bottom). Each short transmitted pulse is followed by a relatively long interval during which the receiver is active. The smaller amplitude echos (A) are from the smaller and less reflective tree, while the larger amplitude echos (B) are a reflection from the more distant but larger and more reflective building. The round-trip time from the transmitter (T) to each object can be used to calculate the distance of each object from the transmitter and from each other.

The strength and direction of a radar echo are functions of the size, distance, altitude, and composition of the object causing the reflection. In general, larger, closer, more reflective objects produce stronger echoes or signatures. The effect of an object's composition on its reflectivity depends on the frequency used. For instance, the absorptive surface coating used on United States and Russian stealth aircraft, while effective at microwave frequencies, has little effect in the 3 to 30-MHz range.¹

Anyone who has listened to transmissions from the OSCAR satellites can appreciate the effect that relative motion has on a radio transmission. The Doppler effect describes this apparent increase in frequency of a signal transmitted from a rapidly approaching object, as well as the apparent decrease in frequency of a signal transmitted from a rapidly retreating object. The apparent shift in frequency is a function of *relative velocity*. It doesn't matter if the transmitter is stationary and the object is moving, or vice versa; the effect is the same. The greater the relative velocity, the greater the shift in frequency. From the following equation

$$\text{Observed Frequency} = \frac{\text{actual frequency}}{1 - (\text{relative velocity/speed of light})} \quad (2)$$

it's obvious that, for appreciable changes in the observed frequency, relative velocity must approach the speed of light.² The Doppler shift associated with speeds approaching 20 or 30 miles per hour is only on the order of 0.1 Hz. To the dismay of many motorists, Doppler radar technology has been perfected to the point where relative velocities of this magnitude can be determined to within a few miles per hour.

HF radar

Many Amateurs have the impression that HF radar (radar operating in the 3 to 30 MHz range) is a recent development. However, although the majority of radars currently in use in the United States operate at UHF and beyond, the earliest radar units relied on HF signals.³ HF radars were used as early as the mid-1920s to measure the height of the ionosphere. The first operational military radar, the Chain Home system, installed by the British in 1938 to detect approaching bombers, operated at 25 to 30 MHz.⁴ This line-of-sight radar system operated in the HF spectrum simply because reliable microwave transmitters of sufficient power didn't exist. It wasn't until World War II that effective microwave systems, duplexers, and other technologies required for VHF, UHF, and microwave radar systems were developed.

The first radars were good for determining range, but not direction. This was mainly due to the wide angle radiation beams provided by relatively small HF antenna systems. By decreasing the wavelength, it was possible to build antennas with narrow rotatable beams. These smaller and lighter narrow beam antennas were also more impervious to jamming because they were relatively insensitive to signals other than those arriving head on. During World War II, the German HF radar system was easily jammed because of the wide beams of their receiving antennas. In comparison, the United States Aircraft Warning Radar

system used 100 to 200-MHz signals, and Japanese bombers used radar operating at 200 MHz. These VHF systems, with their narrow beam antennas, were seldom jammed.

Compared with early HF radar systems, today's UHF and microwave radars have several advantages: better support for weather observation; better low angle coverage; wider bandwidth for fewer problems with interference; narrow antenna beamwidth for improved resolution; and smaller, lighter systems that can be easily mounted on aircraft, tanks, and ships.⁴ However, there continues to be a great deal of interest in HF radars — even those that operate as line-of-sight devices. HF systems are generally less expensive to build, have less critical antenna construction requirements, and suffer less interference from rain and insects than millimeter-wave radars.

Conventional radars that use moderate power microwave signals are line-of-sight devices, which means that their maximum operating range is limited by the curvature of the earth. Although airborne radars can be used to monitor the oceans and remote locations, 24-hour operation of an airborne system is expensive and may be impossible in inclement weather. There is currently a great deal of interest in OTH radar that uses high powered signals in the HF band, because these signals can propagate beyond the horizon and reflect from approaching aircraft and missiles.

Over-the-horizon radar

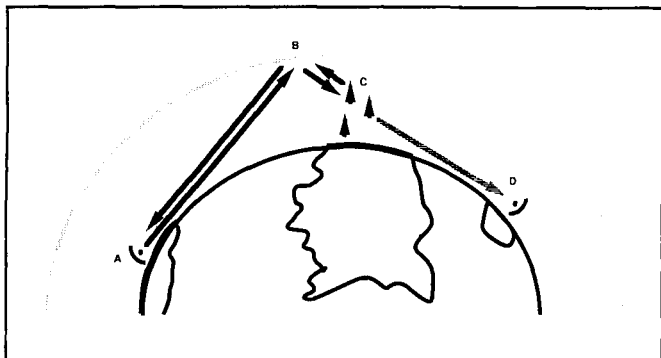
It's somewhat ironic that the very problems which hampered the widespread use of HF radar systems in the late 1930s — namely interference from skywave propagated, long range backscatter echoes from the earth's surface — has led to a renewed interest in HF systems. Today, armed with Doppler radar techniques and digital computer signal processing, the propagation and interference problems of the past can be addressed.

Capabilities

OTH systems provide a number of advantages over conventional radar. The most obvious is the great surveillance range they support, from 500 to over 5,000 miles. Also, unlike conventional radar, OTH can detect aircraft at any altitude between the earth's surface and the ionosphere.⁵ OTH waves are also relatively undeterred by passive stealth techniques that operate at microwave frequencies, like smooth shaping and the coating of aircraft with RF absorbing materials.

In addition to their absolute range, OTH systems are rated for their ability to resolve target range, angle, and velocity. For a single hop backscatter system, the range resolution is typically from 1-1/4 to 25 miles. Angular resolution, largely a function of antenna beamwidth, is typically 30 miles at a range of 1,900 miles. Target velocity resolution, a function of the operating frequency, is typically about 16 miles per hour at 25 MHz.⁶

Although OTH is best suited for aircraft detection, it can also be designed to detect ships and ballistic missiles, or to provide information about wind and wave conditions over the oceans.⁵ An OTH system design is very target specific; it must be designed for the detection of either aircraft, ships, or missiles. The type of waveforms generated by the transmitter and the type of signal processing employed in the receiver system are highly target dependent.

FIGURE 2

OTH radar systems can be designed to operate in either forward or backscatter modes. In backscatter, pulses from the transmitter site (A) are reflected from the ionosphere (B). The earth's surface and targets, like missiles (C), reflect the radar pulses to the ionosphere (B) and to the transmitter site (A). In the forward scatter systems, pulses from the transmitter (A) are also reflected from the ionosphere (B). However, in this system, the receiving antenna is located distal to the area of interest (D).

Operation

Like conventional radar, OTH systems use pulsed RF to illuminate a target area. A major difference between conventional and OTH systems — at least in backscatter systems — is that the ionosphere affects both the transmitted and reflected signals (see Figure 2). OTH receiving systems rely heavily on digital signal processing technology. The composition of reflected radar signals, influenced by the surface of the earth and other objects that reflect HF signals, must be preprocessed by high speed digital computer systems. Only then is the data suitable for display and inspection on a video terminal.

Because OTH illuminates such a large portion of the earth's surface, land and sea echo (clutter) greatly overshadows the returned signal from targets. The dominant source of clutter is the earth's surface. Irregularities on the land and sea, trees, undergrowth, rocky outcrops, buildings, power lines, and towers all contribute to clutter. Natural phenomena, like meteor trails and auroral and equatorial field irregularities, as well as ionospheric instability, add to the clutter.¹ Luckily, Doppler techniques can be used to minimize clutter because the earth's surface is stationary, or at least moves slowly in relation to aircraft and missiles.

OTH radars are Doppler radars. They extract the desired signals from the background noise on the basis of the Doppler shift of signals reflected from moving targets. Signal detection becomes more challenging when the target and background have similar speeds, like ships moving at the same velocity as the sea.⁷ With Doppler processing, the transmitter beam must illuminate the target area for a period (dwell time) sufficient to distinguish moving targets from the background clutter. Depending on the clutter and the nature of the target, the dwell time can range from less than one to over ten seconds.

As in Amateur HF communication, the ionosphere determines the range of operation for OTH systems. Although most systems can use any segment of the HF spectrum, they usually operate near Maximum Usable Frequency (MUF) to minimize propagation losses. The MUF, the highest frequency that can support propagation that relies

on the ionosphere, varies with the state of the ionosphere, time of day, season, and amount of sunspot activity. Ionospheric conditions are usually monitored in real time through vertical sounding of the ionosphere, oblique sounding, and, in a backscatter OTH system, by observing the signals which have been reflected by the earth.

In addition to the problems associated with propagation losses, the ionosphere places additional demands on OTH radar systems. For example, there are usually multiple paths from the radar transmitter to target, as well as from the target to the receiver. In addition, the velocity of propagation is frequency dependent. Together these factors effectively limit the range resolution of an OTH system because short pulses can be distorted to the point of being useless. The refraction of signals by the ionosphere is frequency dependent, allowing a specific area to be illuminated by a narrow range of frequencies. For example, a signal on 10 meters will normally illuminate the earth's surface at a much greater range than a signal on 80 meters. Finally, the ionosphere is in constant flux; propagation characteristics can change unpredictably from moment to moment. An OTH system must be robust enough to handle losses due to polarization mismatch, ground reflection, and the focusing or defocusing of the radar signal due to the ionosphere.

Transmitter and receiver systems

OTH transmitters must provide not only very high power output (on the order of a megawatt or more) but operate over a wide range of frequencies. They must also have the ability to change operating frequency instantly (frequency agility) and produce highly accurate and stable carrier frequencies (for Doppler detection) with high spectral fidelity.⁸ Non-Soviet OTH systems use several transmitters operating simultaneously to meet these requirements.

The transmitted signal can be continuous wave (CW), simple pulse, frequency modulated-continuous wave (FM-CW), chirped pulse, or some other coded waveform.³ Regardless of the type of modulation employed, it's usually desirable to shape the transmitted pulse to minimize the spectral energy contained at frequencies far removed from the carrier (similar to the key click filter in your CW transmitter).³

The main operating variables in the transmitter system are the output frequency, signal bandwidth, pulse repetition rate, and pulse duration. Wide area coverage requires a wide range of frequencies. The radar scanning area (footprint) is moved when the frequency is changed from the low to the high end of the HF spectrum, with higher frequencies used to scan the more distant target areas. Because dwell time must also be considered at each frequency, coverage of a large area can take considerable time. Scan time can be reduced by transmitting and receiving simultaneously at several frequencies, or by transmitting a very wide beam and employing several narrow beam receiving stations.

Good range resolution requires a wide signal bandwidth. A 100-kHz bandwidth, about the maximum that the ionosphere can support at any given instant, corresponds to a range resolution of a little less than a mile.³ The use of wide bandwidth transmitter pulses has the added advantage of reducing echoes from natural targets.³

A low pulse repetition rate avoids range ambiguities. At 50 Hz, for example, the unambiguous range is about 1,900 miles. However, at low pulse repetition rates, Doppler signal

processing becomes more difficult. The pulse repetition rate is, therefore, a compromise between range and velocity resolution. In practice, OTH pulse repetition rates range from several to tens of hertz.⁸ A long pulse duration is required for long range detection. Typical pulse duration varies from hundreds of microseconds to several milliseconds.⁸

In addition to matching the general abilities of transmitter systems — including frequency coverage, frequency agility, and frequency stability — the broadband receivers used in OTH systems must operate in high levels of active interference (from radio transmitters and meteors) and from passive interference (ground and ionospheric reflections). The most sophisticated components of the receiver system, however, don't deal with signal capture but rather with signal interpretation. A single receiver site may require a network of a dozen or more large computer systems, all dedicated to processing the received signals.

Forward scatter and backscatter

OTH radars can operate with either backscattered waves, where a reflected signal is received at or near the transmitter site, or with forward scattering, where the incident and scattered waves propagate in the same direction (see **Figure 2**).⁶ Backscatter OTH systems can detect the presence of a target and measure its location. As in conventional Doppler systems, the transmit/receive time delay determines range. Highly directive antenna systems determine angular coordinates.⁶

Backscatter systems require very high power transmitters (in the megawatt range) for reliable operation, because the losses associated with the ionosphere affect not only the transmitted signals, but the reflected waves as well.

Forward scattering systems can achieve very long operating ranges, on the order of 5,000 miles — about twice the range of backscatter systems. Much lower power transmitters can be used with forward scatter than with backscatter, which is an advantage. Disadvantages include the inability to measure range. Only the presence or absence of a target can be detected. This obstacle can be overcome by using multiple receiving sites, but this is usually prohibitively expensive and complex.

Another constraint associated with forward scatter systems is that the transmitter and receiver sites must be located on either side of the area of interest. For example, to monitor the Soviet Union, the United States installed forward scattering transmitters in Asia, the Pacific, Japan, Taiwan, and the Philippines and receivers in West Germany, Italy, and the United Kingdom.⁶ An additional problem associated with forward scattering is that of maintaining synchronization of transmitter and receiver systems. High stability oscillators in the receiver system that make use of periodic synchronizing signals from the transmitter have been used to help maintain synchronization.

Antenna systems

Antennas for OTH radar must fulfill an almost impossible list of requirements. They must have high directivity and the ability to scan the entire azimuth rapidly; they must also be steerable in elevation. OTH antenna systems must produce high gain, on the order of 20 to 30 dB, to provide both high signal-to-noise and clutter-to-noise ratios. (In this

context, noise is considered to be due to lightning and broadcast transmitters and clutter is considered to be echoes from non-target objects.) Although fulfilling these requirements at a single frequency would be challenging enough, OTH antenna systems must meet the aforementioned specifications for the entire HF band.

Perhaps the most awe-inspiring feature of OTH antenna systems is their sheer size. Systems extending thousands of feet in length and over a hundred feet high are not uncommon. Also, OTH antenna systems often use separate transmit and receive antennas. Although duplexing (using the same antenna for transmitting and receiving) is common in millimeter-wave radars, it's seldom used in OTH systems. Because of the high transmitter power, sometimes on the order of several megawatts, the transmitter and receiving antennas are commonly separated by many miles. By its very design, forward scatter OTH necessitates the use of separate transmit and receive antennas. The high power associated with the transmitters places a number of demands on the construction of the transmitter antenna — including the need for massive feedlines and antenna elements. On the other hand, receiving antennas can be constructed with conventional materials.

Both horizontal and vertical polarization can be used with OTH systems. When vertical polarization is employed, a metallic ground plane is usually placed in front of the antenna to reduce losses — especially if the antenna installation is located inland.⁸ The ground plane for OTH systems most often takes the form of a wire screen extending 150 or more wavelengths in front of the antenna.³ Above ground screens are used to minimize losses caused by snow in places where there's snow on the ground for much of the year.

Installations

There are probably several dozen OTH systems in operation around the globe at any one time. Although there's almost no mention of many of these systems in the literature, several have been described, if only by observers. Perhaps the most infamous OTH system is the Russian Woodpecker (so named because of the characteristic pulse rate of about 10 Hz). This extremely powerful system, located in the Ukraine, was first widely heard during the winter of 1976-77.⁹ Recently, a missile-tracking OTH system based in Siberia has been reported.¹⁰ This system uses a much higher pulse repetition rate than the woodpecker. Its sound has been likened to that of a bumblebee.

Although we tend to attribute all OTH signals to Soviet systems, there are a variety of sources for these characteristic pulses — including several sites in the United States. A system located in Caribou, Maine, for example, uses 21 100-kW CW-FM transmitters, of which seven may be transmitting simultaneously.⁹ The dipole transmitter array, separated from the receive antenna by 100 miles to minimize interaction, is 130 feet high and 2,225 feet wide. The operating range for this backscatter system is from 500 to 1,800 miles.^{6*}

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The Wide Aperture Research Facility (WARF) in Los Banos, California is used for monitoring ocean conditions and for tracking ships in the Pacific and Caribbean. This FM-CW radar system, operating from 3 to 30 MHz, uses separate transmit and receive antennas located 115 miles apart. The receiving antenna, an array of 256 twin whip doublets that extends 1.6 miles, has a beamwidth of only 0.5 degree at 15 MHz.⁷

OTH development isn't limited to the United States and Soviet Union. Australia operates Jindalee, an OTH system near Alice Springs (near the center of the country), to monitor its northern coast.⁵ The transmitter site supports 16 CW transmitters that drive 16 log-periodic antennas. The receiving antenna consists of 492 elements, with an aperture of about 900 feet.⁷ An extensive ground screen is used to maximize transmitter efficiency. The Hall Beach system, located in Melville, Canada produces a 3-MW signal with a log-periodic antenna.⁸

Apparently, each branch of our armed forces conducts its own OTH research. The United States Air Force, which already maintains OTH surveillance of both East and West Coasts, will soon begin transmitting OTH signals from Hanscom Air Force Base in Massachusetts.¹¹ The Navy's MADRE system was developed at the United States Naval Research Laboratory on the Chesapeake Bay in the early 1960s. This 5 to 50-kW system uses a phased array 320 feet wide by 140 feet high. The Table Mesa OTH at Boulder, Colorado, first put into use in the mid-1950s, has reportedly been used to scan the Gulf Coast.⁷

The Pentagon is building a new OTH backscatter network to cover the coastline of the United States. The first unit in this network, built in Maine, ran its first tests last year. Each radar site is capable of tracking targets from 500 to 1,800 miles away and produces 1 MW of FM-CW pulsed radar at angles of 6 to 30 degrees. The Maine unit covers the area from the East Coast of the United States to Cuba, to Greenland, and to Iceland, with an operating frequency range from 5 to 28 MHz. The receiving antenna is 5,000 feet long and uses 28 VAX computers for signal and data processing. An even more elaborate West Coast unit has an 8,000 foot long receiving antenna.¹

Coping with OTH radar noise

Using a highly directive antenna system will minimize the effects of OTH, especially when the radar signal is broadside to the antenna's major lobe. However, because most of us don't have six-element Yagis on every HF band, we need another more general solution. Noise blankers, while not a panacea, can help minimize the effect of OTH QRM.

Most modern transceivers come equipped with noise blankers designed to reduce intermittent noises. Unlike noise limiters — which work by limiting the amplitude of noise spikes to the average signal intensity — noise blankers actually interrupt the IF signal during the exact time of each noise spike. Short duration spikes are handled easily by most noise blankers, allowing the desired signals to stand out clearly from the background noise (the instantaneous breaks in the desired signal aren't usually noticeable).

Unfortunately, the effectiveness of noise blankers diminishes as the noise increases in amplitude and duration. Longer duration high amplitude noise, like the woodpecker noise associated with OTH radar (Figure 3), is handled poorly by simple on/off noise blankers. More capable noise


FIGURE 3



Recording of an OTH signal made on February 10, 1989 at 20:40 UTC (top). At the time of the recording, the signal, which had a pulse rate of approximately 10 Hz, could be heard from 10.495 to 10.509 MHz. Notice the cyclical nature of the signal's peak amplitude, implying that the carrier frequency was systematically changed from pulse to pulse (FM-CW modulation). The illustration at the bottom is a recording of the same signal made at 20:41 UTC. For this recording, the receiver's noise blanker was adjusted to minimize the OTH signal. Note that, although a high quality adjustable noise blanker was used, a considerable amount of the unwanted signal remains.

blankers, with adjustments for duration and blanking level, are best for coping with OTH noise. However, even with the best noise blankers, OTH signals can't be eliminated completely (Figure 3). When using a noise blanker on OTH signals, work with the lowest blanking level possible. Using a higher blanking level than necessary will only add distortion to the desired signal.

Summary

OTH radar, and its characteristic interference, is here to stay. The situation on the HF Amateur bands is likely to become much worse when the new high power systems become active on both east and west coasts of the United States. Because these systems apparently have free reign over the entire HF spectrum, our most immediate recourse as Amateurs is to develop more effective receiver circuitry for rejecting their signals. A long term solution will no doubt require international legislation to limit OTH systems to non-Amateur frequencies. 

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DR-570T

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DJ-160T&DJ-460T

2M H/T is here! And wow!

"Bells & Whistles" is a tame word to use for the new DJ-160T, newest "Magnificent" one from Alinco. Keyboard entry is just one of four ways to enter a frequency in the extended receiver (137-173.995 Mhz) of the DJ-160T. You can store duplex/simplex pairs in any of 20 Memories, or Call Channel, with offsets, and any of 38 encoding subtones. Choose one of 3 scan modes, "Band" "Program" or "Memory" and one of five step ranges in VFO. Priority mode can be used in VFO, Memory or Call. "Dual Watch" allows the DJ-160T to scan 3 seconds alternately on CALL, VFO or one MEMORY. "Pager" is for group or single person alert. Other features include: Auto "Battery Save", Auto "Power Off", and 2-Memory Autodialer. Get 3-watts on standard 700 mah battery, or increased power from built-in DC to DC, or optional 12V battery. The Alinco DJ-160T, now the "Top Gun" with the competition today! DJ-460T for 70cm.



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Smith Design 107 Spectrum Probe

Ever long to own a spectrum analyzer? As a project builder and radio repairer, I sure have! Analyzers are invaluable for a wide range of design and service tasks, and most professional labs have them. Yet the price tags on these electronic marvels have never been in keeping with my "Amateur" status.

Fortunately for those of us on a budget, Smith Design now offers a new device called the 107 Spectrum Probe. For about the price of a 2-meter FM rig, you can team this compact device with any inexpensive oscilloscope and perform many important tests once reserved for expensive and cumbersome laboratory setups.

In terms of technical specifications, the probe's frequency coverage spans from below 1 MHz to more than 100 MHz with a dynamic range of 50 dB or greater. Sensitivity is around 100 μ V, and frequency response is specified at ± 2 dB from 5 to 100 MHz. IF bandwidth is 180 kHz at -3 dB, providing an ultimate resolution of about 0.5 MHz. Maximum CW input is $+15$ dBm or 1 volt at 100 MHz, becoming progressively greater at lower frequencies (the probe has a 10-pF input isolation capacitor). Sweep rate is 6 ms/100 MHz, and horizontal linearity is ± 10 percent. The probe measures 7.5" \times 1" \times 1" and weighs just 2 ounces. External power is supplied by a small wall transformer.



I found the 107 very easy to set up and operate. The probe has no user controls — all adjustments are made at the scope. The 5'6" probe cable has two connectors. There's a line jack for the wall transformer and an RCA plug with BNC adapter for vertical input. Horizontal timing is provided by a sync pulse on the probe's composite waveform.

To obtain a spectrum display, I set vertical gain for 50 mV/div (DC coupled) and triggering for 0.5 ms/div, negative slope. I then repositioned the baseline to the bottom of the screen, pushing the horizontal sync pulse out of view. Finally, I checked the trigger level for stability. "Zero"

MHz was marked on the left of the display by a strong carrier (an internal local oscillator). Each division across the screen represented approximately 10 MHz, with the last corresponding to 100 MHz. Because my scope has x10 sweep expansion, I could zoom in for a closer look at any 10-MHz segment over the probe's 100-MHz range. To check sensitivity, I touched the probe to my finger and brought up several broadcast carriers on the display.

For my first test, I connected a homebrew 20-meter QRP rig to a dummy load and keyed it. I then placed the probe tip near the load, positioning it so the 14-MHz fundamental measured 50 dB, or five divisions, above the baseline (suggested zero reference). At this level, the stronger mixer spurs and a harmonic began to poke up through the noise floor, allowing me to estimate frequency and amplitude. For spotting exact frequencies, I used a signal generator as a tunable marker. This "quickie" sampling procedure yielded nearly identical results to those obtained in a lab evaluation of the same radio.

A lot of the "business" that goes on in RF circuitry is difficult to observe without some form of spectrum analysis. Consequently, I have come to rely on the probe a great deal. Being able to "see" what occurs when two signals mix, when a stage is overdriven, or when a filter is mistuned makes debugging and optimizing much easier. Tough jobs, like checking a mixer-driven transmitter strip for FCC compliance, are simple fare. I've also found the 107 very "test friendly," in that it rarely loads down circuitry and often picks up usable signal levels without physical connection. However, because of this sensitivity, the manual recommends using caution around power amplifiers to avoid burning out the probe's input amplifier. I also observed some desensitizing in high RF fields (the probe case is unshielded).

In addition to functioning as a Hi-Z voltage probe or looped current probe, the 107 interfaces readily with standardized systems. A slip-on adapter converts the probe tip to a BNC-compatible plug. By adding a "T" connector and terminating load, you can plug directly into calibrated generators, attenuators, mixers, and RF bridges. When terminated, the probe is virtually transparent throughout its range.

Although this is a great product, it's important to remember that the 107 isn't a direct replacement for a \$10,000 lab instrument. Dynamic range is 50 dB versus 100, accuracy is ± 3 dB versus 0.3, and coverage is 0 to 100 MHz versus 0 to 1 GHz. But if some tasks are beyond the gambit of the 107, it handles a host of others economically and well. For example, it's a natural for EMI and RFI detection, service work, digital and HF analog design, classroom and laboratory training, production control, and many other real-world applications. And, for the Amateur who builds or repairs ham gear, the Spectrum

Probe is clearly one of the most useful tools I've seen. I recommend it.

Smith Design 107 Spectrum Probes are available directly from Smith Design, 1324 Harris Road, Dresher, Pennsylvania 19025. Credit card orders are accepted and may be placed by calling (215)643-6340. The list price is \$249, including adapters, power supply, and storage case. Significant discounts are available — including a special \$199 direct mail price to Radio Amateurs.

de K1BQT

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The Complete DX'er

Amateur Radio has several very good authors. One of the best is Bob Locher, W9KNI. Bob's ability to spin an interesting yarn about a relatively technical subject is unique. Many of you will remember Bob from his series of articles in *Ham Radio HORIZONS*, the "DXer's Diary." Reading the "DXer's Diary," you shared the joys of working rare DX and the misery of having missed out.

Considerably revised and updated, this new edition of an all-time classic is sure to be a best seller. Bob assumes that his reader wants to work DX. He then sets out to weave a carefully crafted, yet eminently readable, tale of how to do so. From calling CQ to spending hours listening to static crashes on the low bands, Bob gives you the benefit of his years of operating experience and DXCC Honor Roll status.

Special coverage, including operating hints and strategies, is given to working DX in contests — often one of an operator's best opportunities to work new countries. For instance, a good single operator station can work over 100 countries in CQ's World Wide™ contest.

Locher also gives special treatment to working DX through pileups. Read *The Complete DX'er* and you'll have all the tricks of the trade from one of the best at your fingertips. You'll find it's better to listen to a pileup with the rare DX station already in your log than to spend hours calling frantically, and sometimes unsuccessfully!

This book is great reading for the experienced operator and "must" reading for the newcomer. You'll learn how W9KNI has been so successful in his years chasing DX. Who knows, maybe you'll join him on the DXCC Honor Roll someday soon.

The Complete DX'er is available from the **HAM RADIO** Bookstore for \$11.95 plus \$3.75 shipping and handling.

de NX1G

THE HANDY ANTE SWITCH

Frank W. Smith, W4EIN, 2023 Haven Crest Drive, Chattanooga, Tennessee 37421

Several years ago I needed a double-pole, double-throw (DPDT) coax switch in my station. I couldn't find a commercial supplier for such a device, so I coupled two Radio Shack CB-type antenna selector switches mechanically. Basically, they were single-pole, double-throw coax switches. These worked quite well for a number of years — serving primarily to insert a homebrew Class-C amplifier into the antenna feedline when in the send position and to bypass the amplifier in the receive position.

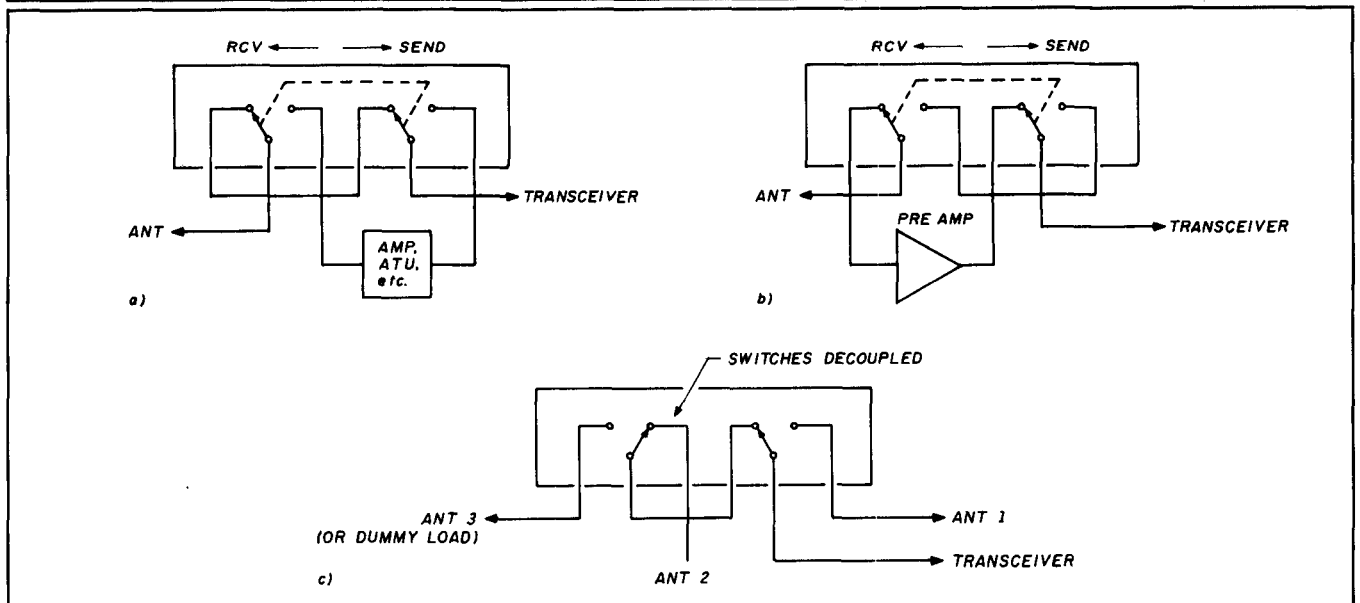
Uses

I discovered that this switching arrangement lent itself to a number of applications. You can insert a preamplifier into

a transceiver input for receiving and bypass it on transmit, bypass the antenna tuning unit on designated bands, and select an antenna or dummy antenna. These applications are shown in **Figure 1**. The switch is also useful for making immediate real-time comparisons under operating conditions — like reception with and without a preamp or effectiveness of a power amplifier with a contacted station.

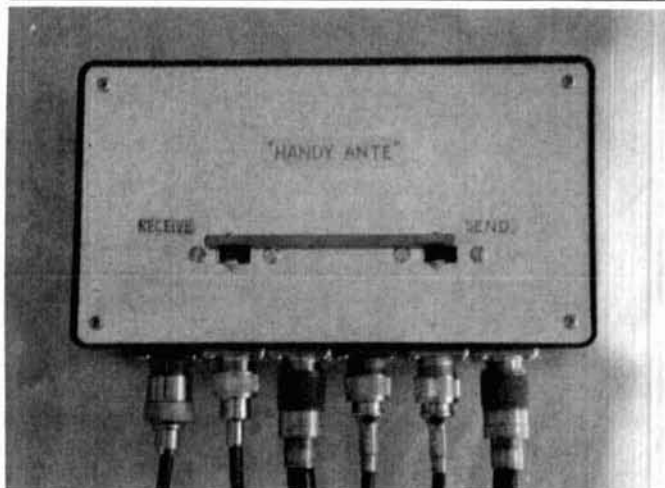
Eventually, wear and tear got the best of the rather puny contacts on the CB switches, and I needed to replace them. But alas, the Radio Shack switches were no longer available, and I couldn't locate any other similar devices. So, I decided to concoct my own. The result was sufficiently successful to warrant passing it on.

FIGURE 1



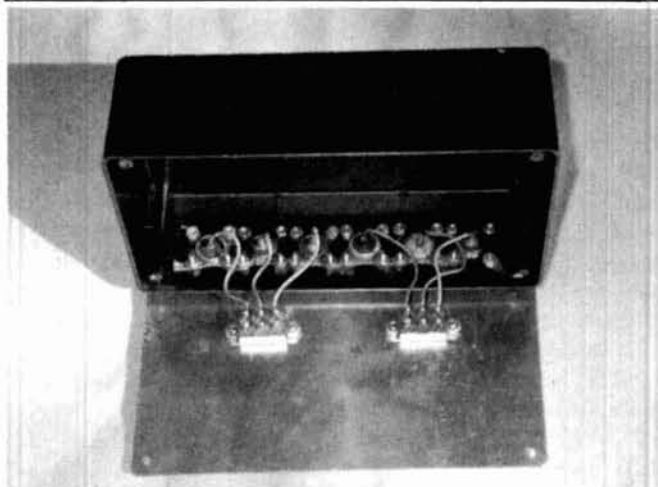
Some typical applications for "Handy Ante."

PHOTO A



The DPDT coax selector switch (dubbed the "Handy Ante") in operating position.

PHOTO B



"Handy Ante" with front panel open. Note lug in lower right corner for grounding panel.

PARTS LIST

Quantity	Description
1	Housing, Radio Shack hobby box no. 270-232, or similar
2	Slide switches, DPDT, 125 volts AC, 3 A, Radio Shack no. 275-403
6	Coaxial cable receptacles, chassis mount, type SO-239, Radio Shack no. 278-201

Description and construction

Essentially, the device consists of two slide switches coupled together mechanically and terminated electrically in six SO-239 coaxial cable receptacles. The construction is quite simple and is shown in Photos A and B. I built it into a Radio Shack 7-3/4 by 4-3/8 by 2-3/8 inch plastic experimenter's box with an aluminum panel. This box is a little large for this application, but it does permit in-line mounting of the six coax receptacles. It's critical to bond

the SO-239 receptacles together electrically; I used copper braid from a scrap of coax. The leads from switches to receptacles should be short.

Switches

The slide switches are double-pole, double-throw, but the poles are paralleled to increase the current capacity. The switches are rated 3 A at 120 volts. Paralleling the contacts should make them good for 6 A, but because they are intended for use in low frequency power circuits, I'd hesitate to give them this rating at RF. I can say from experience that they will handle the output of a 200-watt amplifier and suspect that they will do considerably better.

The switches are coupled by a length of wooden dowel. I used one 1/4 inch in diameter and 3-3/4 inches long. Drill a 5/64-inch hole in each of the plastic switch levers; then mount the switches and carefully mark and drill the dowel to match. I mounted the dowel with no. 2-56 machine screws and hex nuts.

Be sure that both switches are in the same position when you mark the dowel for drilling. If you take reasonable care, the switches will move smoothly and positively from one position to the other, with the dowel used as a handle. I used no. 18 stranded hookup wire to connect the switch. Make sure the leads are short and have sufficient current capacity. If you use a metal panel, connect it to the coax receptacles.

Mount the completed unit solidly to a wall, shelf, or other stable mooring near your operating position. Now all you need to put the "Handy Ante" to work are a few coax jumpers. The coax you choose for these jumpers should have the same characteristic impedance as that of the original transmission line.


Performance

The switch in the circuit at my location seems to have minimal effect on VSWR or general performance of the antenna system. I found there was an increase in VSWR from 1:1 to 1.5:1 at 28 MHz when I inserted the switch in the line. This isn't a major problem and was the maximum change noted between 3.5 MHz and 28 MHz.

A note of caution

If you use the switch to insert a preamplifier or some similar device used mainly for receiving, make sure the device has protection in case you forget to throw the switch to the transmit position to send. I use back-to-back diodes mounted directly across the preamp input terminal from the transceiver. This crowbar-type protection causes the transceiver to transmit into a virtual short circuit and trip off because of its protective system. Other types of protection may be best for different setups.

In closing

You might ask, "Why, in these days of sophistication and automation, should I resort to a manual send/receive switch?" The system is simple and reliable, and lends itself well to on-the-spot comparative testing. The switch also operates easily and quickly. Its only drawback is operation on break-in modes. I use "Handy Ante" on a routine basis to insert my old Class-C amplifier in the feedline when needed. I also use it to insert a 10 or 15-meter homebrew preamp for receiving, combining functions A and B of Figure 1. 

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Microwaves



Bob Atkins, KA1GT

Starting something new is always a challenge. I thought I'd begin this month by introducing myself and talking about the possibilities for this column. As some of you may know, I was the author of *QST's* column on microwave technology from 1980 to 1989. When *QST* decided not to run a regular monthly microwave column late last year, *Ham Radio* graciously offered me the opportunity to join them and continue coverage of the bands above 1 GHz.

I have some ideas about how such coverage might be presented, but I would be interested in what you would like to see. My current idea is to present a mix of operating news, technical presentations, and simple construction projects. These would be aimed mostly at the microwave bands, but would not exclude coverage of topics also related to operation in the VHF and UHF parts of the spectrum. Your input to such a column is *vital* important if it is to present information concerning state-of-the-art developments in microwave equipment and operation. If you have short news items, technical hints, or microwave-related questions, I urge you to send them along to me at the address at the foot of this column. There is a minimum two month lead time between my writing a column and its appearance in *Ham Radio*, so prompt reporting of activity news is especially important.

Ham Radio is always interested in publishing longer technical articles in the main body of the magazine, so if you have such technical material, by all means send it directly to the magazine at the address on the contents page.

It's difficult, in writing a column of this type, to determine the level of knowledge I should assume on the part of the reader. As much as possible, I will aim at general Amateur interest.

There are two reasons for this. First, I want to encourage readership; nothing turns off readers like a page full of equations! Second, I'm not sure I can write at a level more complex than that! Plans for upcoming columns include coverage of optical communications systems including laser transmitters, receivers, and propagation, and articles on the various modes of microwave tropospheric propagation — from line of sight through tropospheric scattering to ducting. Along with these I hope to have some news items. Once again, that depends on reader input. So please write, and between us we'll see what we can do to maintain coverage of the microwave scene.

Please send your questions and comments to me at 103 Division Avenue, Millington, New Jersey 07946.

Meet Bob Atkins

In January's *Editor's Notes* I promised all of you a new column for microwave and VHF/UHF enthusiasts. We know that many of you have missed Joe Reiser's monthly missive "*VHF/UHF World*," so this month we are pleased to present our first installment of "*Microwaves*," written by Bob Atkins, KA1GT. Bob holds a Ph.D. in chemistry and currently works in the field of materials research at AT&T's Bell Laboratories. He was first licensed as G8EKB (a UK no code license) in or around 1969. He was active intermittently on 144, 432, and 1296 MHz until 1976, when he left England for Texas. In 1979 Bob moved to Connecticut, was licensed as KA1GT, and operated SSB/CW on 144, 432, and 1296 MHz and wideband FM on 10 GHz. From 1979 to 1981, he was also active on 432-MHz EME using a homebrew amplifier and antenna system. Bob made his first contacts on 10-GHz SSB/CW in 1981. Later that year he moved to New Jersey, where he can be found on the air from time to time on the VHF/UHF and microwave bands. He enjoys building equipment, and has recently completed construction of a fully relay-switched solid-state 10-GHz all mode transverter system for portable use. We hope you enjoy Bob's column and that you'll send him your comments and suggestions.

KA1STC

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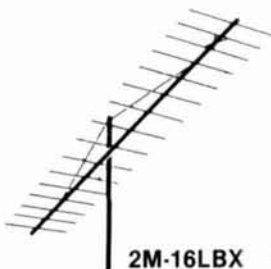
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2M—16LBX

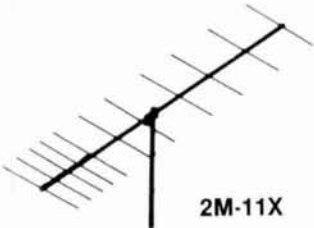
Bandwidth.....144-148 MHz
Gain.....13.2-14.5 dBd
VSWR.....1.2:1
F/B.....20 dB min.
Element Length...40% max.
Boom Length.....28 ft.
Windload.....2.44 sq. ft.
Turn Radius.....15 ft., 5 in.
Weight.....10 lbs.



2M-16LBX

2M—11X

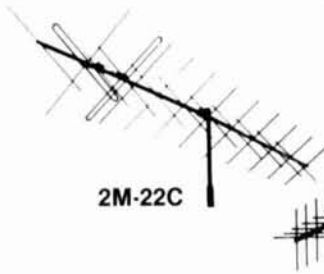
Bandwidth.....144-148 MHz
Gain.....12.5 dBd
VSWR.....1.2:1
F/B.....20 dB min.
Element Length...40% max.
Boom Length.....15 ft., 4 in.
Windload.....1.38 sq. ft.
Turn Radius.....15 ft., 4 in.
Weight.....5.5 lbs.



2M-11X

2M—22C

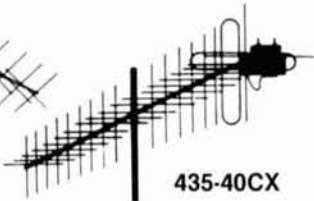
Bandwidth.....144-148 MHz
Gain.....13 dBc
VSWR.....better than 1.5:1
F/B.....20 dB min.
Element Length...41" max.
Boom Length.....19 ft., 1 in.
Windload.....1.85 sq. ft.
Turn Radius.....13 ft.
Weight.....11 lbs.



2M-22C

435—40CX

Bandwidth.....410-450 MHz
Gain.....15.2 dBc
VSWR.....1.5:1
F/B.....20 dB min.
Element Length...13.625" max.
Boom Length.....175.5 in.
Windload.....1.16 sq. ft.
Turn Radius.....105 in.
Weight.....10 lbs.



435-40CX

Antenna gain figures used by MIRAGE/klm are measured according to the National Bureau of Standards. These figures are actual gain listed in dB at free space measured over a dipole. Some manufacturers do not choose to use this standard and therefore indicate gain figures that appear to be higher.

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SWITCHABLE BANDWIDTH CRYSTAL FILTER

CW and SSB filter with one set of surplus crystals

By John Pivnichny, N2DCH, 3824 Pembroke Lane, Vestal, New York 13850

You can design a filter for almost any bandwidth by selecting the proper coupling capacitors and termination resistances. Here's how to build a filter with two different bandwidths (CW and SSB) by switching only the capacitors. The termination resistance is held constant at 220 ohms and doesn't need to be switched.

Ladder-type filters

Simplified design equations for ladder-type filters where all crystals have the same frequency have been published in Amateur Radio literature.¹ **Figure 1** shows a typical ladder filter. (See **Reference 2** for more information.) The crystals are in series with the signal path and are "coupled" by capacitors with one end connected to ground. The filter is terminated with resistances at each end.

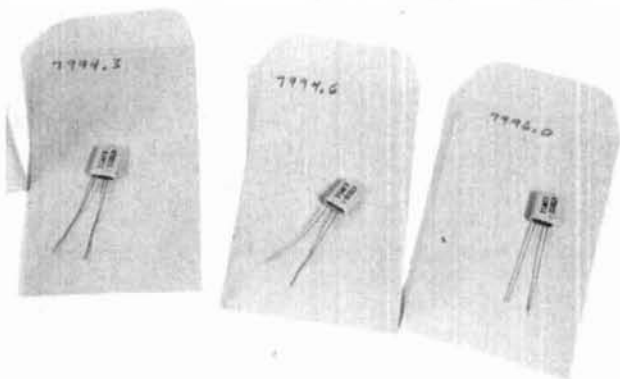
Crystals

My filter uses inexpensive surplus 8.0-MHz crystals in an HC-18 style package with wire leads (see **Photo A**.) I was able to get 50 of them through a swap and checked their frequencies before proceeding.

You can check your crystals by building the oscillator circuit shown in **Figure 2** and **Photo B**. The terminal posts allow for easy connection and removal of each crystal's wire leads. I wrote frequencies to the nearest 100 Hz on a set of small envelopes and placed each crystal in its marked envelope directly after removing it from the test circuit. My envelope system prevents mixups and lets me select appropriate crystals for the filter.

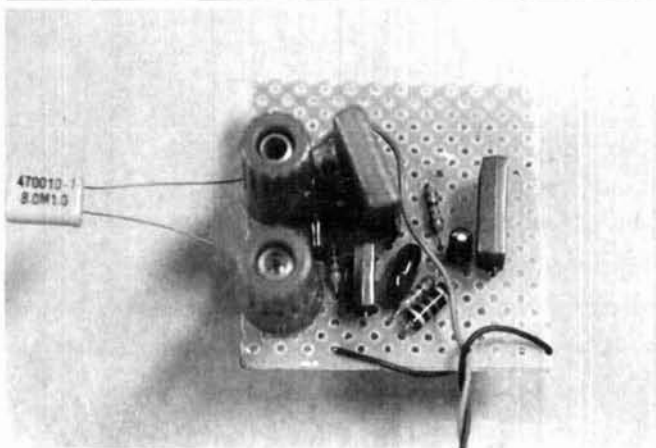
You'll need three crystals with frequencies within 200 Hz of each other. In most cases, it isn't necessary to start with 50 crystals. My crystals turned out to be very low Q, with frequencies ranging from 7995 to 8015 kHz. Nevertheless, I wound up with many sets of three within the required 200 Hz. The surplus crystals, although low in Q, make very satis-

PHOTO A



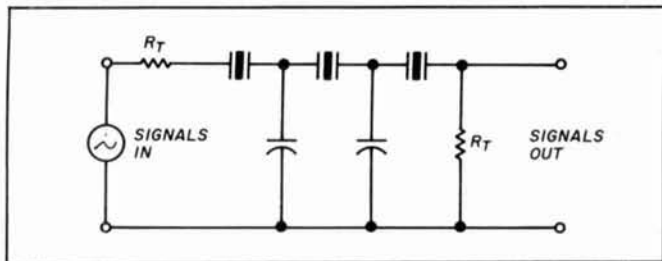
Surplus crystals.

PHOTO B



Oscillator circuit.

FIGURE 1



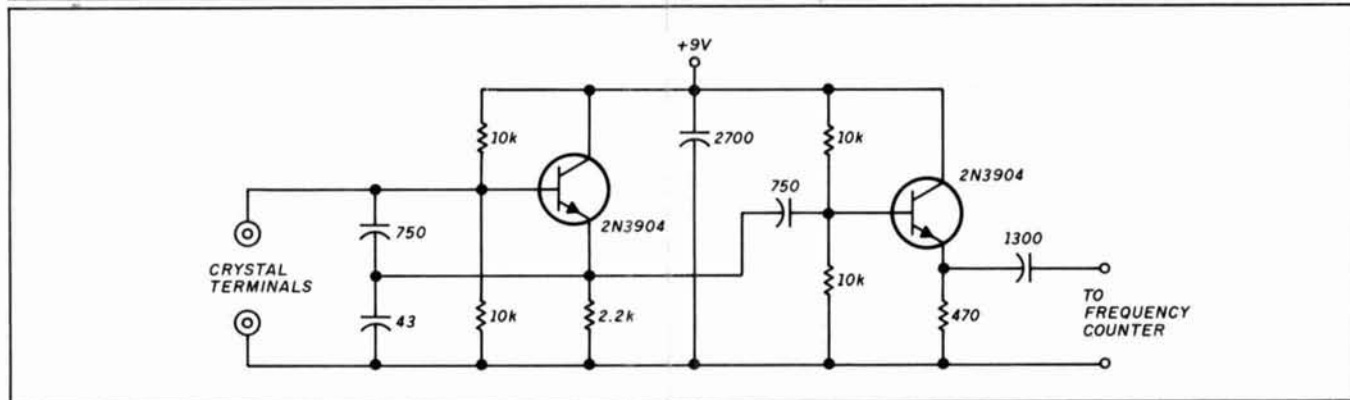
Typical ladder filter.

factory filters. The main difference you'll notice when using low-Q crystals is the higher insertion loss for narrower (CW) bandwidths. You can compensate for this loss in the IF amplifier that follows the filter.

Test circuit

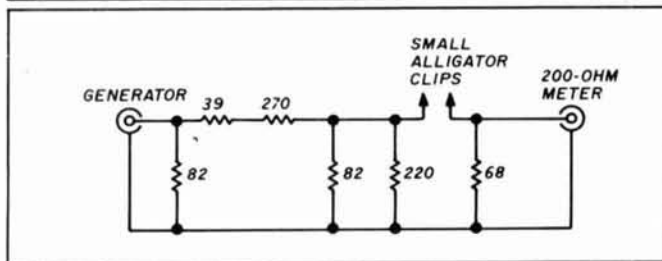
Once you've selected three crystals, you must measure their characteristics more carefully. You'll need a signal source whose frequency is adjustable ± 20 kHz around the crystal frequency and a sensitive RF voltmeter. I used the circuit and hardware of my homebrew tuning dial³ (adding

FIGURE 2



Oscillator circuit.

FIGURE 3



50-ohm test circuit.

240 pF to the tank and adjusting the coil slug), which I tuned to 8 MHz for a signal source. My voltmeter is described in Reference 4. You'll also need a calibrated 3-dB attenuator⁵ and the 50-ohm test circuit in Figure 3.

You must have a 50-ohm system to measure the crystals because the design equations are based on 50-ohm measurements. The pi attenuator at the input of Figure 3 guarantees a source between 50.05 to 51.82 ohms, regardless of your signal source impedance. At the output, I used the 200-ohm attenuator⁵ in parallel with a 68-ohm resistor, followed by a 200-ohm voltmeter.

Following Hayward's procedure¹, insert each of the selected crystals into the test circuit shown in Photo C and tune the signal source for a peak reading on the voltmeter. This peak will occur about 1.8 kHz lower than the frequency measured with the oscillator circuit of Figure 2. Note the reading on the voltmeter. Now replace the crystal with low value fixed resistors (10 ohms is a good starting point) until you obtain the same voltmeter reading. This value will be the series loss (R_s) of the crystal. My surplus crystals measured 27 to 30 ohms; a high quality HC-6/U type will measure 5 to 6 ohms. Record the R_s measurement.

Replace the crystal in the test circuit, checking to make sure the signal source is still set for a peak reading on the voltmeter. Switch in 3-dB attenuation, note the new meter reading, and then switch back to 0 dB. Now tune the signal source above and below the peak frequency until you reach the same new meter reading. Record these two frequencies.

PHOTO C



Test circuit.

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The difference between the two frequencies will be the bandwidth Δf (in hertz) of the crystal at the 3-dB points. My surplus 8-MHz crystals measured 1230 Hz. Hayward reported 294 Hz for his 3.5-MHz crystals and 96 Hz for high quality 3.5-MHz crystals.

SSB filter calculations

Design the SSB bandwidth filter (2.5 kHz) first, using the equations which follow. I took them from Hayward's article.¹ The circuit shown in **Figure 4** is a Butterworth three pole, tuned with series capacitors in the end loops. All capacitors and resistors have the same value.

$$C = 1326 \left[\frac{\Delta f}{B \times 0.707 \times F_0} \right] - 10 \text{ (pF)} \quad (1)$$

$$R_{end} = \left[\frac{120 \times B}{\Delta f} \right] - R_s \text{ (ohms)} \quad (2)$$

where B = desired filter bandwidth in Hz
 F_0 = crystal center frequency in MHz
 Δf = bandwidth measured in test circuit in Hz
 R_s = crystal series resistance measured in test circuit in ohms

Using my crystals with $\Delta f = 1230$ Hz, $F_0 = 8$ MHz. The calculated values are $C = 105$ pF and $R_{end} = 217$ ohms.

You can calculate the crystal's equivalent series circuit elements from **Equations 3** and **4**. A holder shunt capacitance of 5 pF is assumed. (See **Figure 5**.)

$$C_m = 1.326 \times 10^{-15} \left[\frac{\Delta f}{F_0^2} \right] \text{ (farads)} \quad (3)$$

$$L_m = \frac{19.1}{\Delta f} \text{ (henries)} \quad (4)$$

where:

C_m and L_m are the equivalent series capacitance and inductance, respectively.

I used the equivalent circuit to compute the frequency response of the filters in this article aided by a PC circuit analysis program.⁶ With 8 MHz as F_0 , I calculated $C_m = 0.0255$ pF and $L_m = 0.01552$ H. The computed and measured response of crystal no. 3 in the 50-ohm test circuit is shown in **Figure 6**. I also ran the analysis program for the SSB filter with and without the end loop tuning capacitors. These results are shown in **Figure 7**. It appears that the tuning capacitors eliminate the dip in the center of the passband. They also shift the overall response 500 Hz higher on the low frequency side and about 250 Hz on the high frequency skirt. This narrows the bandwidth slightly.

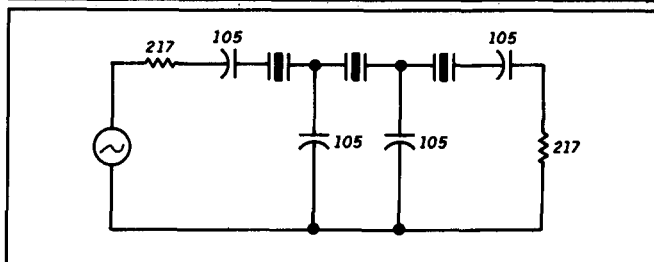
CW filter calculations

A much narrower bandwidth is desirable for CW. I chose 500 Hz, and using **Equations 1** and **2**, calculated $C = 565$ pF; $R_{end} = 21.8$ ohms. Note that Hayward¹ recommends that the crystal unloaded Q (Q_u) exceed the filter Q by a factor of 10.

$$Q_u = \frac{1.2 \times 10^8 \times F_0}{\Delta f \times R_s} \quad (5)$$

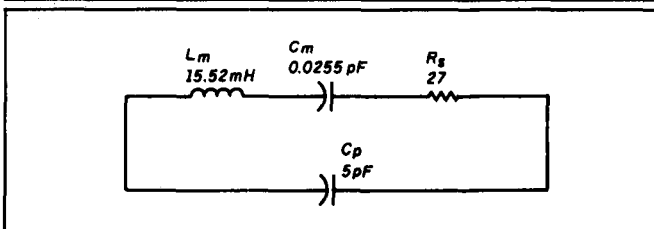
For my crystals, this was $Q_u = 28,900$. A 500-Hz bandwidth at 8 MHz represents a filter of $Q = 16,000$, so a factor of

FIGURE 4



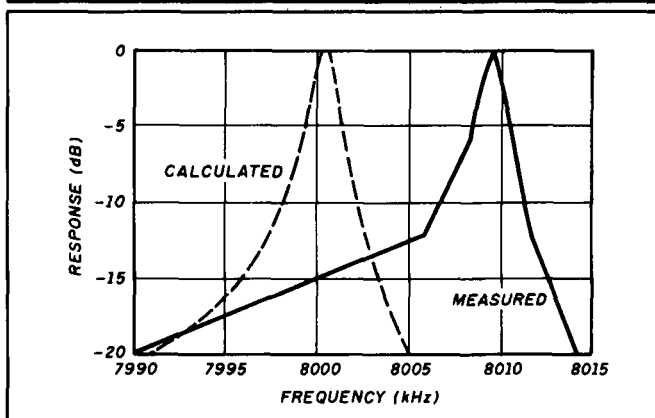
2.5-kHz SSB filter.

FIGURE 5



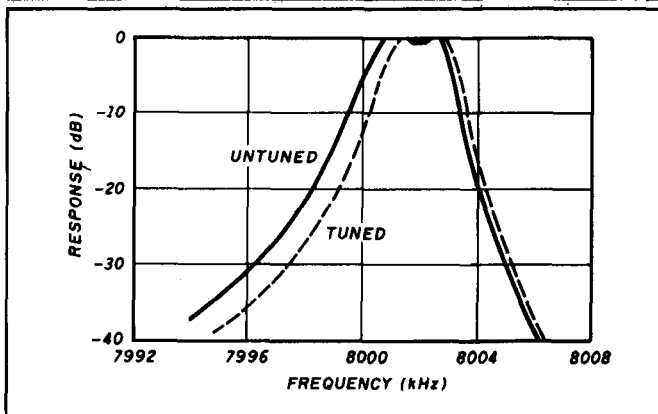
Crystal equivalent circuit.

FIGURE 6



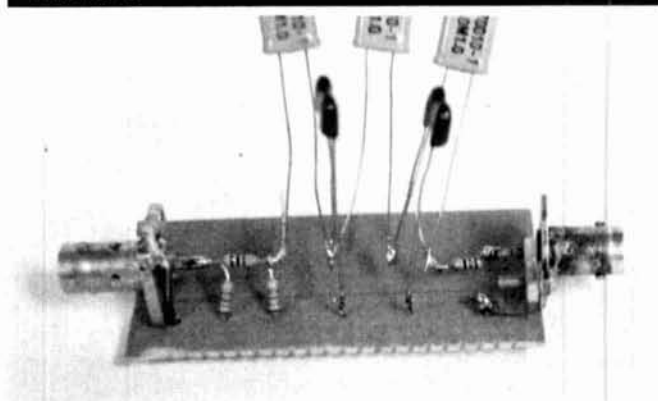
Test circuit response.

FIGURE 7



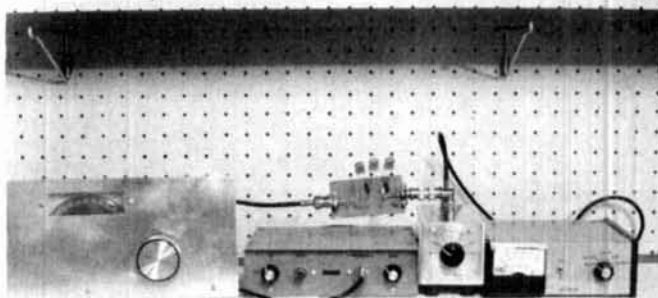
SSB filter, calculated data.

PHOTO D



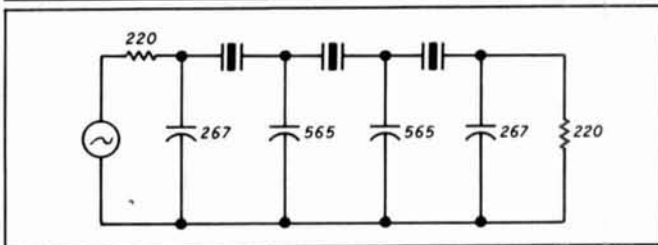
Trial filter.

PHOTO E



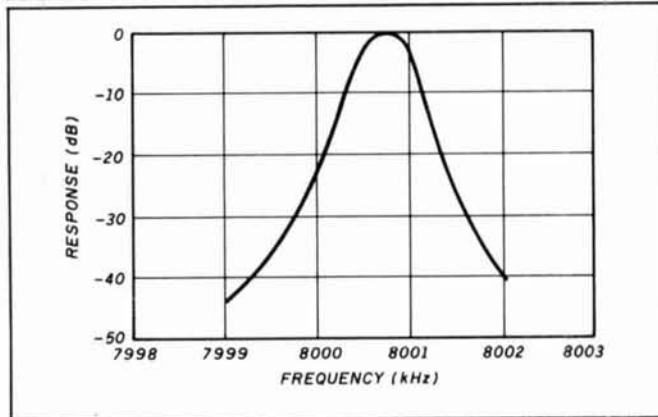
Measurement setup with trial filter.

FIGURE 8



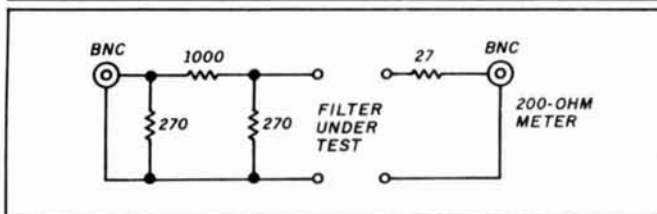
0.5-kHz CW filter.

FIGURE 9



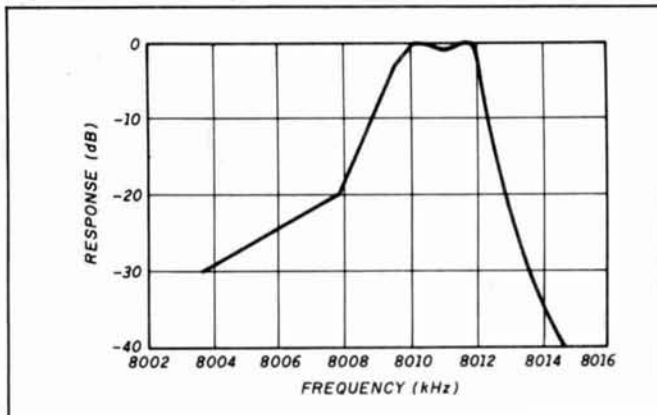
CW filter, calculated data.

FIGURE 10



220-ohm trial network.

FIGURE 11



SSB filter, measured data.

10 isn't possible with these low quality surplus crystals. There's barely a factor of 2 in this case. Nevertheless, with typical Amateur resourcefulness, I used the crystals anyway. As it turned out, they work just fine. However, the insertion loss is somewhat high.

The calculated resistance for the CW filter is stepped up to match that of the SSB filter with a shunt capacitor at the ends of the filter:

$$C_{end} = \left[\frac{1.59 \times 10^5}{R_o F_o} \right] \times \sqrt{\frac{R_o}{R_{end}}} - 1 \quad -5 \text{ (pF)} \quad (6)$$

where R_o is the desired (SSB) resistance. This equation derived by Hayward lets you keep the termination resistance constant and switch capacitor values only when going from SSB to CW bandwidths.

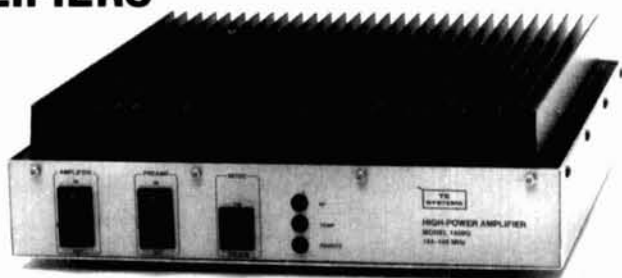
The filter shape and insertion loss aren't changed by this impedance step up. For my crystals, $C_{end} = 267 \text{ pF}$. The filter circuit and its computed response are shown in Figures 8 and 9.

Trial measurements

I verified the computed responses for both filters with the trial measurement setup shown in Photos D and E. I connected the crystals and capacitors for the SSB filter together on a piece of perforated circuit board by temporarily soldering the ends of their leads to the 220-ohm trial network (see Figure 10). This construction method leaves large inductive loops which could alter the filter characteristics. However, it didn't seem to have much effect at the 8-MHz frequency.

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1452G	144-148	25	400	.6	15	13.6	50	UHF
2252G	220-225	25	250	.7	14	13.6	36	UHF
4450G	420-450	10	180	1.1	12	13.6	39	N
4452G	420-450	25	180	1.1	12	13.6	36	N

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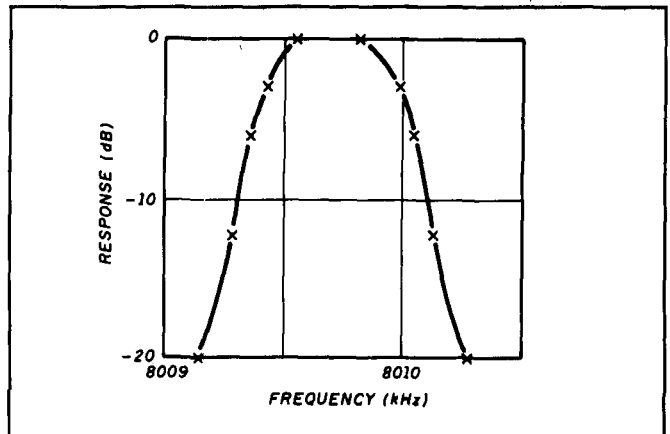


Take measurements by first adjusting the signal generator for the peak response on the RF voltmeter. Add a fixed amount of attenuation with the step attenuator and note the new voltmeter reading. Then switch the attenuator back to 0 dB and adjust the signal generator above and below the peak frequency until the RF voltmeter reading is at the level you noted. Read these two frequencies on the frequency counter. Any error in the RF voltmeter calibration is immaterial with this procedure. Accuracy depends only on the attenuator accuracy and frequency counter calibration.

My step attenuator has 3, 6, 12, 20, 30, and 40-dB steps. The measured results for both the tuned and untuned (series capacitors shorted) SSB filter are shown in Figure 11. Correlation with the calculated data of Figure 7 is very good.

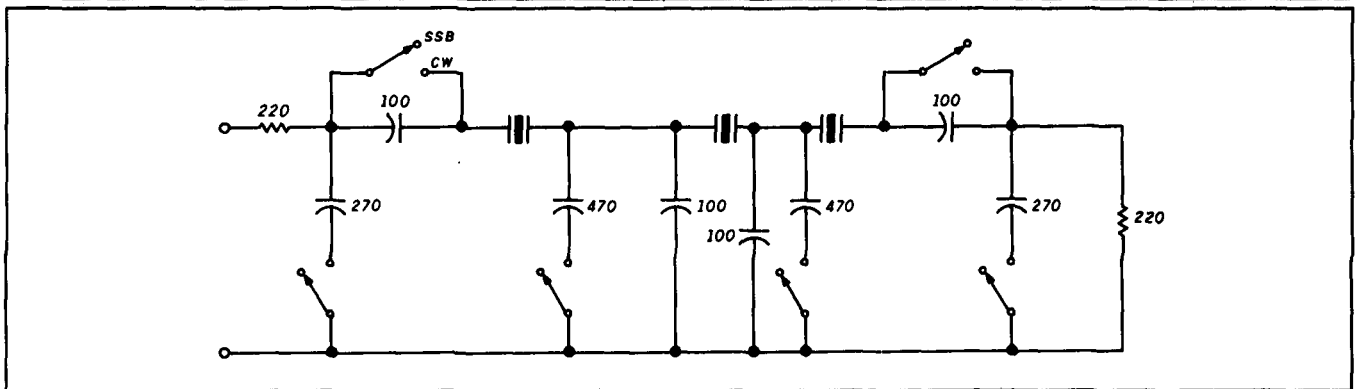
Now solder additional capacitors by their lead ends onto the SSB circuit to form the CW filter circuit. Figure 12 shows

FIGURE 12



CW filter, measured data.

FIGURE 13



Switched bandwidth circuit.

the circuit's measured response. Again, there's good correlation with the calculated data (Figure 9). The measurements also confirm that low-Q crystals can be used to create narrow bandwidth filters.

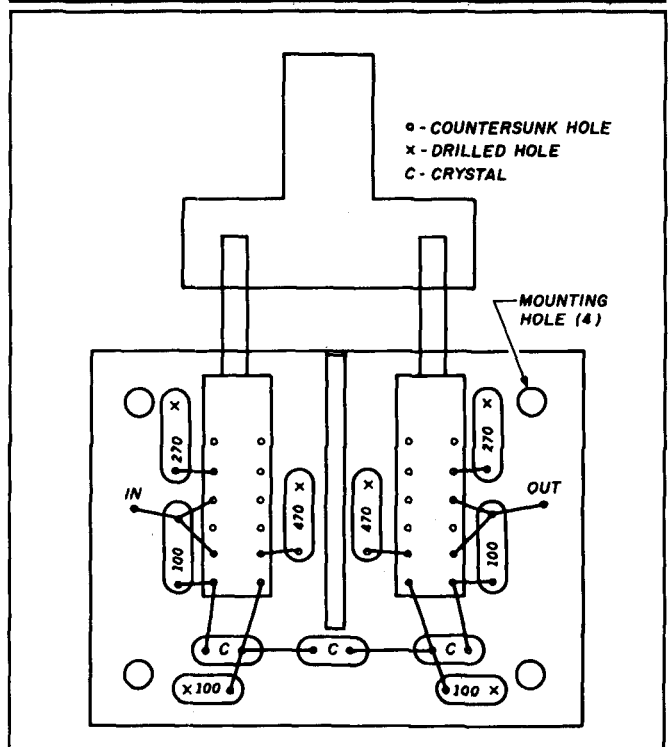
Constructing the switched filter

In a high frequency receiver, you must be able to switch from an SSB to a CW bandwidth. The overall switching circuit is shown in Figure 13. Six sets of switch contacts must close when going from SSB to CW. I used two four-pole push switches (GC Electronics catalog no. 35-492) and mounted them on a 2 x 2.5-inch piece of single-sided copper-clad circuit board. Two unused sets of contacts are available to operate indicator lights, adjust BFO frequency, or adjust the IF gain. As mentioned earlier, the insertion loss of the CW filter is about 8 dB greater than the SSB filter.

Component placement

Mount all components, including the switches, on the copper foil side of the board. Insert the leads in drilled holes, which are countersunk on the top side if the lead connection isn't connected to ground. This copper foil forms an excellent shielding ground plane. Insert component leads to ground into a drilled hole. Then solder the leads to the copper foil with a surrounding bead of solder on the top side. Interconnect the remaining component leads on the bottom side as required, to form the overall circuit.

FIGURE 14

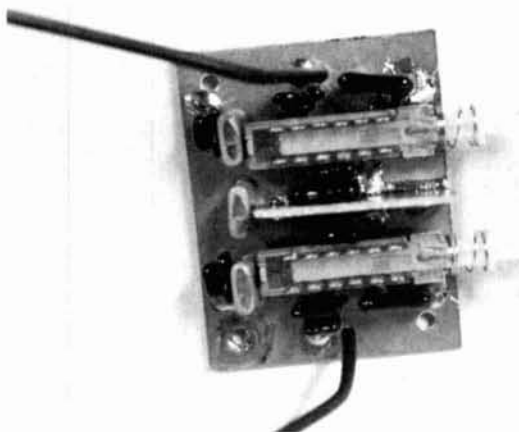


Component locations.

Component locations

Component locations are shown in **Figure 14**. Drill four large corner holes for mounting the filter behind the front panel in your receiver. Use short lengths of miniature coaxial cable for the input and output connections. Use a small piece cut from 1/4-inch Plexiglas™ to form a single push-button which lets you operate both switches simultaneously. Glue it to the push switches with plastic pipe cement. If you look at **Photo F**, you'll notice that I used two dipped silver mica capacitors where you might expect to see one. There's no magic here. I didn't have the correct value on hand and made it up with two other capacitors. There's no need to match capacitor values exactly; I didn't even attempt to do so. I simply soldered in standard value silver micas. Solder a shield between the two switches made from a 3/4 x 1-1/2 inch piece of the single board stock.

PHOTO F




Completed switchable bandwidth filter.

Final check

A final check of the switched filter duplicated the results measured in the trial construction. Even though the component lead length was considerably shorter, and there was shielding provided by the copper planes, I noted no differences.

Acknowledgments

I want to thank Wes Hayward, W7ZOI, for working out and publishing the design equations reproduced in this article. They made the job of constructing a crystal filter a straightforward procedure. Also I'd like to thank my associate Dr. Kischen Kapur, who doesn't have an Amateur license but likes to go to hamfests to buy and sell parts. Dr. Kapur supplied the crystals used in this article. 

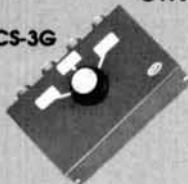
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Simple Circuit Prevents Amplifier Control Relay Failure

The switching requirements of many linear amplifiers exceed the current rating of the relay in the driver stage. Immediate or premature relay failure can result when the contacts stick in the closed or transmit condition.

Once you've replaced the relay in the radio, your choices are buying an intermediate relay assembly (if you can find one), providing your own relay and wir-

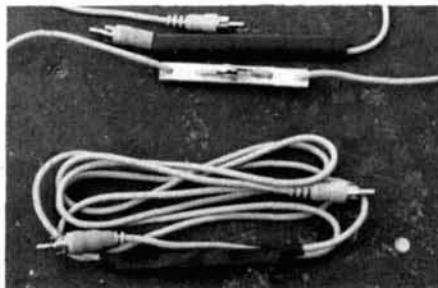
ing, or assembling a solid-state switch.

After looking into all the alternatives, I found my own solution. It's really quite simple. All you need is one transistor, an RCA terminated patch cable, a strip of metal, and some heatshrink tubing (see Photo A).

a coil is related to the rate of change of the current running through it.)

I tried several different transistors and recommend MPSA44s or MPSA45s for switching negative voltages and 2N6519s or MPSA92s for switching positive voltages. The transistor receives all its current from the linear switch line and operates in saturation mode with less than 1 volt from emitter to collector, so very little power is dissipated. High voltage NPN transistors have higher gain than similar PNP transistors. This means the radio's relay is presented with one-fifth the switched current with PNP transistors and about one-twentieth the current with NPN transistors (which I used).

PHOTO A



Simple hardware for providing protection to linear amplifier control relay.

The transistor

You'll need an NPN transistor for switching negative voltages or a PNP transistor for switching positive voltages, as required by your amplifier. The ideal transistor should have a low collector saturation voltage and be a high voltage switching type. Linear amplifiers seldom have transient suppression diodes on the relay coils. This causes spikes of many times the supplied voltage when the radio's relay opens. (Remember the voltage across

Construction

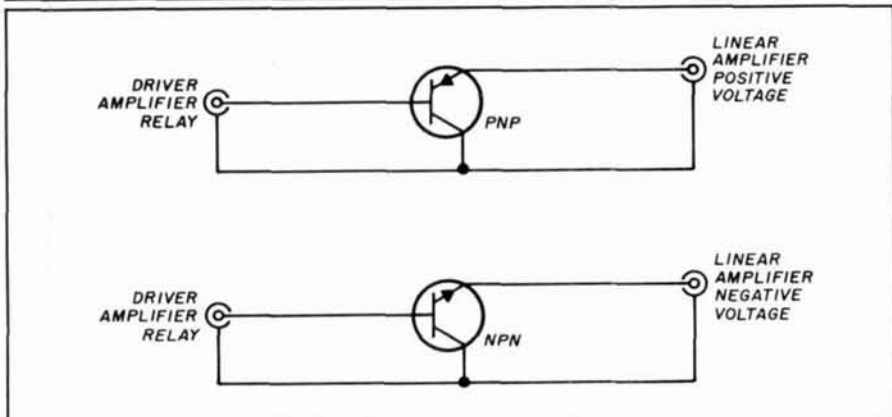
Drill two holes in each end of a 3/8 x 3-inch aluminum strip. Glue the transistor to the center of the metal strip with silicone seal. Cut the patch cable in the center and strip the wires for soldering. Stuff the stripped ends through the outer holes first and then through the inner ones to provide attachment to the metal strip. Solder the wires to the transistor for the circuit in Figure 1. If you were thinking ahead, you've already stuffed the heatshrink onto one of the cables so you can slide it over the assembly and shrink the tubing for protection. For added protection, cover the transistor and connections with silicone seal. I shrunk a small band on the linear's end plug, using blue for positive voltages (PNP) and black for negative voltages (NPN).

The switching cable is quiet and fast. It's simple to install and leaves your accessory socket free. You might consider using it to increase relay life, even where immediate or premature failure isn't evident.

The assembled NO7G+ or NO7G- cable is available for \$15 plus \$2 shipping and handling. Send your requests to me at 525 26th Avenue S., Seattle, Washington 98144.

David Smith, NO7G

FIGURE 1



Solid-state switch protects relay in driver amplifier.

Stop Your TH3 Junior Drooping

Recently I took possession of a TH3 Junior Yagi which was looking rather tired. I tried the old method of giving the tubing in the the elements half a turn and, although it looked better, it still drooped.

I made up three tubes (as in Figure 1) using the bolt in place of the anchor bolt in the element-to-boom bracket. The dowels are 5/8 inch in diameter and 12 inches long.

The braided rope (nonconductive) is

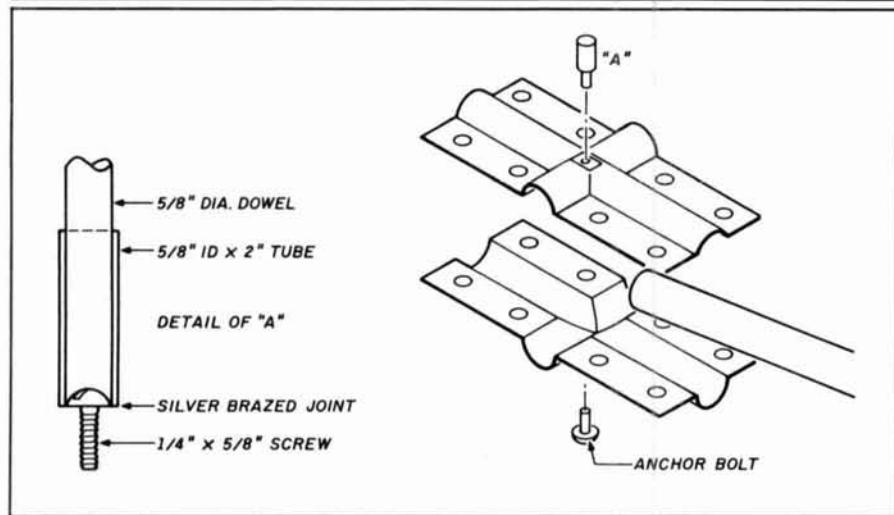
5/32 inch and is tied off outboard of the 15-meter traps (see Figure 2). Allow the rope to stretch under tension before putting the Yagi on the tower.

Perhaps this idea could be employed on the bigger Yagis if you use stronger materials. It works well on the TH3 Junior and deters large birds from perching on the elements.

By Arthur Brean, VK6SY

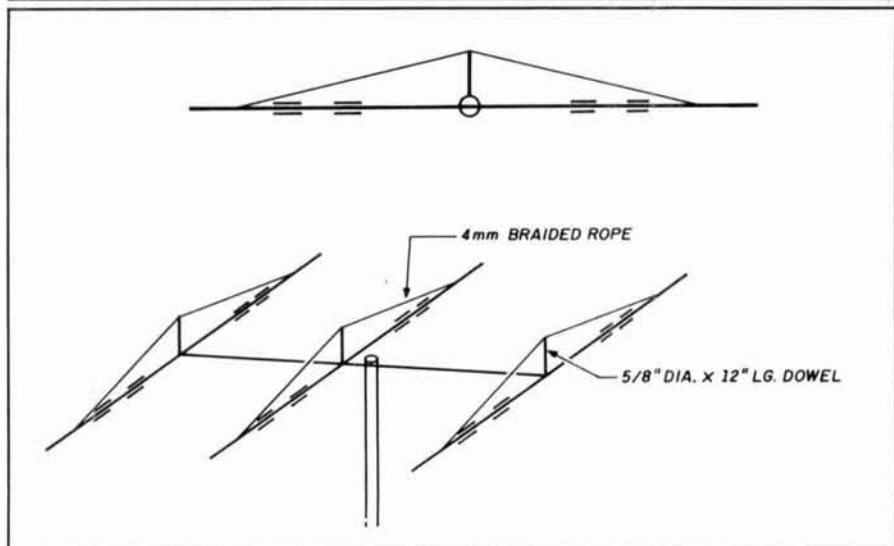
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FIGURE 1



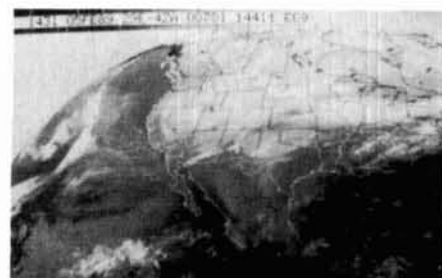
Boom-to-bracket element.

FIGURE 2



Braided rope (nonconductive) tied off outboard of the 15-meter traps.

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INEXPENSIVE HALF-WAVE 2-METER MOBILE ANTENNA

By Glen Noble, WE7C, 4840 Schindler Road, Fallon, Nevada 89406

I needed a 2-meter mobile antenna, so I thought: "Why not build it myself? How hard could it be, anyway?" A search of the garage yielded an old gutter-mount CB antenna which could provide a mount and coax, but the whip looked hopeless for 2-meter use.

The hardware

I went to Radio Shack to look for a suitable whip and was pleasantly surprised when I found a stainless steel CB whip exactly the right size for a half-wave 2-meter antenna.

Now that I had the new half-wave whip and the old CB gutter-mount frame, my project was on a roll. While browsing at Radio Shack, I discovered that gutter-mount frames and other styles, like mirror mounts and single-hole types, were available at low cost without attached CB antennas.

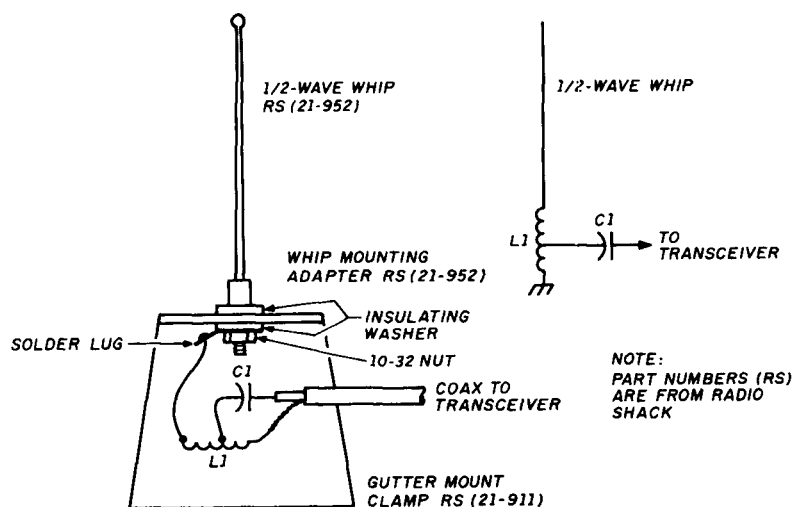
The impressive performance of my AEA "Hot Rod" half-wave handheld antenna heightened my enthusiasm for this project. I completed my design after consulting the RSGB *VHF/UHF Manual*.¹ It showed me how to turn my growing collection of parts into a usable antenna by adding a coil of wire and a small capacitor.

How it works

The main obstacle when building a half-wave whip antenna lies in matching its high input impedance to 50 ohms. Because the input impedance is high, you can't simply connect one end of the coax to the whip while grounding the other end, as you would for a quarter-wave antenna. However, if you add a six-turn coil tapped at two turns, along with a small capacitor, you can form a matching network that transforms this high impedance to the required 50 ohms.

The coil is the principal means of transforming impedance; the capacitor compensates for a reactive component introduced by the coil. Figure 1 shows the connections and parts used. The resulting antenna has good bandwidth and covers most of the 2-meter band with acceptable VSWR.

FIGURE 1



C1: 10-15 pF 500V MICA CAPACITOR

L1: 6 TURNS No.18 WIRE TAPPED AT 2 TURNS. 3/8" DIAMETER FORM, 1 1/4" LONG

Pictorial and electrical schematic of the 2-meter half-wave antenna.

PHOTO A



Matching coil is located under the mount to inhibit radiation.

When building this antenna, I did deviate from the usual construction methods by locating the matching coil under the mount, where it can't radiate (see **Photo A**). This coil is usually placed above the mount, where it can contribute to signal strength on both transmit and receive. However, casual comparisons of this antenna with my AEA "Hot Rod," with the coil above the mount haven't revealed any major differences. Installing the coil under the mount does simplify construction significantly.

It's interesting that the half-wave antenna is thought to perform better than a 5/8-wave whip when a good ground plane isn't available. This could be quite helpful on a gutter-mount antenna, where half the ground plane is missing.

Construction

The stainless steel CB whip, Radio Shack part no. 21-952,* provides the 39-inches necessary for a half wave on the 2-meter band. My gutter mount was an old Radio Shack CB antenna similar to the current part no. 21-909. These parts cost about \$6 each.

Start your construction by modifying or fabricating an insulating washer to insulate the base of the whip from the top of the mount. A similar washer is needed to insulate the nut and lug from the bottom of the mount. The top washer of my gutter mount had a lip that I fitted into the mounting hole to center the whip in the opening. The bottom washer was a typical flat plastic one. I had to thin these washers so the 10-32 threads on the whip would extend far enough through to allow me to attach a nut. I used a thin blade saw to cut one of the washers in half; you can also sand or grind them down. You may have to make these washers if you can't salvage them from a used antenna.

Next, wind the coil on a 3/8-inch diameter wooden dowel form. You need six turns of no. 18 wire, 1-1/4 inches long, with a tap at two turns. The series capacitor doesn't appear to be critical in terms of capacitance. The RSGB manual suggested 15 pF, but the 10 pF I had available worked fine.

*Replaces Radio Shack part nos. 21-904/908/940. Ed.

Finally, wire the assembly as shown in **Figure 1**.


Adjust the completed antenna for minimum VSWR. You can do this by compressing or expanding the coil turns, as well as adjusting the whip length. When you're satisfied with the VSWR over the band, use epoxy glue to cement the coil turns firmly in place and waterproof the wooden dowel form. It's worthwhile to monitor the VSWR when attaching the coil, as the epoxy may have a slight effect on the VSWR. Tweak the coil, if necessary, to optimize the VSWR before the epoxy hardens.

After the epoxy hardens, use a dab of silicone sealant to hold the whole coil in place and waterproof the coax. It's a good idea to leave a little extra whip length within the mounting adapter. This will allow for some adjustment when you mount the antenna on your the vehicle, or if the sealant has any effect on your VSWR. **Photo A** shows the completed antenna.

Closing remarks

With all the half-wave whips and mounts available, and all the used CB antennas lying around, it's easy to build VHF mobile antennas. It should also be possible to convert an AM/FM antenna using this approach. The Radio Shack half-wave whip screws into the base of the AM/FM antenna on my Toyota. The addition of a coil and capacitor at the base of the antenna in the fender well should produce good results on 2-meters.

The only real problems in converting CB antenna parts to VHF mobile use appear to come from adapting the various mounting thread sizes and the possibility of molded-in base loading coils.

Next time you need a VHF mobile antenna give this approach a try. In less than a day you can have the satisfaction of homebrewing a quality antenna using inexpensive, readily available parts. 

REFERENCES

1. G.R. Jessop, G6JP, *VHF/UHF Manual*, Fourth Edition, Radio Society of Great Britain, 1985, page 8.33

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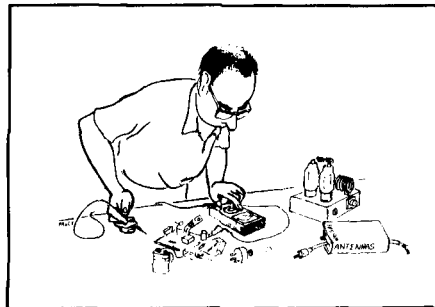
THE MN ANALYSIS PROGRAM (THAT WAS THEN. THIS IS NOW)

In 1935 Marshall Mims, W5BDB, described a startling new beam antenna concept in *QST* magazine.¹ It was a two-element rotatable Yagi made of aluminum tubing elements. The antenna had remote control rotation and a direction-indicating system. Nothing like it had ever been seen before in Amateur Radio!

The Mims beam took over a year to perfect; four designs were built and discarded. Number five worked, after a fashion. Number six seemed better. By the seventh model, the beam was ready for action. It had quarter-wave spacing between the driven element and the reflector, and provided about 2.7-dBd gain and a front-to-back ratio of 10 dB.

Or so Mims thought. Running his 1935 design through a modern 1989 computer program shows that the pioneering ham band Yagi provided 4.3-dBd gain and a front-to-back ratio of about 12 dB. Mims made a pretty good beam and an excellent guess about its characteristics. His data was based upon measurements made by interested hams, using the S-meter of their receivers. (*Editor's note: Actually, "R-meters" were used in those days.*)

Antenna design and measurement techniques had a crude and shaky beginning but were refined over the decades. What was once thought to be rather routine field measurement proved quite tricky and complicated if reproducible antenna measurements were desired. Over the years the military and manufacturers built some large and expensive antenna ranges to measure the characteristics of HF and VHF antennas, but Amateur



antenna measurements were rough-and-ready.

The computer enters the picture

The advent of the powerful digital computer soon provided a new insight into antenna operation. In 1968 an analysis technique known as "moment methods" was publicized.² This scheme dealt with the investigation of electromagnetic fields using computer techniques. Those who remember their high-school integral calculus will no doubt recognize the concept. But the new idea didn't catch fire until computer power was generally available at low cost.

The moment method provided the know-how to translate theory into practice. The job at hand was to provide a good computer analysis program for transmitting antennas.

The birth of the home computer brought a new level of antenna analysis. Some computerized methods for calculating antenna properties were based upon FORTRAN programs which used simple approximations for mutual and self-impedance to calculate element currents in arrays. The magnitude and phase of the element currents were then combined to produce moderately accurate radiation patterns. One of the most popular and well known of these programs was outlined and used by James Lawson, W2PV, in his series of articles³ and his book.⁴ Another contributor to this field was Stanley Jaffin, WB3BGU.⁵

About the same time, the Lawrence Livermore Laboratory at Livermore, California was manipulating a different,

more powerful antenna analysis program — Numerical Electromagnetic Code (NEC) — a mainframe program which would be used to analyze electromagnetic fields.⁶

A derivation of the program (MININEC) was refined at the Naval Ocean Systems Center, Point Loma, California. It applied specifically to antenna analysis using IBM personal computers. Unfortunately, it wasn't user friendly and was too complex for everyday Amateur use. Even so, it was a gigantic step in the right direction.

The K6STI MN antenna analysis program

Brian Beezley, K6STI, took the MININEC program and modified it for general Amateur work. He retained the original antenna modeling algorithm but optimized the code for higher performance. He then massaged the program to make it more applicable to Amateur service. The latest version of K6STI's program, MN 2.0, is the subject of this column.

The MN approach to antenna analysis can be illustrated by considering a straight conductor, like the dipole shown in **Figure 1**. The dipole is broken into segments for examination. The number of segments depends upon the complexity of the antenna and may be chosen by the user. In this case, ten segments are used. More segments would give higher accuracy to the computations, but would also increase computer time.

Each pair of wire segments defines a rectangular current pulse. As shown in the illustration, the current is modeled as uniform within each pulse. The current pulses are centered on segment boundaries and are the same length as the segments. The amplitude of the pulse closely approximates the amplitude of the current in the dipole at that point, as long as enough current pulses are used. The collection of pulses approximates the dipole current, which in this case is the classic half

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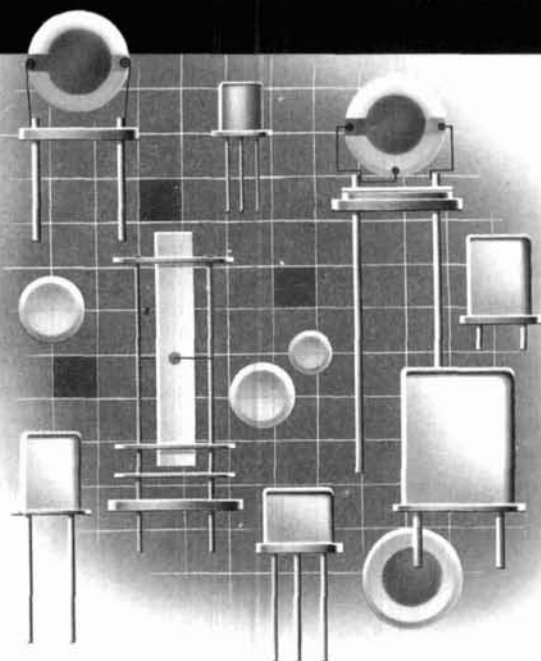
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TABLE 2

The X-Y coordinates of element ends.						
Element	End point no. 1 Coordinates			End point no. 2 Coordinates		
	X	Y	Z	X	Y	Z
Reflector	-10.7	-17.32	0	-10.7	17.32	0
Driven	-1.28	-16.6	0	-1.28	16.6	0
Director	10.7	-15.67	0	10.7	15.67	0

full pulses per element is the number of segments minus one. Thus, there are nine pulses in the reflector. There are also nine pulses each in the driven element and in the director. The source is located by counting the pulses, starting at one end of the first element of the antenna. In this case pulses are counted starting with the reflector, followed by the driven element. That is, nine reflector pulses plus one-half the number of driven element pulses specify the source. This falls at pulse 14 (nine reflector pulses plus five driven element pulses). Half pulses aren't counted. (If an odd number of pulses were chosen for the driven element, the feedpoint would be off center.)

Thus, the antenna has one source and the feedpoint is at pulse number 14. There are no loads in the antenna (inductors, capacitors, or resistors), so this portion of the MN file may be left out. It would be used if a triband Yagi with traps were being modeled.

Into the computer!

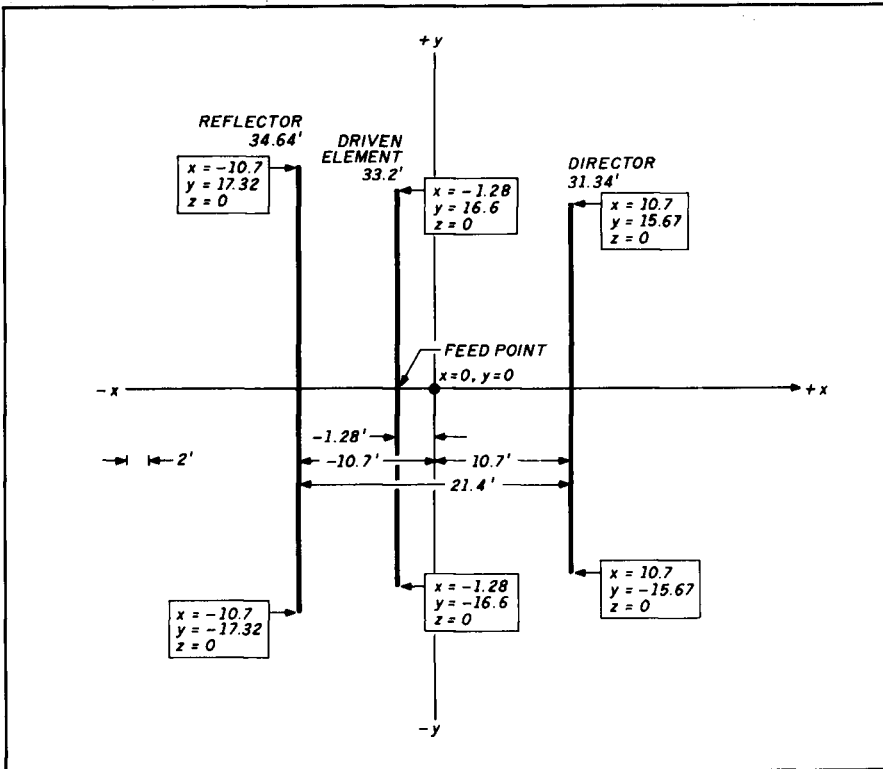
Are you with me? Now that we have all the information in hand, all that remains is to use an ASCII text editor or word processor to place it in a computer file in the stylized form shown in Table 3. Line numbers have been added for reference, but they shouldn't be in the final file.

Line 1 specifies a title for the antenna. This one is the "Three-element 20-Meter Yagi." Line 2 tells whether the antenna is in free space or modeled over the ground. I've chosen free space for this example, so MN looks for those words. If we had specified a Z dimension (like +45) to all elements, it would indicate that the antenna is 45 feet in the air.

Line 3 specifies the analysis frequency. In this case it will be 14.175 MHz, the center of the 20-meter band. Line 4 tells the number of "wires," or elements, and the unit of measurement. We're using three elements measured in feet. This completes the introductory phase.

Now to the element specifics. Three computer lines are required (lines 4, 5, and 6), as there are three elements in the antenna. Each line (starting with the reflector) specifies the number of segments, the X, Y, Z coordinates of the wire (element) tips, and the wire diameter (chosen as 1 inch or 0.083 feet).

FIGURE 3

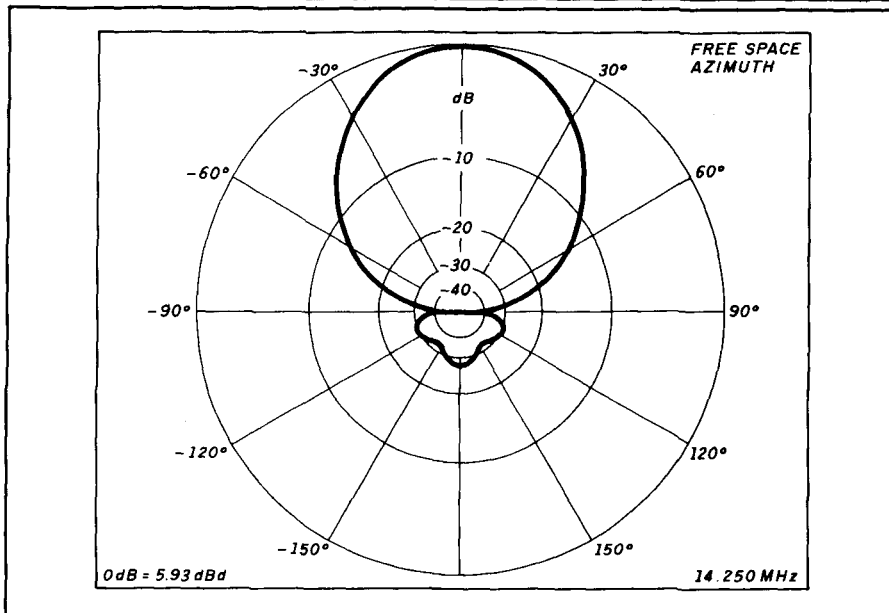


The 20-meter beam laid out in X-Y coordinates.

TABLE 3

Antenna data file for input to MN. Line numbers are for reference only and shouldn't be in the final file.					
1	3	element 20-meter Yagi			
2	Free	space			
3	14.175	MHz			
4	Three	wires, feet			
5	10	-10.7, -17.32, 0	-10.7, 17.32, 0	0.083	
6	10	-1.28, -16.6, 0	-1.28, 16.6, 0	0.083	
7	10	10.7, -15.67, 0	10.7, 15.67, 0	0.083	
8	1	source			
9	14				
10	0	loads			

FIGURE 4



Azimuth plot of four-element Yagi by MN program.

entered into the MN program, which is ready to run. I'll discuss the running of the MN program and show the results of this study in my next column. *Editor's note: There have been instances of (earlier versions) MN failing to load. This is characteristic of any big program which won't fit into available memory. When sufficient memory is provided, MN runs normally. The current version gives the message "out of memory" if it detects too little space.*

Antenna comparison with MN

MN can come up with some surprising answers. It's interesting to observe the power gain and field plot of a popular 20-meter four-element Yagi on a 26-foot boom. DXers favor this antenna. MN shows that the power gain of this array is 5.93 dBd with a front-to-back ratio of 27.17 dB at the design frequency. The field plot is shown in **Figure 4**.

A two-element 20-meter Yagi on an 8-foot boom is shown in **Figure 5** for comparison. The power gain at the design frequency is 5.05 dBd and the front-to-back ratio is 6.37 dB.

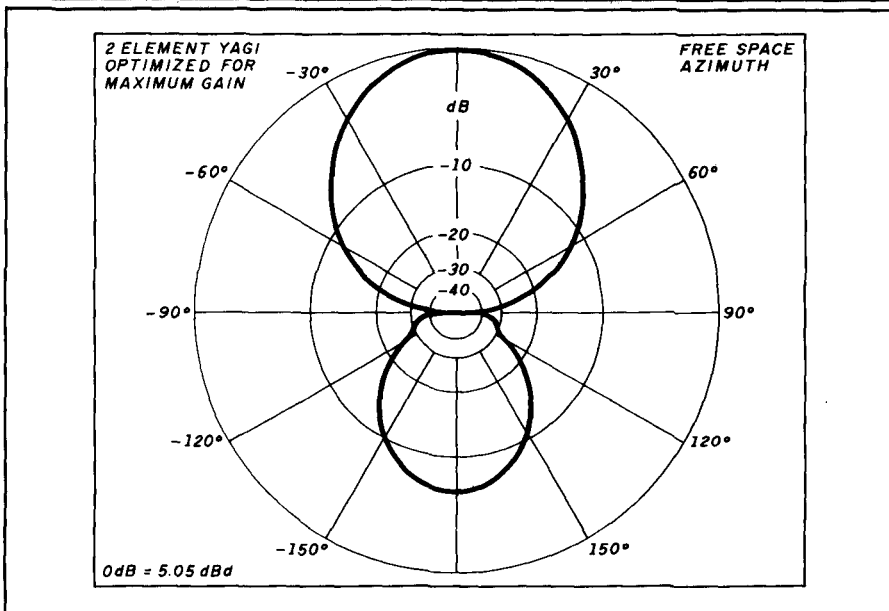
Consider that the boom of the four-element array is 18 feet longer than that of the little two-element beam. What did the extra 18 feet achieve? The big beam has a gain of only 0.88 dB over the small one. The real advantage is in front-to-back ratio — a whopping increase of 20.8 dB!

If you're only interested in power gain, it's a waste of space, time, and money to put up the big array. But if you need front-to-back ratio, there's no doubt as to which antenna is the logical choice.

After running various antenna designs on the MN program, I find it has proven decisively what I had known intuitively for many years — there is no free lunch. Bigger antennas are generally better in most respects than small ones. When you optimize a design for one characteristic (say, maximum gain), you suffer in another area (poorer front-to-back ratio or bandwidth). The "optimum" design depends upon your definition of optimum.

Fortunately, there's a computer program which can help you make some difficult optimization decisions. I'll discuss this interesting program "further down the log."

FIGURE 5



MN plot of two-element Yagi shows good gain but poor front-to-back ratio. In comparing Figures 4 and 5 note that the gains for 0 dB (shown in the lower left) are different.

Line 5 lists the number of segments (10), the coordinates of the left and right-hand tips of the reflector, and the element diameter in feet. Lines 6 and 7 list the same for the driven element and director. Line 8 specifies the number of sources (1) and line 9 gives the drive pulse number (14). Line 9 can also set the applied voltage and current. But these values are optional and,

if not entered, the computer picks a nominal default value to make things work out. Because there are no loads, line 10 is left out.

The MN program allows dimensions in feet, inches, millimeters, etc. and permits wire gauges to be used for element diameter. We needn't concern ourselves with these items now.

The completed antenna file is

The "Dead Band" quiz — The winner is: K2KQU!

"A snowplow begins to clear a roadway at noon on a day during a steady snowfall. The plow moves two miles during the first hour and one mile during the second. What time did the snow begin to fall?"

This brain buster from Andy Loomis, KE0UL, really stirred up a mathematical storm! Andy found the problem in a C.R.E.I. electronics course text illustrating the use of differential equations. Runner-up Bob Phelps, NB7G, found the problem in *Differential Equations* by Agnew, 2nd edition, McGraw-Hill Book Company, New York. As Bob points out, the solution doesn't require the use of differential equations, but it does require calculus. Bob gets honorable mention for his concise explanation of the mathematics of snow plows and his insight into snowstorms.

Those snowplow drivers who realized that the speed of the plow at any instant is inversely proportional to the snow depth at that instant were on the right track!

Chief snowplow driver: Al Cohen, K2KQU

The snowfall began at 11:22:55 a.m., or 0.618034 hours before noon. Al notes that of all the quantities involved, only one (the snow depth) is a linear function. He says if the information given is placed in graph form, it's easy to jump to the conclusion that a graph of these points would place the starting time of the snowfall at 11:00 a.m., provided the plow moves 2 miles the first hour and 1 mile the second. Not so! The real graph of progress is a curve, not a linear function. Further, the speed of the plow is inversely proportional to the depth of snow, and the plow moves equal volumes of snow in equal periods of time.

Calling the speed V , the situation is expressed in the following form:

$$V = \frac{K}{C(t + H)} \quad (1)$$

where t is the hours after noon and H is the hours before noon when the snowfall began. C is the rate of snowfall (a constant). K is a second constant

which, hopefully, will disappear during the calculations.

Speed is related to distance traveled by integration over the time period, so Equation 1 can be converted to an equation in "X" by integrating both sides:

$$\int_{t_1}^{t_2} V dt = \int_{t_1}^{t_2} \frac{K}{C(t + H)} dt \quad (2)$$

$$= \frac{K}{C} \int_{t_1}^{t_2} \frac{dt}{t + H}$$

Because distance is the integral of speed, the integral of V from t_1 to t_2 is: $(X_2 - X_1)$.

On the right side, the integral of $dt/(t + H)$ is the natural log of $t + H$. By simplifying:

$$(X_2 - X_1) = \frac{K}{C} \text{Log}_e \frac{(t_2 + H)}{(t_1 + H)} \quad (3)$$

By further manipulation — substituting the given data for t_1 and t_2 , and remembering that two times the log of a number equals the log of the square of the number — K drops out along the way, the logs drop out, and the result is:

$$\frac{1 + H}{H} = \left[\frac{(2 + H)}{(1 + H)} \right]^2 \quad (4)$$

Expanding and clearing terms, the final formula is:

$$H^2 + H - 1 = 0 \quad (5)$$

where H = number of hours before noon.

This can be solved by the quadratic formula to obtain:

$$H = \frac{-1 + \sqrt{5}}{2} \quad (6)$$

which yields $H = 0.618034$ hours before noon, or 11 hours, 22 minutes, and 55.078 seconds.

Al points out that Equation 5 and the value of H look familiar. The number H is one less than its reciprocal, which follows directly from dividing Equation 5 by H :

$$H = \frac{1}{H} - 1 \quad (7)$$

This is the "Golden Ratio" of Greek

architecture — the ratio of length to width that's said to produce a rectangle most pleasing to the eye.

Finally, Al considers the "Fibonacci Series," a series of numbers that begins with 0,1. Each term thereafter is the sum of the two terms before it:

0,1,1,2,3,5,8,13,21,34,55,89.....etc.


As this series gets longer, the ratio of two consecutive terms gets closer and closer to the "Golden Ratio" and to the term H in the snowplow problem. As interesting sidelights, the truncated series of only 12 terms has converged to $55/89 = 0.61798$ and the 15th term of the series, 377, is the impedance of free space.

Al says this mysterious series of numbers shows up repeatedly in nature in such places as the pattern of markings on animals and the radial directions of bud sites on tree twigs! And, in some fashion, it also describes the physics of plows slogging through steady snowfalls!

Congratulations to Al Cohen, K2KQU, winner of the Dead Band quiz on snowplows!

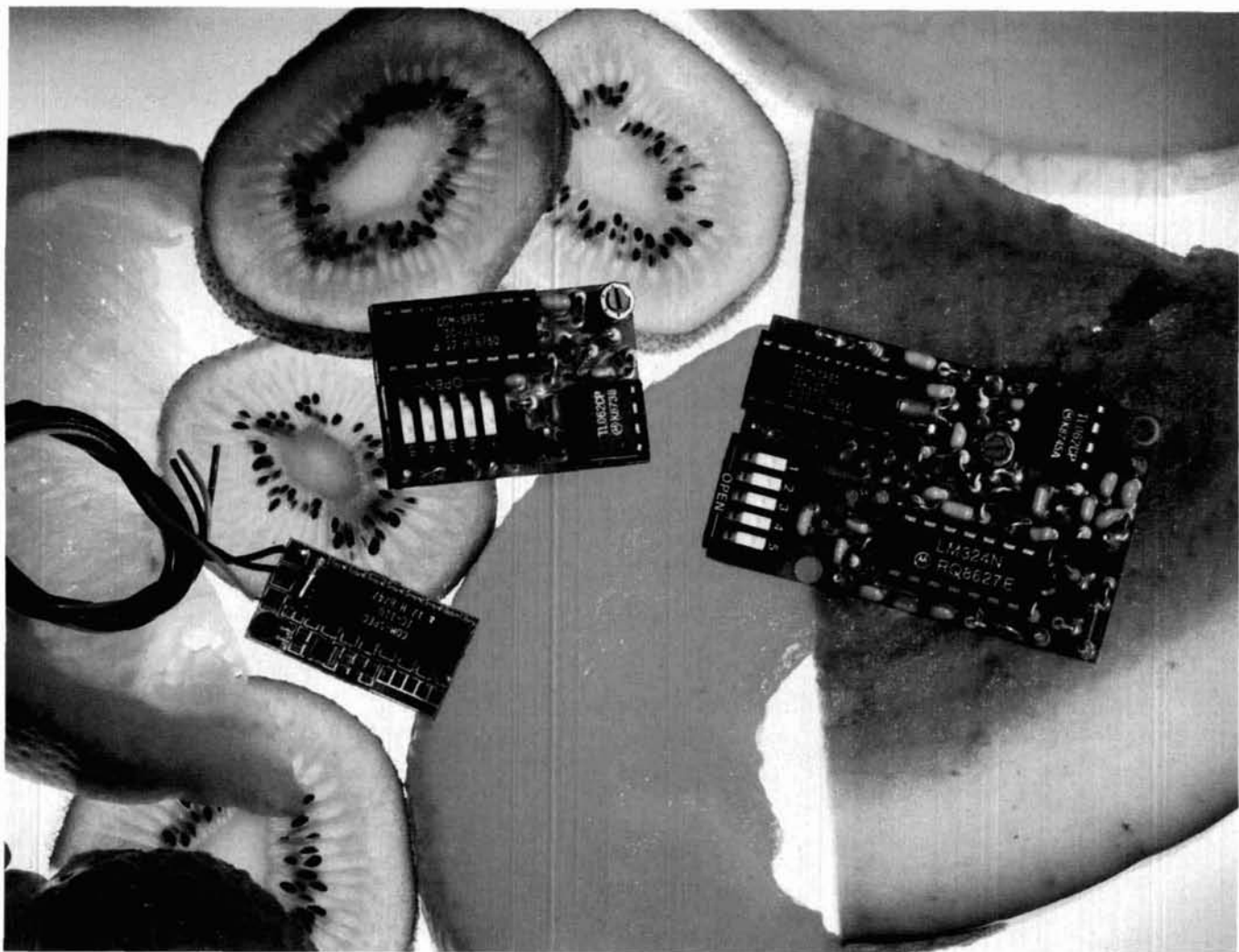
Additional congratulations go to the following people who calculated the correct time: K0NT, K1KG, K6MV, NB7G, K0IXH, WD8KBW, AE2P, W4ZPS, W2IMU, KD9CM, WA2BSR/WB1GQU, KI6WX, K5RA, KB1FZ, NJ2P, KA1RCV, W6EL, G3RUH, W7HGS, and AH6HU.

Others who came close are: K5IU, KZ0Y, N4HUR, N4TXV, K7FC, NV7X, K9MFI, KA7D, W9WSS, KN6W, AI2S/NW2I, N0IBA, K4EF, N4TMI, WE8J, W7XU/0, W6TAD, N0ICS, ND6O, WE0F, W4UGW, W3EBY, NQ0V, K4KC, KA5MXX, K9AY, W1BG, KB0DON, KD8KF, K5ESV, KC8RF, N9DEO, W6EBY, K4ZLE, N2DT, KJ6GR, WA1SOO, WO8E, N6NS, NF7J, WB6JZY, K4WV, WB6LPS, KA8D, K10F, KC6P, KB4LJV, KI7L, W7TLK, KA3ENQ, AA2V, W1BG, AA6CT, and W6JRZ.

Thanks all! I hope you enjoyed this little quiz. 

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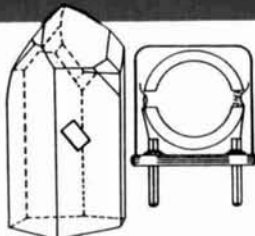
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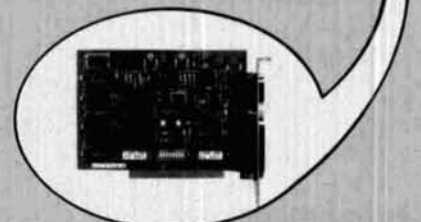
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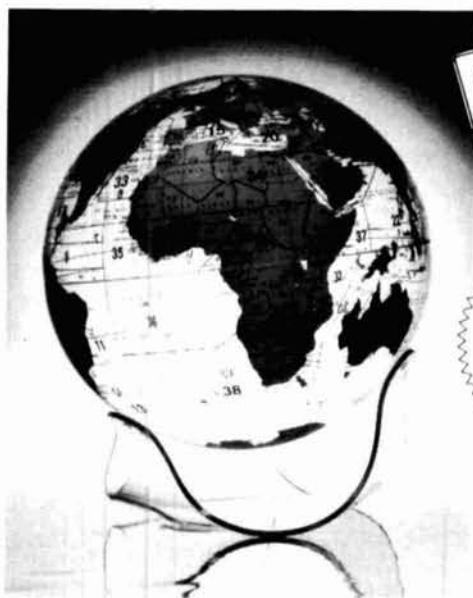
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PERSONAL MESSAGE CENTER

Receive pages from your home station

By Roger Owens, AA4NX, 2042 Old Big Cove Road, Owens Cross Roads, Alabama 35763

My Personal Message Center (PMC) functions as a reverse autopatch, without the phone line, when connected to a radio transceiver.

Transmit page mode

I designed the PMC so I could receive pages from my home station. If another ham operator wants to reach me, he presses the "page" button on the PMC. This initiates a CW message (for example, AA4NX/RMT), which is generated every 10 seconds and stored in EPROM. If I don't respond within ten pages, the PMC terminates the call. If I'm available, I answer the page by entering a two-digit preselected access code. The code terminates the CW message, letting me activate the push-to-talk line to the transceiver. The microprocessor of the PMC does all the housekeeping functions. It doesn't allow PTT from the radio to be active for over 30 seconds at a time, and it identifies the station every 15 seconds of activity time on the transmitter. The call is terminated if the transmitter times out or the control operator enters a # sign.

Once the call termination (#) is received, the PMC identifies the station (AA4NX/SK) in CW, and disables the PTT to the transmitter. Two indicator LEDs operate during the page mode. One is the page LED, which blinks at a 1-second rate during the page mode. This tells you the device is active. The other LED, the PTT indicator, turns on whenever the PMC activates the radio transmitter.

Receive page mode

The PMC can also receive page commands from a mobile station. Enter your access code followed by the command *1. This activates the "callme" LED indicator which blinks at a 1-second rate, and a beep tone which is on for a second and off for a second. The PMC remains in this mode until it receives another *1 command or someone at the home station pushes the page button. Either of these commands terminates the callme LED and the beep tones. If the mode is terminated because the page button is activated, the PMC goes to transmit page mode operation.

DTMF squelch

The PMC acts as a DTMF controlled squelch. A set of relay contacts plugs into the external speaker audio jack, breaking the audio line. The DTMF squelch is activated any time a valid access code is received, and during all operating modes.

Remote site operation

If you need to make field measurements (after installing a new antenna system, for instance), you can access the PMC from your mobile or handheld using remote site operation mode. Activate this mode by entering your access code followed by *0. When the command is accepted, the PMC will:

- Activate the transmitter for 5 seconds of dead air time.
 - Send the CW message AA4NX/RMT.
 - Allow an additional 5 seconds of dead air time.
 - Send the CW message AA4NX/SK.
- After performing these tasks, the PMC returns automatically to its idle state.

User activated ports

The PMC has four DTMF-activated outputs capable of controlling external devices. These outputs are for driving relays, etc., and are active high. To activate these ports enter your access code, followed by *2, *3, *4, or *5. To deactivate, repeat this step.

Radio interface

Bringing the receive audio out from your radio to the PMC requires just one modification. The best place to tap the audio is at the input to the volume control. At this point the audio

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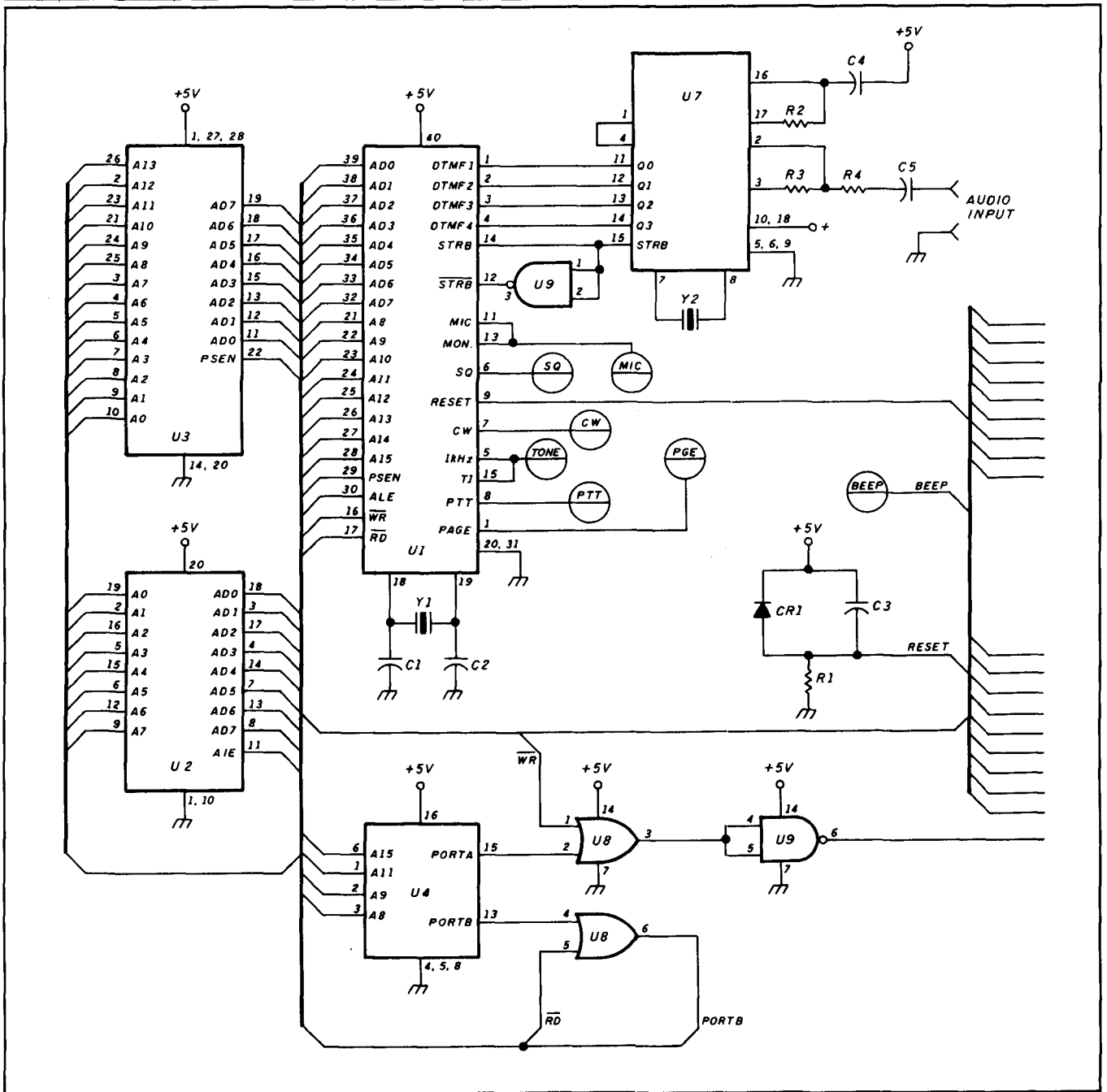


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FIGURE 1A



Schematic of the Personal Message Center.

remains constant, regardless of the volume control setting.

Plug the DTMF squelch into the external speaker jack. Place a normally open switch (SW1) in parallel with the DTMF squelch relay contacts. To monitor the frequency, simply close the switch.

The transmitter audio line is unbroken from the microphone to the radio. Feed the Personal Message Center CW audio parallel with the microphone audio. This is a capacitively coupled output; it should not affect normal operation.

Break the push-to-talk line and feed it to a DPST switch (SW2). Attach one side of the switch to the PMC push-to-talk input. Connect the output relay contacts of the PMC to the push-to-talk input on the radio. This gives push-to-talk control

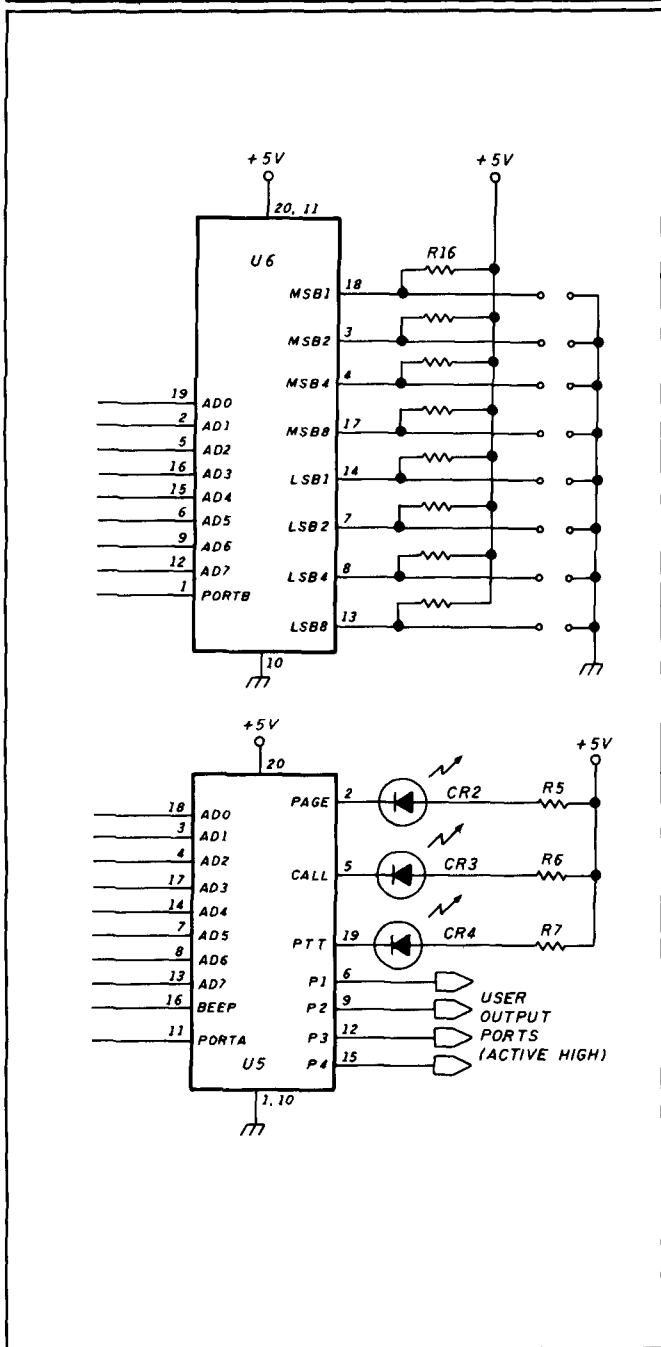
to the remote station operator. Hook the other pole of the switch to the radio push-to-talk line for normal push-to-talk operation.

Circuit description

The Personal Message Center is basically a digital device (schematics shown in Figures 1A and 1B). The only adjustment you'll need to make is to the beep level which connects to the microphone.

The microprocessor is an 80C31(U1) 12-MHz, 8-bit processor. The memory is a 27C64A 8K × 8 EPROM (U3) addressed by a 74HCT373 octal latch (U2). The system is reset on power-up by CR1, C3, and R1. External ports are accessed by the

FIGURE 1A



decoder 74HCT138 (U4) and associated logic gates. The two-digit access code is read through a 74HCT373 (U6). The actual selection is made by shorting or opening the appropriate bit. Each is binary weighted. The first four positions select the most significant digit of the access code; the second digit is selected by the upper four.

Example:

Shorting position	1	2	3	4	5	6	7	8
Open/short	0	0	0	S	0	0	S	S
Binary	1	1	1	0	1	1	0	0
Access code	7				3			

The output control port 74HCT373 (U5) handles the page LED, beep, and user control functions.

The beep output control pin and a NAND gate (U14) drive

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Designator	Part	Description	Quantity
CAPACITORS			
C1,C2		20 pF	2
C3		6.8µF 16 volt	1
C4,C5,C7,C9		0.1µF 50 volt	4
C6		220µF 16 volt	1
C8		1000µF 35 volt	1
INTEGRATED CIRCUITS			
U1	80C31	8-bit microprocessor	1
U2,U5,U6	74HCT373	8-bit latch	3
U3	27C64A	8K × 8 EPROM	1
U4	74HCT138	2 to 8 MUX	1
U7	8870	DTMF decoder	1
U8	74HCT32	Quad or gate	1
U9	74HCT00	Quad NAND gate	1
RESISTORS			
R1	8.2 k	1/8 watt, 1 percent resistor	1
R2	309 k	1/8 watt, 1 percent resistor	1
R3,R4,R14	100 k	1/8 watt, 1 percent resistor	3
R5,R6,R7	560 k	1/8 watt, 1 percent resistor	3
R8,R9	10 k	1/8 watt, 1 percent resistor	2
R11,R12	2 k	1/8 watt, 1 percent resistor	2
R13	100 k	1/8 watt, 1 percent resistor	1
R15	1 k	Pot	1
R16		10 SIP resistor	1
SEMICONDUCTORS			
CR1,CR2,CR3,CR4,CR6,CR7	1N4148	G.P. diode	6
CR5	DF04	Bridge rectifier	1
Q1,Q2	2N3904	NPN transistor	2
Q3	2N3906	PNP transistor	1
Q4	7805	+5 volt regulator	1
RELAY (Miscellaneous)			
RLY 1,2		1A05	2

PNP transistor Q1, which produces the drive for an 8-ohm 2-inch speaker. The 1-kHz tone signal is generated by the microprocessor. This tone is also applied to another NAND gate (U14), activated by the microprocessor CW control bit. The output of this gate supplies CW audio to the transmitter through its associated components.

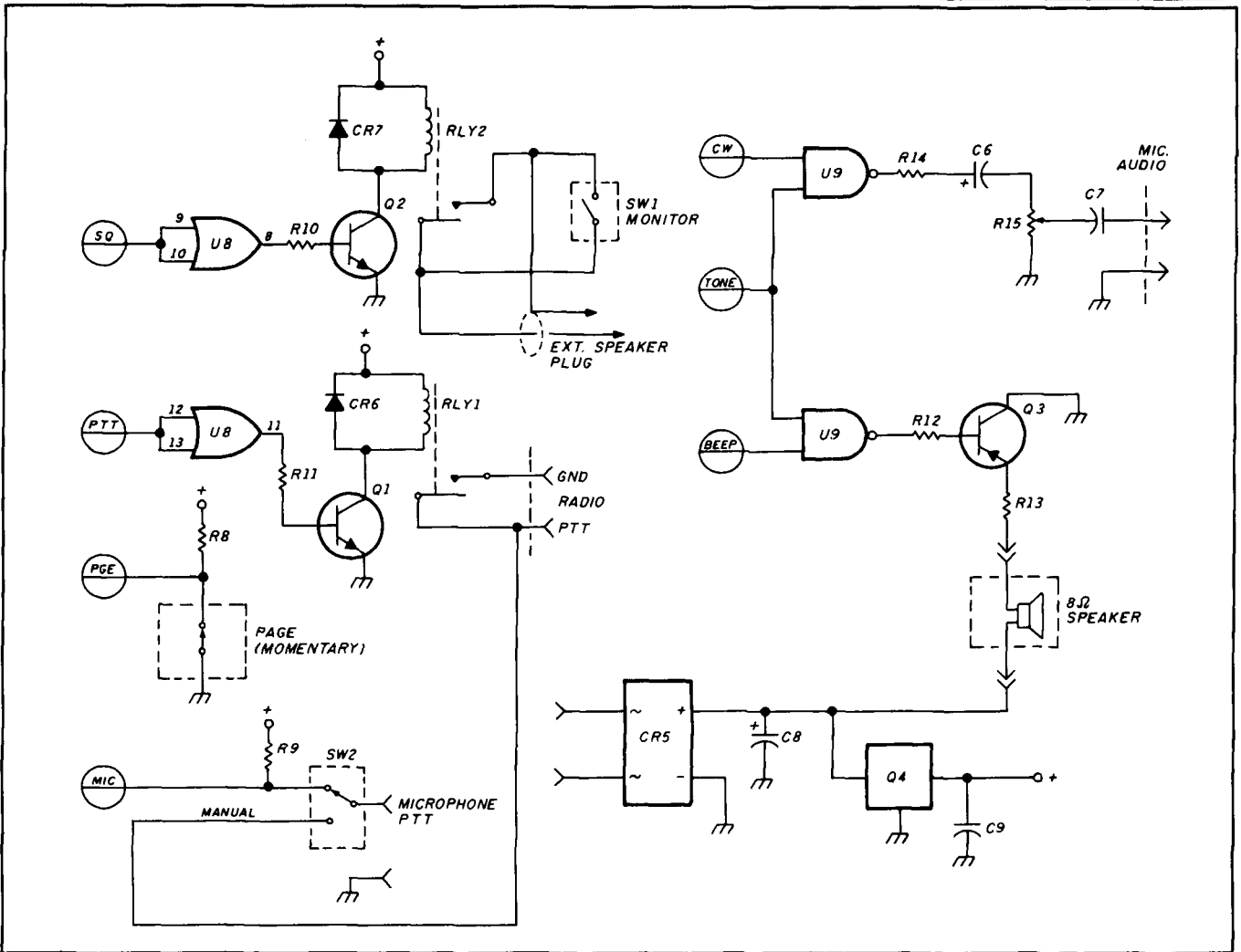
The DTMF receiver is a GTE 8870 device (U7). Receive audio is applied to the chip through C5, R3, and R4. Output detection delay is controlled by C4 and R2. When valid DTMF tone is detected, a strobe is applied to the microprocessor, along with a 4-bit code. The microprocessor decodes this information and performs the task requested.

The microprocessor controls the activation of both the push-to-talk and DTMF control squelch relays. Both circuits are identical and are buffered by OR gates (U8).

The push-to-talk line supplied from the microphone switch, which is active low, is a microprocessor input control line. The page push-button switch which uses a normally closed (momentary) push button is another. This furnishes an active high signal to the microprocessor to initiate the paging sequence.

The power supply consists of a rectifier bridge (CR5), filter capacitor (C8, C9), and +5 volt regulator (Q4). Complete, assembled pc boards are available from Valley Communica-

FIGURE 1B



Additional detail of the schematic of the Personal Message center.

tions, P.O. Box 277, Owens Cross Roads, Alabama 35763. (Write for price information.) PC boards with EPROM cost \$27. Programmed EPROMs are \$8.00. (Please specify call.) Other components may be purchased separately.

Note: Carefully review FCC Rules concerning remote control operation (Part 97.79) and reverse autopatches. This device may not be operated below 220.1 MHz under FCC rules concerning remote control links. Ed.

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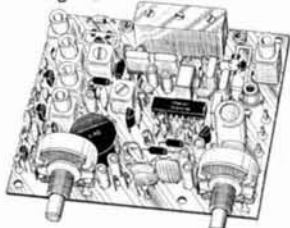
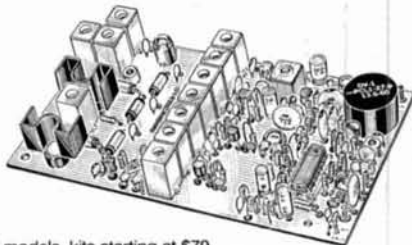
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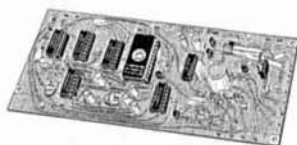
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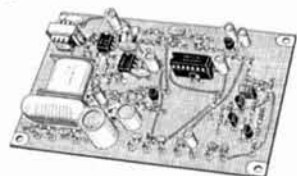


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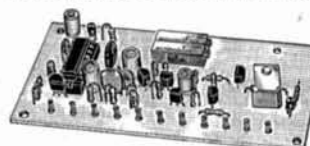
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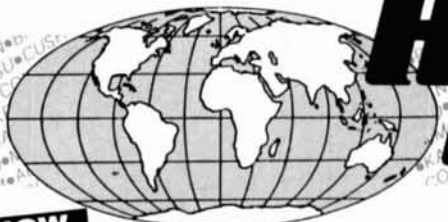
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CA-2X4SR	146 MHz 446 MHz	3.8dB 6.2dB	150W	3' 4"	MOBILE
CX-901	146 MHz 446 MHz 1.2 GHz	3.0dB 6.0dB 8.4dB	150W	3' 6"	BASE/REPEATER
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A DEEP NOTCH RESONANT FILTER

Filter using a 2N2222 transistor provides notch greater than 30 dB

By Douglas A. Kohl, WØTHM, 417 6th Avenue, N.E.,
Osseo, Minnesota 55369

The ideal notch filter passes zero signal voltage over a very narrow bandwidth. The signal at all other frequencies will pass through the filter unaffected. In practice, there will always be some signal voltage that passes through the filter at the notch frequencies.

One of the techniques commonly employed in active notch filters using op amps subtracts a portion of the input signal from the output of a frequency modified stage! When a fraction of the input signal is matched to equal the voltage present at the frequency of the notch, the depth of the overall filter output at the notch frequency will increase and the output may approach zero voltage.

The same principle can be applied to resonant LC filters, which can work at radio frequencies much higher than most op amp notch filters. The simple series-resonant filter shown in **Figure 1A** exhibits a low output voltage, e_o , at the resonant frequency:

$$f_{notch} = \frac{1}{2\pi \sqrt{LC}}$$

At resonance the notch voltage is *not zero*, due to the effective wire resistance (R_w) in the inductor and other coil/capacitor losses. The value e_o results from the voltage divider action of R_1 and R_w (see **Figure 1B**), where e_g is the input signal voltage applied to the filter.

Note that the capacitor and inductor reactances cancel at resonance, leaving only R_w effectively in the circuit. Below the resonant frequency, the capacitive reactance becomes much larger than R_w . The divider fraction in the formula approaches unity, making e_o nearly equal to the input. Similarly, at frequencies above resonance, the inductive reactance increases and e_o increases (see **Figure 1C**).

Q versus notch depth

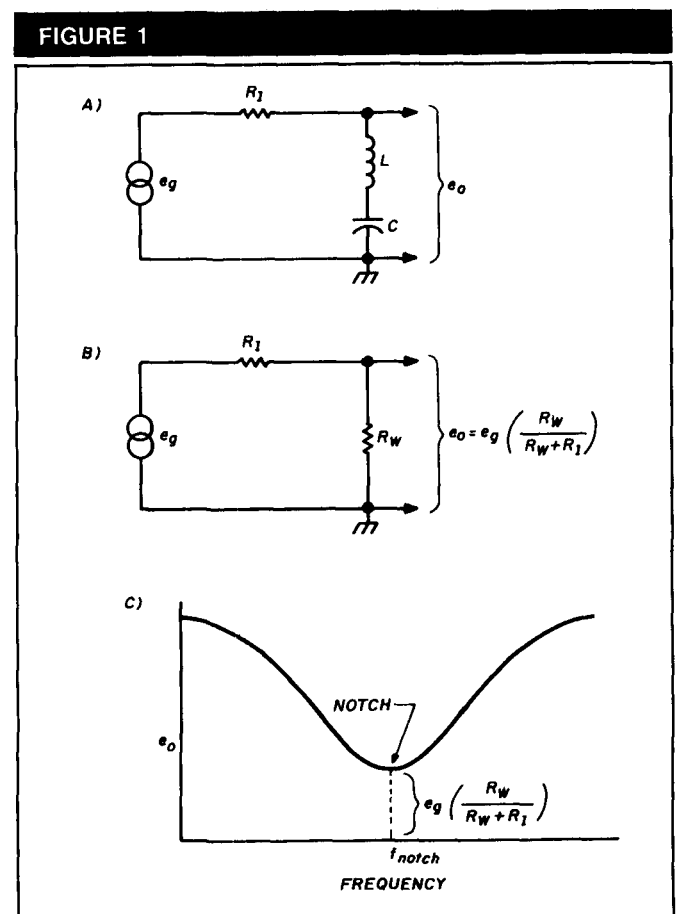
In a series-resonant circuit, the current that flows at resonance must pass through coupling resistor R_1 as well as

the source of voltage, which has some internal resistance (see **Figure 1A**). These additional resistances lower the Q of the circuit accordingly:

$$Q = \frac{\text{inductive reactance}}{\text{all circuit losses}} = \frac{2\pi F_{notch} L}{R_w + R_1 + R_g} \quad (1)$$

where: R_g is the internal resistance of the signal source.

To increase the circuit Q and make the notch narrower in frequency,* an additional resistor (R_2) is added to provide a low resistance path for the circulating current between the capacitor and inductor at resonance. In **Fig-**



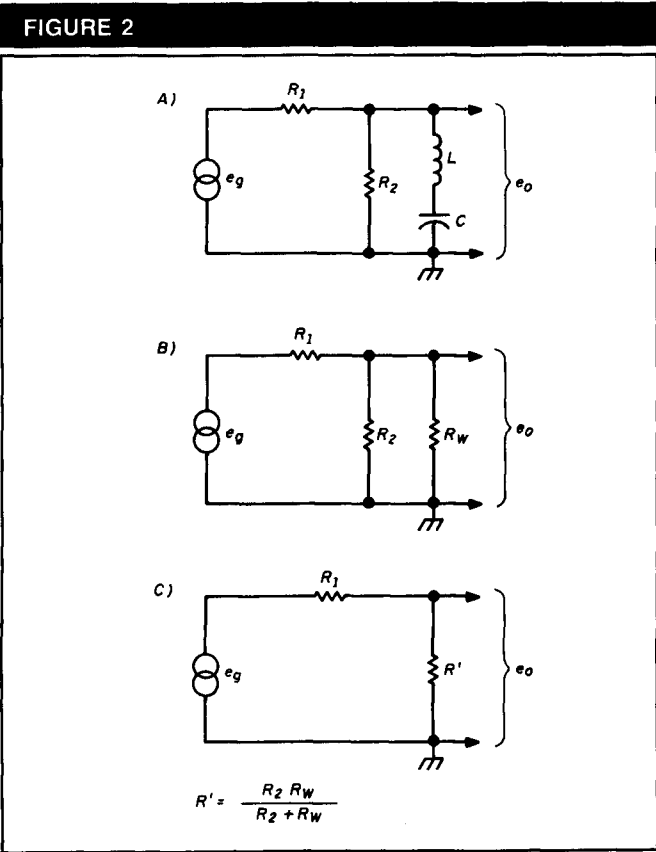
A conventional series-resonant filter and equivalent circuits.

$$* Q = \frac{F_{notch}}{\text{bandwidth}}$$

Figure 2A, the smaller R_2 is made the higher the Q. At resonance, the circuit acts as if coil resistance R_w is in parallel with R_2 (Figure 2B), which can be combined into one equivalent resistance, R' , for analysis (Figure 2C).

To make the filter input resistance high, you'd choose R_1 to be much larger than R_2 . Thus:

$$C_o = e_g \frac{R'}{R_1} \quad (2)$$



High Q series-resonant filter and equivalent circuits at resonance.

This would make e_o in the notch much smaller than e_g .

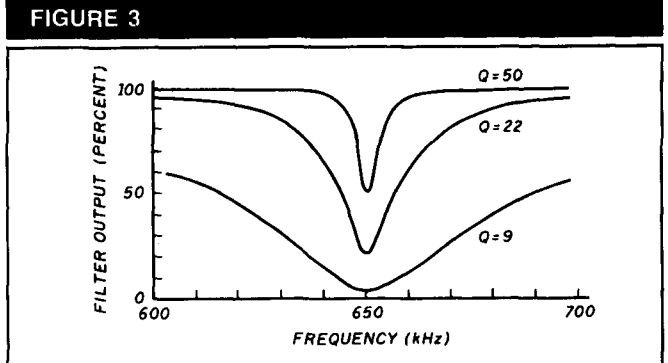
The results of the circuit in Figure 2A are shown in Figure 3. The bandwidth of the notch is the smallest for the highest Q condition ($R_2 = 10$ ohms). However, the output voltage at resonance is relatively high; that is, the notch depth is shallow.

When $R_2 = 100$ ohms, the circuit Q decreases to 9, but the output voltage at the notch is much smaller. The circuit design problem becomes either a narrow bandwidth with a shallow notch, or a wide bandwidth with a deep notch. The circuit shown in Figure 4 solves the problem and has both a narrow bandwidth and a deep notch.

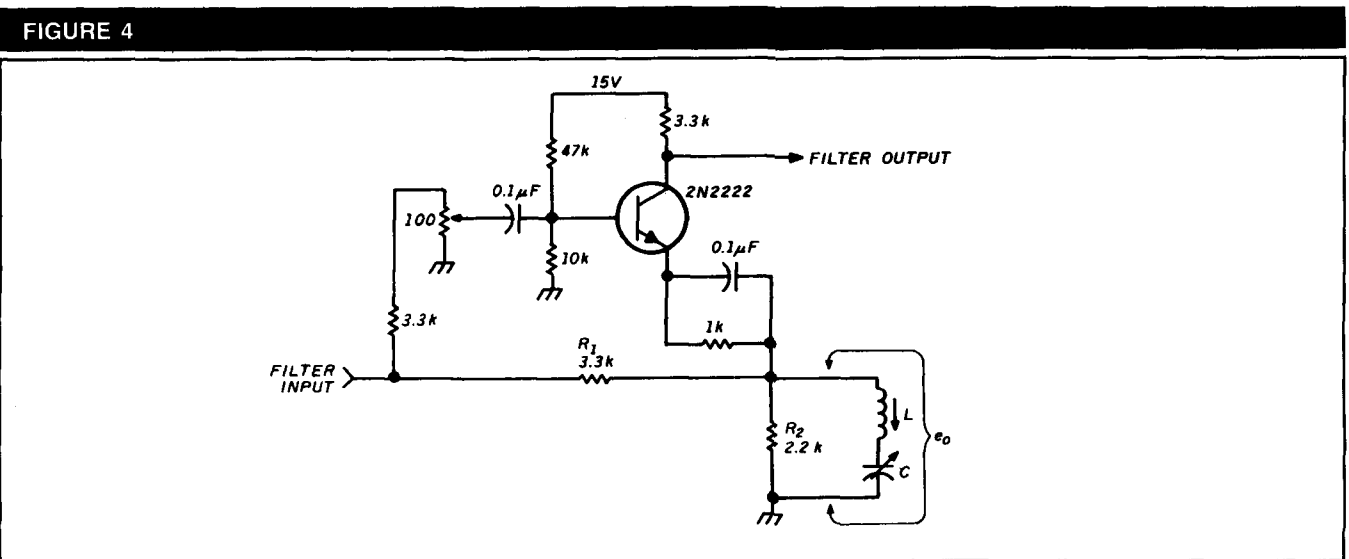
The circuit

Figure 4 contains the series-resonant circuit discussed previously with e_o , the voltage drop across it, identified. When R_2 is small to achieve high Q, e_o at resonance is about 85 percent at frequencies above and below the notch. See the upper curve of the normalized frequency response shown in Figure 5. Even though it has very high Q, the depth of the notch is so shallow that it would be a very poor filter without the rest of the circuit. The phase of the notch voltage at e_o is the same as that of the filter input signal.

A 2N2222 transistor extracts the difference between the input signal and e_o , canceling e_o from the output at the



Frequency response of high Q series-resonant filters.



Schematic diagram of notch filter.

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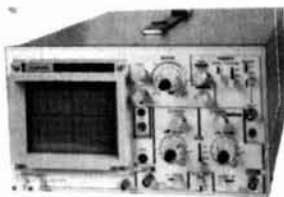
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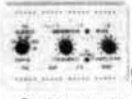
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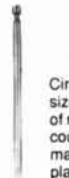
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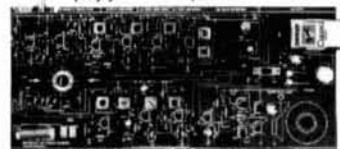
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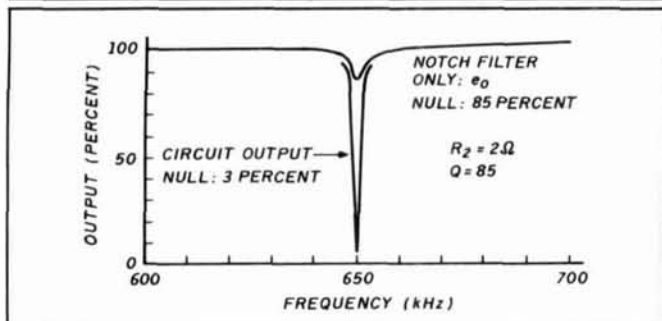
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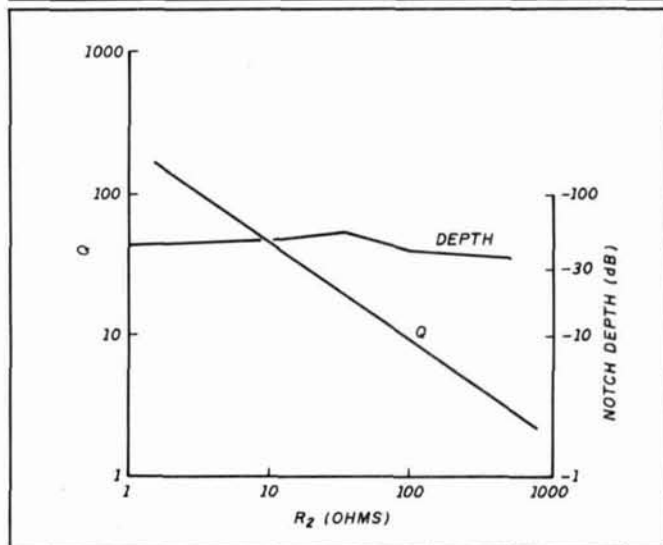
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FIGURE 5



Depth of notch improvement of the circuit.

FIGURE 6



Circuit performance for Q values from 8 to 98.

notch frequency. To do this, use a pot to adjust the amount of the input signal appearing at the base of the transistor to match exactly the e_o voltage present at the emitter at resonance. The filter output voltage at the notch center will typically drop to less than 3 percent. See the lower curve of Figure 5 for particulars.

To change the notch frequency, you may change either L or C. Because the Q varies with frequency, you'll also need to readjust the pot to achieve the deepest notch performance for each different frequency.

Results

The maximum Q attainable depends on the loss characteristics of both the inductor and capacitor you chose. Experimental data presented in Figure 6 show the effect on the circuit Q as R_2 is varied. The coil is an air-core RF inductor.

You can adjust the circuit to provide a deep notch exceeding 30 dB over a wide range of Q values or bandwidths. The voltage divider loss at the series resonant circuit, e_o , was restored at the filter circuit output as a result of transistor gain. *hr*

REFERENCES

1. Frank R. Dungan, *Linear Integrated Circuits for Technicians*, Breton Publishers, North Scituate, Massachusetts, 1984, page 192-3.

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EFFECTIVE NOISE TEMPERATURE

PART 1: INTRODUCTION AND BACKGROUND

Understanding optimum noise performance

By Michael E. Gruchalla, P.E., 4816 Palo Duro NE, Albuquerque, New Mexico 87110

The concept of effective noise temperature is used to specify noise characteristics of low noise amplifiers and other low noise systems. It's generally understood that the lower the value of the effective noise temperature, the "better" the amplifier. Just exactly what is the effective noise temperature and why is it used instead of the more traditional noise figure specification? I'd like to discuss how the effective noise temperature specification is defined, its relation to the noise figure, and some of the pitfalls that you may encounter when trying to apply this specification.

Thermal noise

First, you must look at thermal noise and how it affects an electronic circuit. In 1928, Johnson showed that noise power is available in a system simply by virtue of the system temperature.¹ This is thermal noise, often called Johnson Noise because of his initial investigation of the phenomenon. Johnson showed that the noise power available to a *matched load* from a thermal source is a function of the absolute temperature, the noise bandwidth, and Boltzmann's constant (Equation 1).

$$P_n = k T BW_n \quad (1)$$

where:

- P_n = noise power (W) available to the load
- k = Boltzmann's constant
= 1.38×10^{-23} W-s/°K
- T = Absolute temperature (°K)
- BW_n = Noise bandwidth (Hz)
= $\pi/2 \times$ signal bandwidth (see text)

Only resistances contribute this thermal noise; ideal reactive components don't. However, all real devices exhibit some resistive component which contributes thermal noise power to a circuit.

Spot noise

Often the noise is specified in terms of a 1-Hz noise bandwidth — noise power per Hz, noise voltage, or current per root-Hz. This is called the spot noise and is given by Equation 2, which is simply Equation 1 with a 1-Hz noise bandwidth.

$$P_n(\text{spot}) = k T \quad (2)$$

It's important to remember that the noise performance of many devices — like bipolar transistors, FETs, and amplifier diodes — is often specified in terms of spot noise at various operating frequencies. This information is provided because the noise performance of typical devices isn't the same at all operating frequencies. If you use a spot noise value alone for noise computations, without the noise bandwidth of the application, you'll get a noise level much lower than the level that can actually be obtained.

Uncorrelated noise voltages

Noise is a statistical quantity. As such, noise voltages and currents can't be combined by simple addition. Only the mean square values may be added. The mean square value is simply the square of the rms value. Because noise powers are functions of the mean square of noise voltages or currents (for example, power is equal to the rms voltage squared divided by a resistance), noise powers may be added directly where necessary. However, when unrelated (*uncorrelated*) noise voltages (or currents) are to be added, the combined voltage is the square root of the sum of the mean squares of the individual voltages. For instance, if you have a 1-mV rms noise source in series with a 2-mV rms noise source, the total noise voltage due to the two sources in series is 2.24 mV rms (the square root of 5 mV),* not 3 mV.

Noise bandwidth

Noise bandwidth is the bandwidth used in noise computations. It's shown here as BW_n . The noise bandwidth is related directly to the signal bandwidth and the nature of the transfer function of the system under consideration. (The mathematics involved in the development of the noise bandwidth is beyond my intent here. For those interested

* $\sqrt{5 \text{ mV}^2} = 5 \quad \sqrt{5 \text{ mV}} = 2.24$

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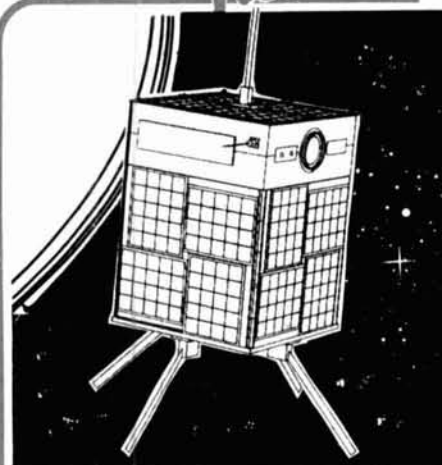
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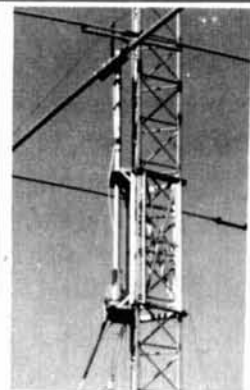
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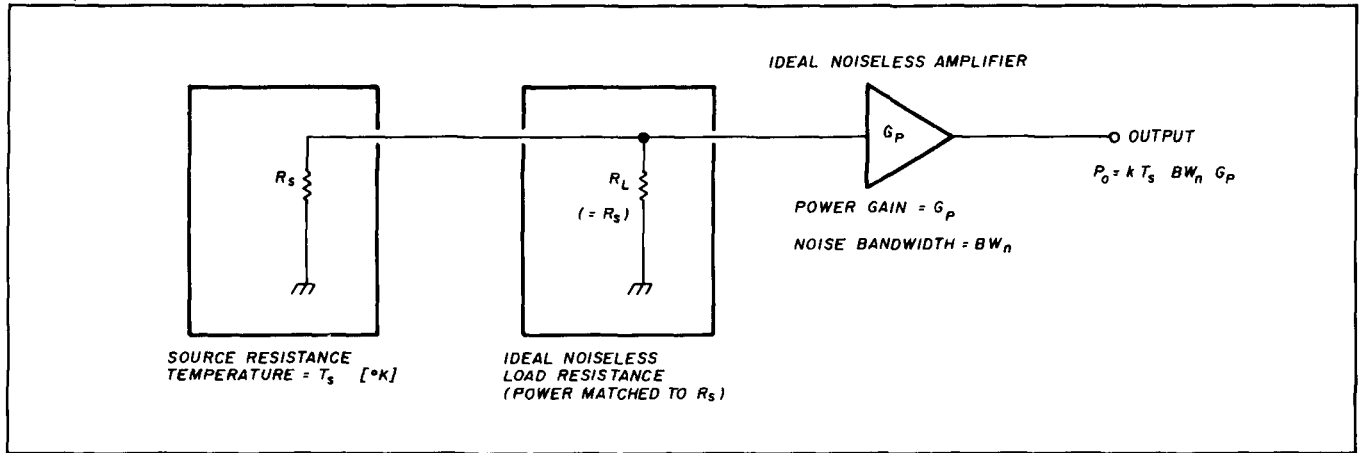
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FIGURE 1



“Warm” resistive source with matched load.

in digging deeper, Reference 1 is a very good introductory text on noise phenomena.) When the amplifier has a relatively flat passband response, and the bandpass cutoff characteristic follows a single time constant characteristic or rolls off at 20 dB/decade, the noise bandwidth is very nearly $\pi/2$ times the signal bandwidth. This approximation is generally sufficient for many noise calculations. In cases of very low noise systems and very accurate noise calculations, the exact noise bandwidth must be measured for the specific system to obtain accurate computations. For this article, I used a single time constant cutoff and the noise bandwidth is defined as $\pi/2 \times$ the signal bandwidth.

Temperature definitions

The temperature in Equation 1 is the absolute temperature in degrees Kelvin. The absolute temperature in $^{\circ}K$ is 273° greater than the temperature in $^{\circ}C$. Absolute zero is $0^{\circ}K$ and $-273^{\circ}C$. The value of $290^{\circ}K$ ($17^{\circ}C$) is commonly used as nominal room temperature for electronic applications.

Thermal noise power is the result of atomic thermal agitation of a system. This is a basic principle of physics and can't be altered by circuit design or other techniques. The only way to reduce the thermal noise power in a specified bandwidth is to reduce the temperature.

The “Ultraviolet Catastrophe”

Equation 1 implies an interesting paradox. If you make the bandwidth infinitely wide, Equation 1 suggests that the available noise power will also be infinite. In early noise research, this led to what was termed the “Ultraviolet Catastrophe.”² This paradox came about because the classic mechanics of the time couldn't explain the noise processes at very high frequencies (ultraviolet wavelengths). The advent of quantum mechanical understanding led to principles which showed that the available noise power per Hz is reduced at very high frequencies. However, Equation 1 is quite adequate for frequencies below about 1000 GHz.

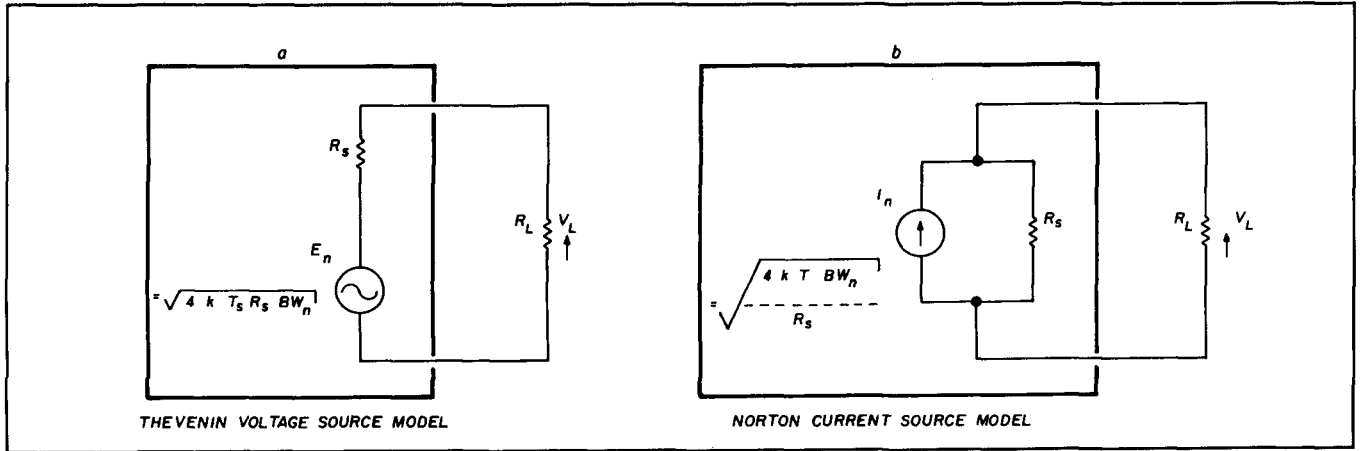
Thermal noise, then, establishes a fundamental lower signal level limit in processing and measurement of signals (amplification, for example). You can define the minimum signal that can be processed as that signal level which results in a signal-to-noise ratio of unity. (Remember, this

is the signal level required to just be able to discern that a signal is present.) Any signal in a system that's smaller than the thermal noise at that point in the system will be lost in the noise. You can examine a simple resistor to find the signal level that results in a signal-to-thermal noise ratio of unity with this resistor as the source resistance. Consider a resistor of value R_s at an absolute temperature T_s . Let the noise bandwidth of the “viewing” instrument of the source (amplifier noise bandwidth or measurement system noise bandwidth) be BW_n . The thermal noise power available from this resistor to a matched load is given by Equation 1. For a simple resistor, the matched load is a resistance of equal value — in this case, R_L . Figure 1 shows a “warm” resistor, your source, loaded into a “noiseless” matched load. Actually, there's no such practical noiseless load. It's used here as a mathematical tool to examine the noise processes.

Signal bandwidth

The noise power available to the matched load from the warm source resistance, given by Equation 1, is also the signal power level at which a power signal-to-thermal noise ratio of unity would be obtained with “perfectly noiseless processing.” Any signal lower than that level will be below the noise and essentially unrecoverable. You can do only two things to reduce the fundamental noise limit in this simple system — reduce the temperature of the source or reduce the noise bandwidth. In practice, the source temperature is usually determined by processes beyond your control. The signal bandwidth, and therefore the noise bandwidth, is generally selected to pass the required spectral contents of the signal being processed. This means that the thermal noise can't be reduced below the fundamental limit given in Equation 1. However, you can draw one important conclusion. In low noise designs, it's important to incorporate a signal bandwidth that's just wide enough to pass the signal of interest with the needed fidelity. If a wider bandwidth is used, no additional signal improvement is gained but additional noise is added. For example, if you want to pass a DC-1 MHz signal, and use a bandwidth of DC-10 MHz, the noise power will be a factor of 10 dB higher than if you used the minimum bandwidth of 1 MHz.

FIGURE 2



Thermal noise source circuit models.

Warm resistor noise model

Johnson's fundamental noise theory lets you find the value of noise power available at the matched load in **Figure 1**. Using this information, you can derive a noise model of the warm resistor. The resistive noise source may be modeled as a voltage source in series with its characteristic resistance (a Thevenin model) or as a current source in parallel with its characteristic resistance (a Norton model). These are exactly equivalent in linear systems, but the voltage source model is generally the model used. These two circuit models are shown in **Figure 2**. Now you know the noise power delivered to the noiseless matched load from **Equation 1**. The voltage that must be impressed across that load to produce that power is easily computed:

$$P_r = E_r^2/R = k T B W_n$$

SO:

$$E_r = \sqrt{P_r R} = \sqrt{k T B W_n R} \quad (rms) \quad (3)$$

A similar computation may be performed for the current in the load.

$$P_r = I_r^2 R = k T B W_n$$

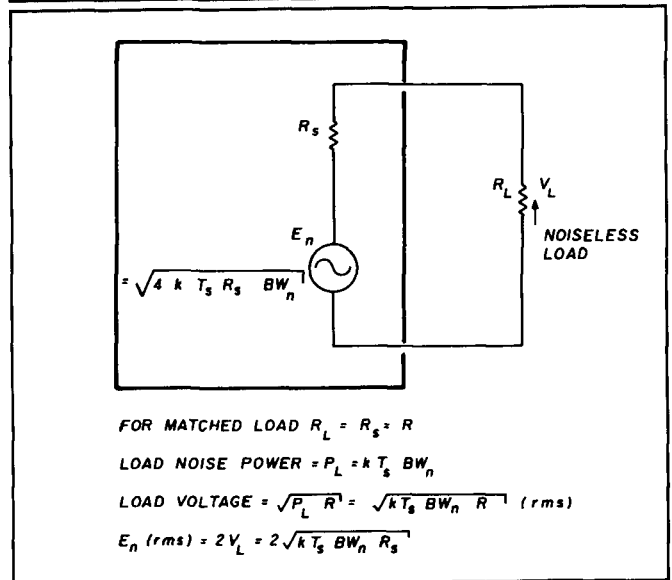
$$I_r = \sqrt{k T B W_n / R} \quad (rms) \quad (4)$$

These results are the noise voltage and current, respectively, delivered to the matched load from the warm source. From these you can compute the equivalent voltage and current sources in **Figure 2**. The computation steps for the Thevenin source model are shown in **Figure 3**. Because the source and load resistances are equal, the equivalent source voltage is simply twice the voltage delivered to the load (one-half of the source voltage is dropped across the source resistance and one-half across the load resistance). The same is true of the equivalent current source in terms of current, so:

$$E_n = 2\sqrt{k T B W_n R} \quad (rms) \quad (5)$$

$$I_n = 2\sqrt{k T B W_n / R} \quad (rms) \quad (6)$$

FIGURE 3



Thermal source Thevenin potential computation.

Matched and unmatched loads

The noise model values were derived from Johnson's findings for matched loads. However, these are true linear models and represent true resistive sources. They may be used accurately in any application independent of the load they are driving. In cases where a matched load is being driven, the maximum power is delivered and is that given by **Equation 1**. For unmatched loads, the delivered power will be less than the matched case and must be computed directly from the specific circuit element values.

Limiting value of source voltage

The noise voltage given by **Equation 5** is the minimum possible noise voltage that can be achieved for a resistor at the specified temperature and noise bandwidth. If this resistance is the source resistance of some device, it is a limiting value of source voltage. Consider that you are attempting to amplify small signals with a 1-MHz signal bandwidth from a 50-ohm room temperature source. Using

Equation 5, the equivalent thermal noise voltage of the resistor may be computed:

$$E_r(50 \text{ ohms}) =$$

$$2 \sqrt{(1.38 \times 10^{-23} \text{ W-s/}^\circ\text{K}) (290^\circ\text{K}) (\pi/2 \text{ MHz}) (50 \text{ ohms})}$$

$$= 1.12 \mu\text{V(rms)}$$

Thus, the signal for which the signal-to-noise ratio at the source is unity in this example is 1.12 $\mu\text{V rms}$. If a unity signal-to-noise ratio is defined as the point at which the signal is just lost in the noise, this thermal noise level is the limiting value of signal level — or about 1 μV in this example. This is the best that can be achieved in minimizing noise. Even with a perfect, noiseless amplifier with a noise figure of 0 dB or noise temperature of zero degrees, this would be the limiting noise. If the signal is below this limiting value, you can't process it out of the noise — even with a perfect, noiseless amplifier.

Part 2 will discuss noise figure phenomena and amplifier equivalent noise temperature. **hr**

1. C.D. Matchenbacher and F.C. Fitchen, *Low-Noise Electronic Design*, Wiley, New York, 1973
2. S.P. Parker, *McGraw-Hill Dictionary of Scientific and Technical Terms*, Third Edition, McGraw-Hill, New York, 1984.

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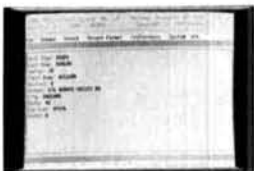
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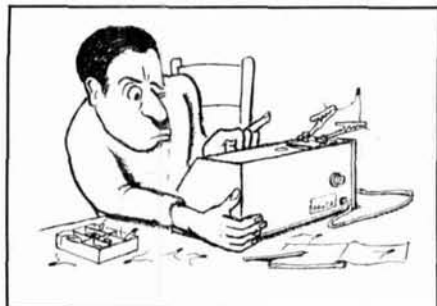
REVIEW OF A LOW COST COMMERCIAL SPECTRUM ANALYZER

One of the most popular topics I've covered in this column concerned a build-it-yourself spectrum analyzer. A lot of readers put together those kits and modified them considerably. I've published some of my modifications, and those of a few readers, in this column over the past two years. A few months ago I told you that a small manufacturer of low cost spectrum analyzers was going to make one of their products available for review in this column. Judging from my mail, that announcement interested a lot of you. Well, Penntek Instruments (14 Peace Drive, Lewistown, Pennsylvania 17044 (717)248-2507) came through with a loaner instrument, so here's the promised review.

What is a spectrum analyzer?

For those who may have missed the earlier articles, I'll review briefly what a spectrum analyzer is and how one can be used in communications work. The spectrum analyzer is a special form of swept superheterodyne receiver with a frequency domain output (*amplitude versus frequency*) instead of the time domain output usually found on oscilloscopes. To display the output of a spectrum analyzer you must use an X-Y oscilloscope, rather than the standard Y-time type. Most common oscilloscopes can be used in X-Y mode if they have a horizontal input, or are dual beam models in which an X-Y mode is provided on the vertical function selector.

An internal sawtooth waveform causes the receiver tuning to sweep from one end of the band to the other. The same sawtooth waveform is also directed to the X-input (horizontal) of the oscilloscope, where it's used to

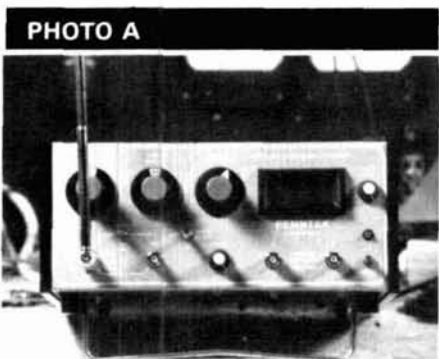


sweep the CRT beam from left to right. As the receiver and scope beam sweep through the range, the detected output is fed to the Y-input (vertical) of the oscilloscope. This causes an amplitude-versus-time display to appear.

Penntek Instruments' Model SA-500E spectrum analyzer (**Photo A**) converts any oscilloscope that has either an X-Y mode or allows access to the horizontal amplifiers (which is the method used on older oscilloscopes). The display outputs from the SA-500E are scaled to 1 volt/division; the X-output is ± 5 volts and the Y-output is ± 4 volts. The horizontal sweep speed (X-output) is variable from 1 to 40 Hz.

Penntek Model SA-500E spectrum analyzer

The SA-500E has a frequency range of 0 to 550 MHz, with a digital dial accuracy of 2 percent of full scale. This instrument is basically a swept triple conversion superheterodyne receiver. There are twelve switch-selected sweep (or "span") widths from 50 MHz/division down to 20 kHz/division in the



Penntek SA-500E spectrum analyzer.

popular 5:2:1 sequence ratio. When the SA-500E is set for the 50-MHz/division span width, the entire spectrum from 0 to 550 MHz is displayed on the oscilloscope. There's also a 0-MHz/division sweep width, in which the SA-500E operates as a manually tuned receiver or wave analyzer. The frequency resolution or IF bandwidth has two options — wide (200 kHz) and narrow (10 kHz). Bandwidth selection is coupled to the sweep width control. The input sensitivity is better than $2 \mu\text{V}$ and the receiver has a 70-dB dynamic range. An RF input attenuator allows selection of up to 70-dB attenuation in 10-dB steps, while an IF attenuator offers a variable range of 0 to 22 dB. Allowable input levels are 0 to +13 dBm RF, 1 volt AC, and 0 volts DC. There are three IF frequencies used in the SA-500E: 700 MHz, 62 MHz, and 10.7 MHz. The SA-500E offers two crystal marker frequencies to help calibrate the dial, 5 and 50 MHz.

The Penntek SA-500E is small and doesn't take up a lot of space on the bench. Its dimensions are 10-3/8 x 11-3/4 x 5 inches and it weighs under 3 pounds. The cabinet is gray and black.

Using the SA-500E spectrum analyzer

The SA-500E is intended for operation with an external oscilloscope. The X and Y outputs of the spectrum analyzer must be connected to the X (horizontal) and Y (vertical) inputs of the oscilloscope. The scope input attenuators must be set to 1 volt/division or 2 volts/division, whichever gives the best display. You then turn the center frequency control to the center of the band of interest and set the span control to the desired sweep width.

Photos B and C show the SA-500E display that appeared when I connected my Measurements Model 80 signal generator to the input of the spectrum analyzer. The signal generator's output frequency was approximately 100 MHz, plus or minus the

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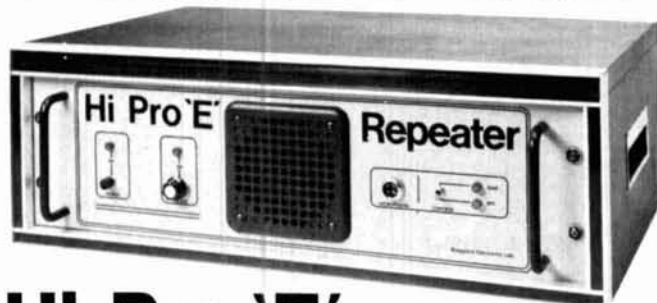
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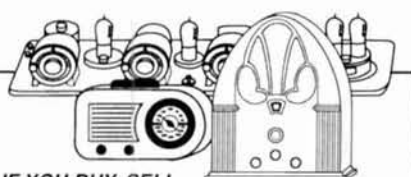
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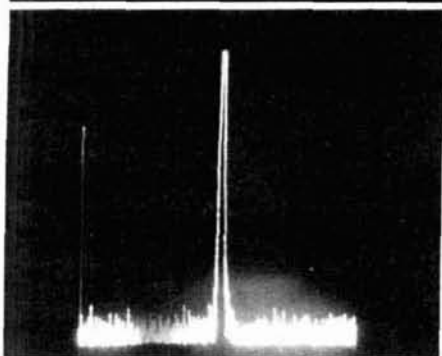
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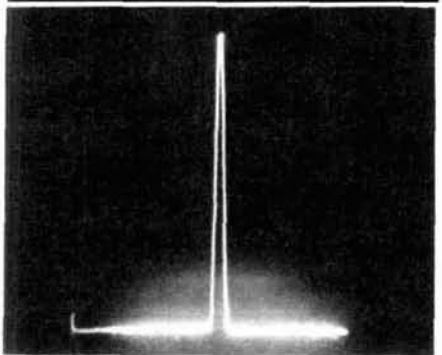
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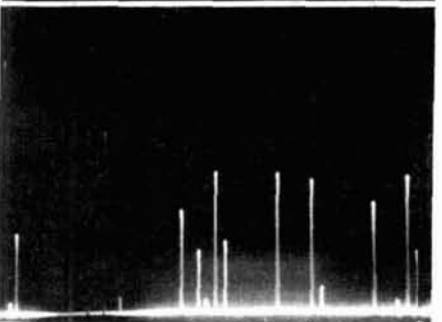
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PHOTO B

Signal generator output signal without video filter.

PHOTO C

Signal generator output signal with video filter.

PHOTO D

Display of 2-meter band signals near Washington, D.C.

30-year old calibration. The signal in Photo B shows the display that resulted when I selected no filtering on the SA-500E. When I selected the video filter, the baseline noise disap-

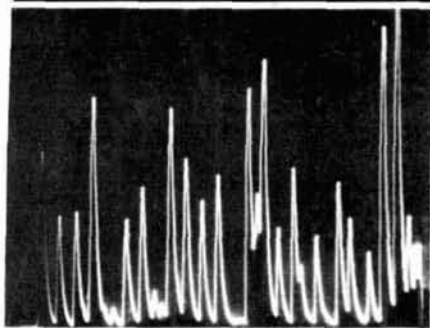
peared and the trace in Photo C resulted. This display was taken with the span control set to 0.05 MHz/division, so the trace represents a 5-MHz slice of spectrum centered on 100 MHz.

After I played with the process of displaying a signal from my signal generator on the scope, I connected the RF input of the SA-500E to the 2-meter ground plane antenna that serves as my access to the local repeater world. Photo D shows a 1-MHz/division slice of spectrum centered on 146 MHz. A large number of local repeaters, mobiles, and fixed stations showed up on the display. The spikes representing these signals appeared and disappeared as the stations went on and off the air.

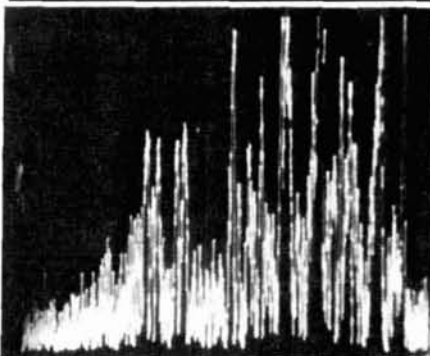
Monitoring the 2-meter band on my old Kenwood transceiver revealed that the centermost spike was from the 146.91-MHz repeater in Fairfax, Virginia. That spike disappeared every time the repeater went off the air. Interestingly enough, most of the time I could also see a much smaller spike ride up and down with the repeater spike; I believe this represented mobile rigs accessing the repeater. Because the repeater receiver is a lot more sensitive than the spectrum analyzer and installed at an elevated location, it could hear more mobile signals than the spectrum analyzer. This accounts for the fact that the mobiles didn't always show up on the scope screen. Only those mobiles that were relatively local would show up.

I tuned the SA-500E center frequency control to the middle of the FM broadcast band (88 to 108 MHz) and looked at the scope display. Photo E shows the FM band viewed from my location in Virginia, near Washington, DC. The large spike on the right is from my Model 80 signal generator — which I'd failed to turn off before retuning the spectrum analyzer.

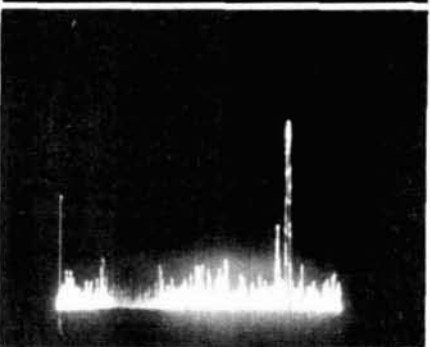
No one who looks at Photo E can doubt that the FM broadcast band is crowded in my area, but wait until you see the HF spectrum. No denizen of 20 meters will deny that HF is overrun with signals; Photo F demonstrates that lamentable fact rather conclusively. This trace represents the HF spectrum centered on 14 MHz, with a sweep span that lets you look about 5 MHz above and below that frequency. The big center group of spikes represents the 20-meter band, while those to the right seem to show a fairly active inter-

PHOTO E

Display of the FM broadcast band signals (88 to 108 MHz) near Washington, D.C.

PHOTO F

High frequency (HF) bands on the spectrum analyzer shows why 20-meters is so maddening.

PHOTO G

"Spur" turned out to be the local oscillator of an FM broadcast receiver sitting next to the spectrum analyzer.

national broadcasting band. I'm sure that the WWV time and frequency standard station is somewhere in that mess!

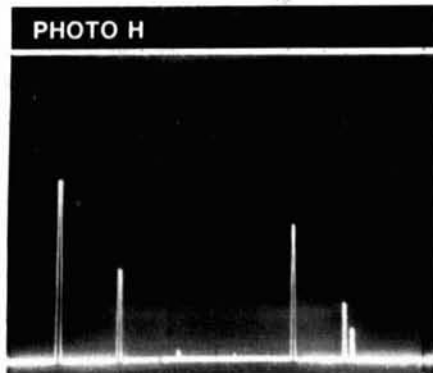
Photo G perplexed me a little bit at first. Initially I thought the signal was a

spurious reponse of the SA-500E (most spectrum analyzers have at least a few). But the signal didn't seem to behave like a spur. Then I found the problem. I listen to radio while working at the word processor and at my workbench. Those who know me personally can attest to my love of (blush) bluegrass music — the more traditional the better. I had the radio tuned to 88.5 MHz (WAMU in Washington, DC) for the Saturday bluegrass show. The SA-500E antenna was disconnected and I was using only a short, shielded scope probe to supply input signal from my signal generator output. But even when the signal generator was turned off, this little signal persisted.

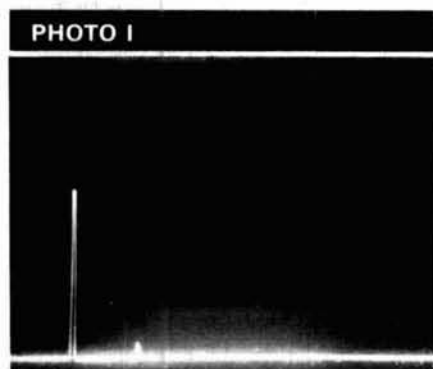
On a hunch, gleaned from an awful troubleshooting experience many years ago, I turned off the FM broadcast receiver on the workbench. I'm sure you can guess what happened. The "spur" disappeared. The signal displayed in **Photo G** is the radiated local oscillator signal from my FM receiver. It was on a frequency of $88.5 + 10.7$ MHz, or 99.2 MHz. The rest of the FM band didn't show up because of the short "antenna" represented by the partially shielded scope probe.

As I mentioned in November's column, I saw this same problem more than ten years ago when I worked at a major university medical center. Nurses in the post-coronary care unit complained that one patient's electrocardiograph (EKG) signal was showing up on two channels of the radio telemetry display. The problem turned out to be an unshielded FM receiver located too close to the telemetry receiver's local antenna. The receiver was "intermodding" (for lack of a better term) with the other signals in the system, forcing the weak EKG radio signal onto a different frequency. Penntek tells me that a lot of hospital biomedical equipment shops have bought their spectrum analyzer. I wish I'd had one that night a decade ago when I was trying to figure out the solution to that perplexing problem. Now it's easy to understand why airlines don't like FM radios on board. The local oscillator of FM receivers falls into the aviation band when the FM radio is tuned to the high end of the FM broadcast band!

While it's fun to look at the spectrum through an instrument like the Penntek SA-500E, there must be a practical use for it before most of us will invest in one. There are several uses of interest to the



Output spectrum of a frequency tripler circuit when output had previously been peaked.



Output spectrum of a frequency tripler circuit after tuning with aid of spectrum analyzer.

Amateur. In addition to identifying intermod problems, it's also possible to use the SA-500E to tune transmitters and other RF devices. Amateurs typically tune RF circuits for maximum output or to null some interfering signal. The SA-500E offers a means for doing this job easily.

Photos H and I show the responses of a VHF oscillator that's part of a converter project I'm building (see **Figure 1**). The local oscillator (LO) frequency is generated in a crystal oscillator at 43 MHz and then tripled to 129 MHz. The tripler is a diode multiplier circuit

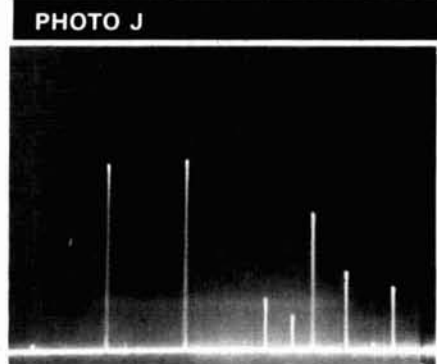
followed by a tuned RF amplifier at the 129-MHz output frequency (a circuit derived from, but not exactly like, an old *ARRL Handbook* project).

There are really two issues involved when you're tuning the desired output of the frequency multiplier. One concern is to maximize the output signal so it can be used to drive the mixer. The other is to minimize the harmonics and other spurious responses. These signals can interfere with the proper receiver operation.

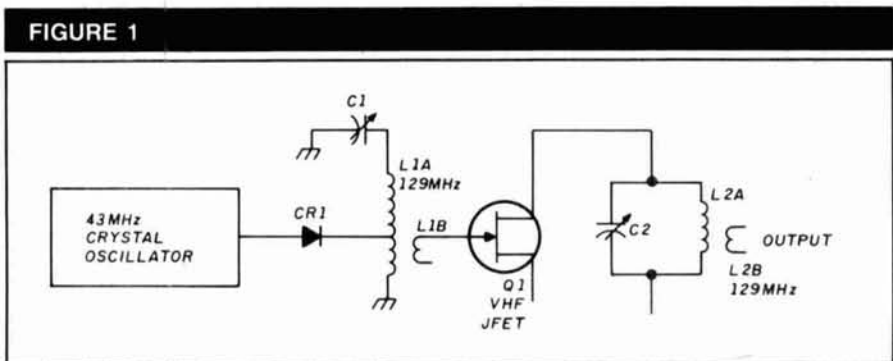
In **Photo H** the leftmost spike is the 129-MHz output signal, while its second harmonic (258 MHz) is the third spike from the left (second highest). There are three other spikes in the spectrum. I'm not sure where they came from, but they weren't supposed to be there. The spacing on the scope display suggests that they might be other harmonics of 43 MHz, or mixing products of the 43 MHz plus its other harmonics.

Photo I shows what results when you tune the multiplier (C1 in **Figure 1**) and the output of the amplifier (C2) properly. The main signal at 129 MHz lost a little amplitude, but the harmonics and/or spurs disappeared. This is what it's supposed to look like.

Photo J shows another application



Bandsweeper mixer output signal when one oscillator was acting "funny."



VHF tripler used to generate the signals in Photo G.



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RS-7A	5	7	3 3/4 \times 6 1/2 \times 9	9
RS-7B	5	7	4 \times 7 1/2 \times 10 1/4	10
RS-10A	7.5	10	4 \times 7 1/2 \times 10 1/4	11
RS-12A	9	12	4 1/2 \times 8 \times 9	13
RS-12B	9	12	4 \times 7 1/2 \times 10 1/4	13
RS-20A	16	20	5 \times 9 \times 10 1/2	18
RS-35A	25	35	5 \times 11 \times 11	27
RS-50A	37	50	6 \times 13 3/4 \times 11	46

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VS-35M	25	15	7	5 \times 11 \times 11	29
VS-50M	37	22	10	6 \times 13 3/4 \times 11	46
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RS-10S	7.5	10	4 \times 7 1/2 \times 10 1/4	12
RS-12S	9	12	4 1/2 \times 8 \times 9	13
RS-20S	16	20	5 \times 9 \times 10 1/2	18

of the spectrum analyzer in adjusting Amateur projects. I'm building a bandsweeper — a sweep signal generator that will cover an entire HF Amateur band at one setting. The project uses two low band VHF voltage-controlled oscillators (VCO) heterodyned together in a Mini-Circuits balanced mixer. One will be fixed at a center frequency and then swept with a sawtooth (the digital sawtooth generator reported in this column earlier); the other will be fixed tuned to a frequency that heterodynes the sweep oscillator to the desired Amateur band. Both oscillators are adaptations of an oscillator circuit that was used in some commercial test equipment.

When I looked at the 15-meter band amplitude versus time display on my oscilloscope, there was a lot of disconcerting distortion of the signal that I believed should be nearly sinusoidal. Looking at the 15-meter output of the project on the SA-500E (Photo J) showed the cause of the distortion. There were a large number of mixer products and harmonics present.

I discovered two problems that needed attention. First, the oscillator feedback control (C1 in Figure 2) was improperly adjusted. Second, I needed either a low pass or a notch filter on the mixer output. Photo K

shows what happened when I adjusted the feedback control to a point between its initial adjustment and the point where oscillation ceased. (If anyone can propose a reason for this spectrum, I'd appreciate it. Just write to me at the address at the end of the column.) The trace in Photo K shows that the mixer product is still in the output, so I'll add a filter before I publish the circuit in this column.

As you can see, there are several possible Amateur applications for the spectrum analyzer. You can use it to adjust the tuning of projects in progress. You can also use it to test transmitters for harmonic radiation. Federal Com-


munications Commission rules, common sense, and plain old decency require that we suppress harmonic emissions as much as possible. Whether the equipment is homebrew or store bought, it's the transmitter's licensee who's responsible for the correct operation of the unit. The spectrum analyzer helps because it lets you test transmitters for spurious emissions.

Finally, you can use the spectrum analyzer to locate sources of RF that are interfering with your own operations. It may also be helpful in locating TVI sources. It's quite possible for Amateurs to be blamed for TVI that isn't their fault — neighbors can be like that. (In my Novice days, an irate neighbor blamed me for TVI even when I was asleep, and before my first Heathkit transmitter arrived in the mail! I guess it was those 500-THz Martians who were using my receiver dipole. Seriously, though, a spectrum analyzer helps make such hidden signal hunts a lot easier.) I'm sure many other applications will surface once a few Amateurs start using this instrument.

Conclusion

The Penntek SA-500E spectrum analyzer is a viable and useful tool for Amateurs. It costs about the same as a decent HF SSB rig (\$1,495), so it may be a little pricey for many people. But if you're into serious technical work, or if your club would consider buying an instrument for everyone to use, you should consider this little analyzer. Penntek will be happy to send you a spec sheet and answer any reasonable technical questions.

NOTE: Several readers have written to me, or approached me at hamfests, to suggest a panadapter project. A panadapter is like a spectrum analyzer, but tends to be fixed tuned to the IF frequency of a ham receiver and sweeps sufficiently to cover the entire band to which the receiver is tuned. Some receiver and transceiver manufacturers make these units as add-on adapters. Does anyone have any ideas for an Amateur panadapter centered on either 8.83 or 9 MHz? If so, then drop me a line, or better yet, write an article yourself and submit it to *Ham Radio!*

Joe Carr, K4IPV, can be reached at POB 1099, Falls Church, Virginia 22041. He would appreciate your questions and recommendations for this column. 

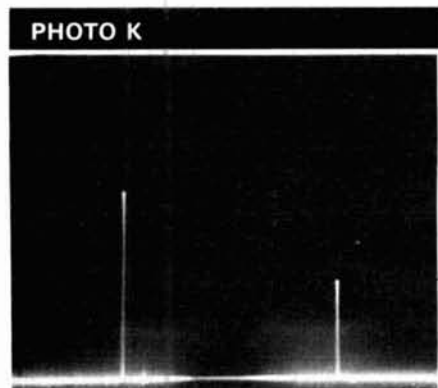
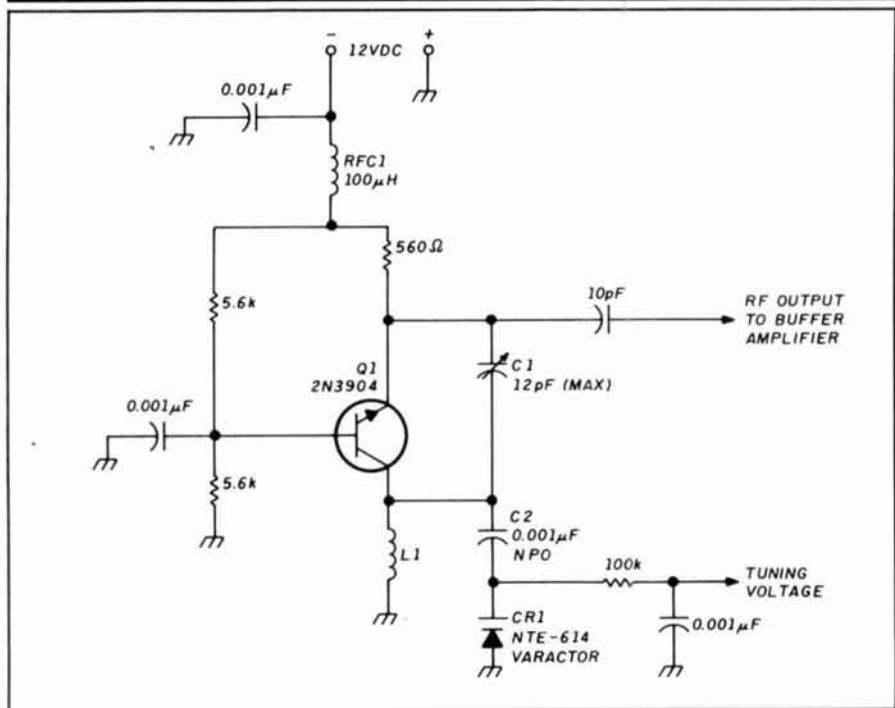


PHOTO K
Bandsweeper mixer output signal when the feedback level was readjusted.

FIGURE 2



Oscillator used in the bandsweeper project.

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Pomona Electronics offers all six of its current family of Quad Clip test adapters in kit form, packaged in a convenient carrying case. Units are designed to fit PCC or PLCC ICs with "J" leads. The kit, Model 5515, includes one each of Pomona's 20, 28, 44, 52, 68, and 84-pin Quad Clip test adapters.



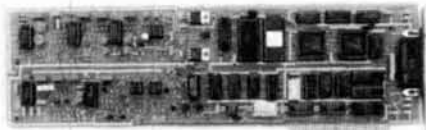
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QRZ Industries of Piedmont, South Carolina announces the uVB-1 natural voice digitizer. The uVB-1 is a single message voice digitizer which can store up to 8 seconds of voice (expandable to 32 seconds with built-in memory expansion port). The unit has two auto repeat modes. One has rapid adjustable repeat for contesting (0 to 20 seconds); the other has a programmable long repeat mode for repeater IDs (up to 20 minutes). A built-in monitor circuit, capable of driving headphones or a speaker, combines digitized and live audio with an auxiliary audio input. There are separate adjustments for balancing live voice and digitized audio, and setting auxiliary audio gain, 600-ohm output level, and monitor output level. To program your voice into the uVB-1 simply speak into the microphone while the unit is in record mode.

The uVB-1X is similar to the uVB-1, but it uses CMOS static RAM for extremely low power consumption and provides on-board memory lithium battery backup. A special power-down circuit ensures that memory contents are maintained when power is lost.

Model VB-8A

The VB-8A is the next generation replacement for the popular VB-8. Model VB-8A has the basic features of the VB-8, along with a number of improvements and enhancements. Featuring eight soft-sectored messages, the VB-8A is capable of recording up to 100 seconds of digitized speech. The audio quality has been improved by increasing the digitizing rate to 40 kHz and adding 14 poles of audio filtering. A built-in monitor amplifier is standard and an automatic serial number inserter (which uses the operator's own voice) has been added.

JUNKBOX VARIABLE CRYSTAL OSCILLATOR

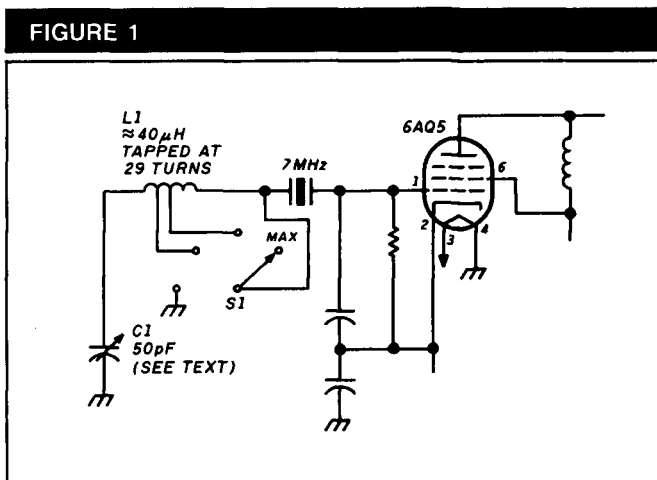
By Charlie Tiemeyer, W3RMD, RD 2, Box 427-C, Chestertown, Maryland 21620

I built an 80/40 Meter Junkbox rig and needed a compatible crystal oscillator. I've come up with one that lets you "rubberize," or move your crystal frequency several kilohertz above and below the fundamental. Most of my "rocks" are the FT-243 type, but I'm sure any type you may have can also be shifted.

The rig mentioned above uses a 6AQ5 oscillator tube, but any other oscillator tube should be adequate. My well-modified Globe Chief Model 90A uses a 6AG7 oscillator tube and this VXO works in it equally well. The only requirement is that your oscillator tube have a grounded grid through the crystal.

VXO circuit

Figure 1 shows the schematic diagram of the Junkbox VXO. Note that it's only necessary to install or insert coil L1 and variable capacitor C1 in series between the crystal and ground. If you have room, it would be more convenient to add these components to your rig. I built my oscillator in a minibox approximately 4" wide x 3" deep x 2-1/2" high for outboard use. Before this, however, I did install a VXO in the transmitter itself.



Schematic of the Junkbox VXO. The 6AQ5 oscillator circuit is described in Amateur handbooks.

If you don't have a minibox, you can get small metal cabinets in various sizes from Radio Shack. Two possibilities are their 3-1/2" x 2-1/8" x 4" cabinet (catalog no. 270-251) and their 4" x 2-3/8" x 6" cabinet (catalog no. 270-252). With a little ingenuity, you could probably use an empty coffee can, or something else of reasonable size. I suggest strongly that you use the smallest possible enclosure so that all leads are as short as possible. This will minimize unnecessary capacitive reactance. Make sure your enclosure is well shielded to reduce TVI.

Some years ago I built an inboard VXO for my rigs using a slug-tuned coil. A company called Calectro once made them in various ratings. I recall using one to cover 5 to 25 μ H. It worked pretty well, but is no longer available. The amount of frequency shift wasn't as great with these slug-tuned coils as it is with the coil used in this circuit.

The LC circuit

Basically I use 7-MHz crystals, but I can adjust the plate circuit for output at 7, 14, or 21 MHz. I can also use 3.5-MHz crystals and swing the frequency — although not as much as I can with the 7-MHz crystals. L1, which is about 40 μ H, works in series with C1 and the ground side of the crystal to swing the frequency. You can adjust the circuit to furnish either capacitive or inductive reactance to the crystal with these two components in a series-resonant circuit. This means C1 can tune from slightly above to considerably below the crystal's normal frequency.

About 35 pF is all that's normally necessary for full range with enough inductance to achieve the desired frequency shift. I didn't have a 35-pF variable capacitor, so I'm using a 50-pF variable. It works okay, but uses only the upper 30 to 35 pF. If you close or fully mesh the plates of a capacitor that's more than 35 pF, you'll notice that your crystals won't oscillate. However, the circuit will oscillate with a different tap on L1 and also with 80-meter crystals. A little experimentation is required.

The inductance of L1 is around 40 μ H. To determine this inductance, refer to the chapter "Electrical Laws and Circuits" in the *ARRL Radio Amateur's Handbook*, under the heading "Calculating Inductance." I happened to have an old Globe Chief tank coil 1 inch in diameter, using about no. 18 enameled wire, close wound. I use 58 turns, which approximates 40 μ H, and a tap at 29 turns — the half-way point.


Using coil taps

You may have a larger diameter coil or form. This might be a little better, but it requires fewer turns to achieve 40 μ H. It may be wise to have at least two taps, perhaps several, which should be close to the lower end of L1.

By using taps, you can reduce the shift or use 10-MHz crystals for the 30-meter band. The inductance is critical; too little gives insufficient shift, while too much produces a too fast tuning rate. It's best to use just enough inductance to shift to the desired frequency with the entire 30 to 35 pF of C1. Of course, as I mentioned before, if you have more capacitance in your variable capacitor you'll use only the upper 30 to 35 pF.

My L1 coil has just *one* tap because I use a bat-handle toggle switch to short half the coil. I found that by using this method with my 50-pF variable capacitor, I can tune a little higher or lower than the crystal frequency — depending on the switching arrangement used. You must experiment to achieve your desired results, but about 40 μ H of inductance is necessary.

There you have it, friends of the ether. I'm sure you will have satisfactory results with this VXO. If you have only a handful of crystals, you can change frequency up or down without being completely "rock bound." I find this method gives crystal stability with VFO capability.

You may find a few things to enhance this circuit. I'd appreciate any comments you have to offer. 

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HT MOBILE COMPANION

Handheld accessory uses just three circuit boards

By Peter A. Lovelock, K6JM, 1330 California Avenue, Santa Monica, California 90403

Handheld VHF and UHF transceivers (HTs or handy talkies) are popular with Amateur operators. Many use HTs for mobile operation; this gives you two radios for the price of one. However, using an HT in this mode has disadvantages. Low audio comes into a tiny speaker and may be drowned out by traffic noise before it reaches your ears. You might also experience battery discharge in the middle of a QSO.

The Mobile Companion alleviates these difficulties, and does so in a compact unit. One interconnecting cable to the HT provides the following features:

- Regulated 10.5-volts DC at 1 A to power the HT and save the battery.
- A constant voltage charger, which quickly replenishes a dead battery pack.
- A receiver audio booster, which delivers 2 or 8 watts to an external speaker.
- An optional PA system for emergency use and monitoring the receiver when you're not in the car.

A full schematic is shown in **Figure 1**. The battery charger and 8-watt amplifier (which doubles for receiver audio boost and PA operation) are included.

Not everyone will want all the features. I used three small pc boards for the audio amplifier and battery charger — which offer simple, flexible choices.

Some HTs have a jack accepting up to 16 volts DC directly from the car battery. This doesn't protect against transient switching spikes or auto regulator failure, which may place a harmful voltage on the HT. The LM7805 regulator provides damage insurance; it helps relieve filter alternator whine and other noise. The regulator is set for 10.6 volts DC (the most common supply for HTs) and maintains regulation with the car battery to 12 volts DC. This means a sacrifice in RF output power for 5-watt RF HTs using 12-volt battery packs.

I needed a small under-the-hood PA speaker for use with our local emergency team. I replaced the HT's 2-watt amplifier board with an 8-watt commercial module (see parts list). Two miniature relays switch the amplifier input and output for radio or PA mode.

Functional description

SW1 and SW2 (DPDT slide switches) select the radio and PA modes, as shown in **Table 1** and **Figure 1**.

Setting SW1A in the R position connects the mic to the HT mic input for normal radio operation; in the P position it connects the mic to preamp Q1 for PA mode. Putting SW1B in the P position turns on flashing LED CR1 and energizes K2 to connect J1 (PA speaker) to U2 amplifier output. Coil K1 is also supplied voltage, but SW2A in the R position keeps K1 de-energized and the amplifier input connected to R8 (radio gain control), for radio monitoring through the PA speaker. Placing SW2A in the P position allows K1 to energize through SW2B, connecting the amplifier input to R10 (PA gain control), and operating the PA from the mic. In the P position SW2B causes LED CR2 to light, showing PA mode.

TABLE 1

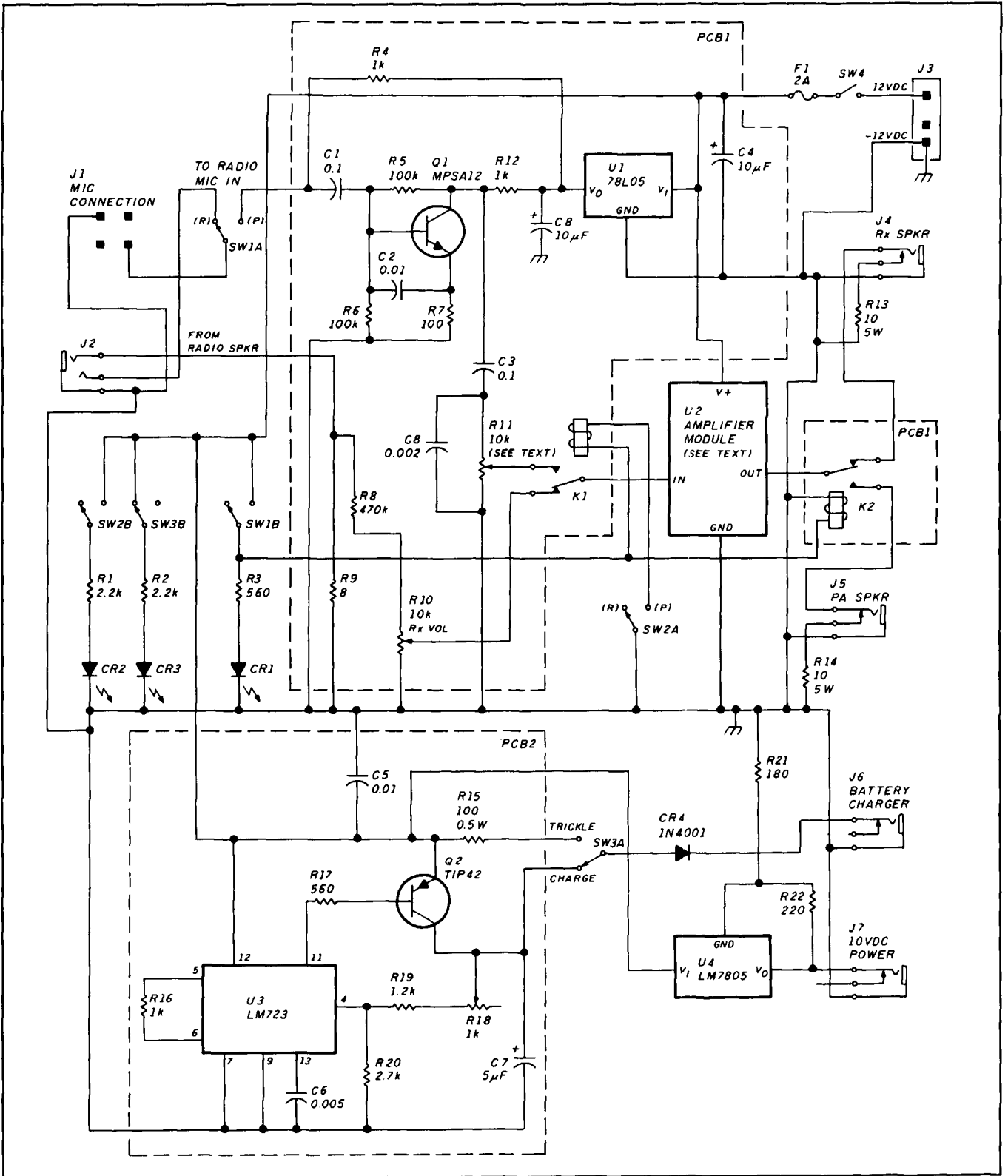
Mode select arrangement		
SW1	SW2	Mode
Radio	PA	Radio
Radio	RX	Radio
PA	PA	Mic/PA
PA	RX	RX/PA

Preamp Q1 provides gain for the mic to drive the 8-watt amplifier to full output. It's configured for a high-impedance dynamic or Electret™ (capacitive) mic. R4, a 1-k resistor, is required only with an Electret mic and should be excluded when you're using a dynamic or ceramic mic. VCC 5 volts for the preamp is supplied by U1, a 78L05 regulator (which also helps isolate car-generated noise from the mic line).

The charger circuit was originally developed by Joe Moell, W0OV. It comprises Q2 — a low-saturation PNP pass transistor TIP42A — controlled by U3, a μ A723 voltage regulator. Do not substitute another transistor for the TIP42A. The charger circuit is an adjustable constant-voltage type.

A fully discharged battery pack will draw high current (about 750 mA) at first, tapering down to approximately 25 mA as the battery pack is fully charged. SW3A selects full charge or a trickle charge of 10 mA; SW3B controls CR3 to show when the full-charge status is on.

FIGURE 1



Schematic of the Mobile Companion.

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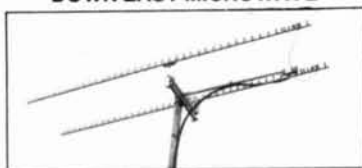
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PARTS LIST

Reference	Description	Radio Shack	Quantity
PC BOARD 1, CONTROLLER			
Q1	Transistor MPSA12		1
U1	Voltage regulator 78L05		1
K1,K2	Relay, micro, SPDT, 12-volt coil	275-241	2
R10	Trimmer potentiometer, 10 k	271-335	1
R4-R12	Resistors, 1/4 watt, composition		9
C1,C3	Capacitor, ceramic, 0.1 μ F, 16 volts DC		2
C2	Capacitor, ceramic, 0.01 μ F, 16 volts DC		1
C4,C9	Capacitor, 10 μ F, 25 volts DC		2
C8	Capacitor, ceramic, 0.002 μ F		1
PC BOARD 2, CHARGER			
Q2	Transistor, PNP power, TIP42		1
U3	IC voltage regulator, LM723	276-1740	1
R16,R17,			
R19,R20	Resistor, 1/4 watt, composition		4
R18	Trimmer potentiometer, 1 k	271-333	1
R15	Resistor, 100 ohms, 1/2 watt		1
C5	Capacitor, ceramic, 0.01 μ F, 16 volts DC		1
C6	Capacitor, ceramic, 0.005 μ F, 16 volts DC		1
C7	Capacitor, tantalum, 5 μ F, 16 volts DC		1
	Insulated mounting kit for Q2	276-1373	1
PC BOARD 3, OPTIONAL 2-WATT AUDIO AMPLIFIER			
U4	IC amplifier, LM380		1
C1	Capacitor, ceramic, 120 pF, 10 volts DC		1
C2	Capacitor, electrolytic, 47 μ F 10 volts DC		1
C3,C4	Capacitor, ceramic, 0.1 μ F, 16 volts DC		2
C5	Capacitor, electrolytic, 470 μ F, 16 volts DC		1
R1	Resistor 2.7 ohms, 1/2 watt, composition		1
CASE COMPONENTS			
U4	IC voltage regulator 7805	276-1770	1
CR1	LED, flashing	276-036	1
CR2,CR3	LED, mini, red	276-026	2
CR4	Rectifier, diode, 1N4001	276-1101	1
SW1, SW2,			
SW3	Switch, submin slide, DPDT	275-407	3
R1,R2	Resistor, 2.2 k, 1/4 watt		2
R3	Resistor, 560 ohm, 1/4 watt		1
R11	Potentiometer, 10-k, 2-watt AF taper	271-1721	1
R13,R14	Resistor, 10 ohm wire wound, 5 watts		2
F1	Fuse holder, 5x20 mm fuses	270-362	1
J1	Standard 4-pin mic socket	274-002	1
J2	3-wire stereo jack, open	274-279	1
J3	Polarized, 3-pin power socket		1
J4,J6	Submin 2-wire jacks	274-292	2
J5	Min 2-wire jack, closed	274-296	1
J7	DC power jack, coaxial	274-1563	1
	Insulated mounting kit for U4	276-1373	1
U2	Optional 8-watt audio amplifier module, GC Electronics, catalog J4-596.		
	Case, aluminum 3" x 3" x 4"		1
Set of 3 pc boards etched and drilled may be obtained from: R. & R. Associates, 3106 Glendon, Los Angeles, California 90034, \$6.75 ppd.			

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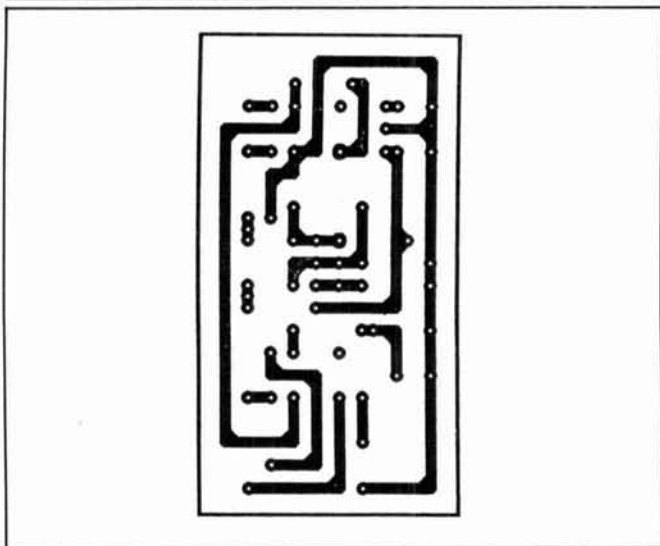
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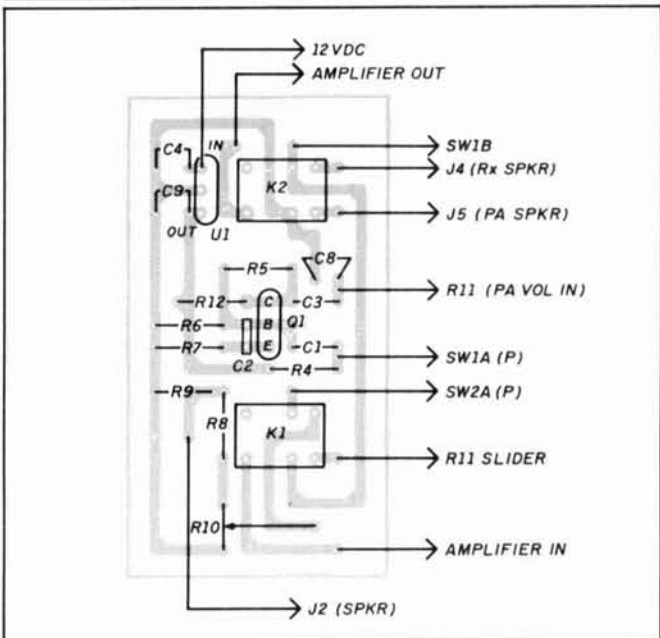
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FIGURE 2A



PC board 1, controller, etch side.

FIGURE 2B

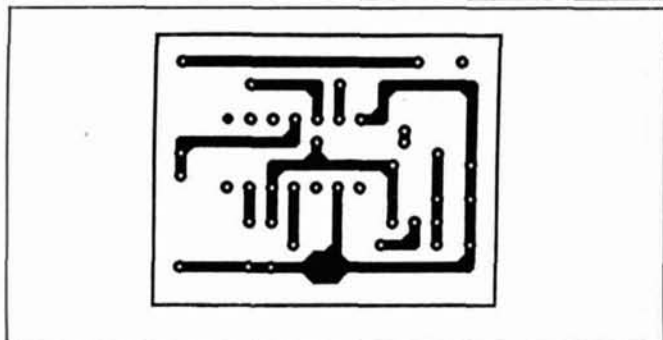


PC board 1, controller, component layout.

Construction

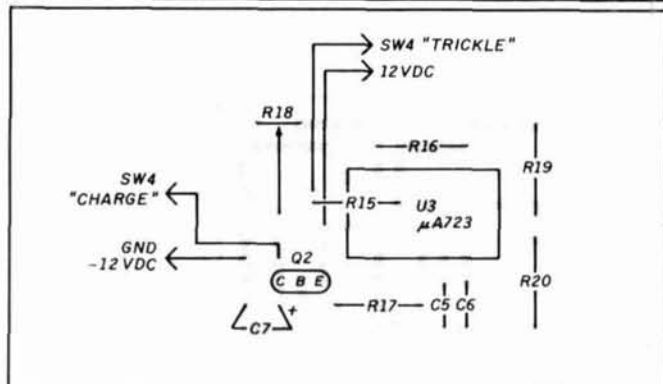
Most of the circuitry is on two boards. You'll need a third board for the 2-watt audio amplifier if you use it instead of a commercial amplifier module. Board 1 includes the mic preamp, K1 and K2 relays, receiver audio-gain control R9 (preset trimmer), and 5-volt regulator U1. Board 2 includes all the parts for the battery charger. Both boards can be assembled on perfboard using point-to-point wiring. For those who prefer them, pc board layouts are shown in Figures 2, 3 and 4. R10 (PA volume), SW1, SW2, SW3, and J1 and J2 are panel mounted.

FIGURE 3A



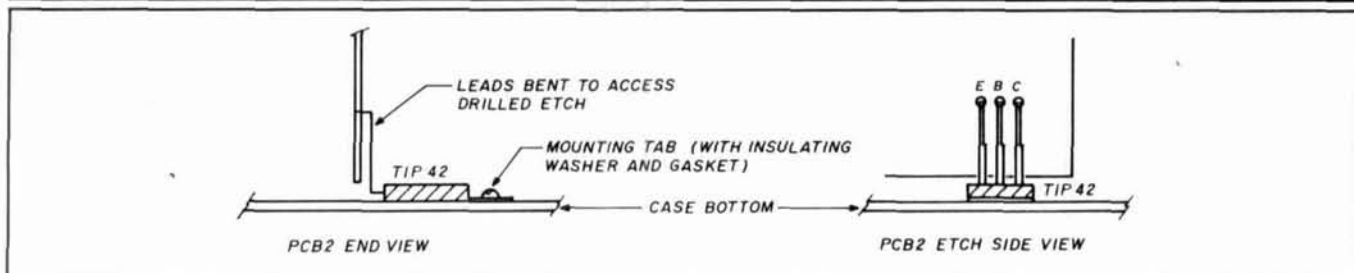
PC board 2, charger, etch side.

FIGURE 3B



PC board 2, charger, component layout.

FIGURE 3C



Mounting TIP42 transistor to pc board 2.

When assembling the boards, allow adequate lengths of input/output wire (no. 22 insulated) to reach the destination point on another board or panel control. Use miniature shielded wire (where shown) for connections to the amplifier input and output. It's important to avoid ground loops, particularly with the high gain associated with the preamp and 8-watt amplifier. **Figure 5** shows a "star-type" grounding hookup. This hookup avoids daisy-chain ground loops, which can cause oscillation. The incoming -12 volts DC power line connects to a ground lug mounted to the case with external tooth washers and hardware for low-resistance connections. A no. 20 stranded, insulated wire is run from this lug to the V-input on the amplifier board or module. This is the main

ground bus; all other signal grounds must be connected to it to avoid instability.

Components located on the boards are shown within the dotted lines on the schematic, **Figure 1**. If you want to use pc boards, you can make them using the pc board drawings or obtain them from the source shown in the parts list. If you decide to use point-to-point wiring, these drawings will help you with component layout. Board 1 (controller) is shown in **Figure 2A** (etch side); **Figure 2B** shows the component layout and interconnecting wires. **Figures 3A, B, and C** show board 2 (charger), and **Figures 4A, B, and C** the 2-watt audio amplifier. For convenience, the parts list is subdivided by board number or chassis/panel mounting.

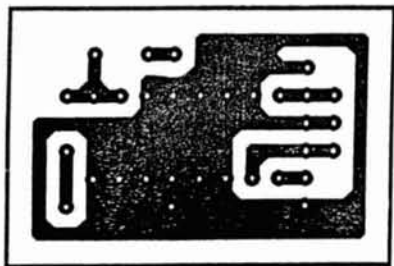
Assemble all resistors and capacitors for board 1; then solder them and clip excess leads. *Note that R10 (PA volume) shown in the **Figure 1** schematic is not mounted on the board, but on the front panel.* It's connected to the board with miniature shielded cable. Connect the shields to the board ground at one end and to the R10 ground lug at the other. Finally, assemble active components Q1 and U1, solder, and clip excess leads.

Follow the same steps for assembling and soldering board 2. However, note that TIP42 power transistor Q2 is connected to the board on the etch side. The leads are bent so that the device can be mounted on the case bottom for heat sinking. **Figure 3C** illustrates this assembly. Q2 is mounted to the case with a mica insulating washer and no. 4 screw shoulder washer, so that the collector tab is insulated from the case. The mounting hardware is available as a kit, shown in the parts list. After you've put everything together, check your work with an ohmmeter to make sure that Q1 tab is insulated from the case. If it isn't, the Q2 output will be shorted to ground and may damage the device.

LEDs CR1, CR2, and CR3 are mounted on the front panel. Mounting-hole size depends upon the LEDs used. The mounting holes should allow the LEDs to fit snugly when inserted from the rear of the panel. Once the diodes are installed, put a drop of Crazy Glue™ on the rear panel and also on the LED. (This anchors the LED to the rear panel.) Connect one lead of R1, R2, and R3 to the respective LED anode lead. Do this by twisting the leads, soldering them, and snipping off the excess. Connect the LED cathode leads in common with short lengths of no. 22 wire, and then to the case ground. Connect the other LED leads directly to SW1, SW2, and SW3, as shown in **Figure 1**.

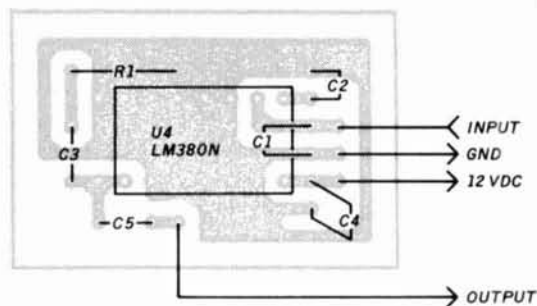
Figure 6A shows a hole-drilling guide. The case I used is 3" x 3" x 4" aluminum. Radio Shack no longer stocks this size, but catalogs the same type in size 3"

FIGURE 4A



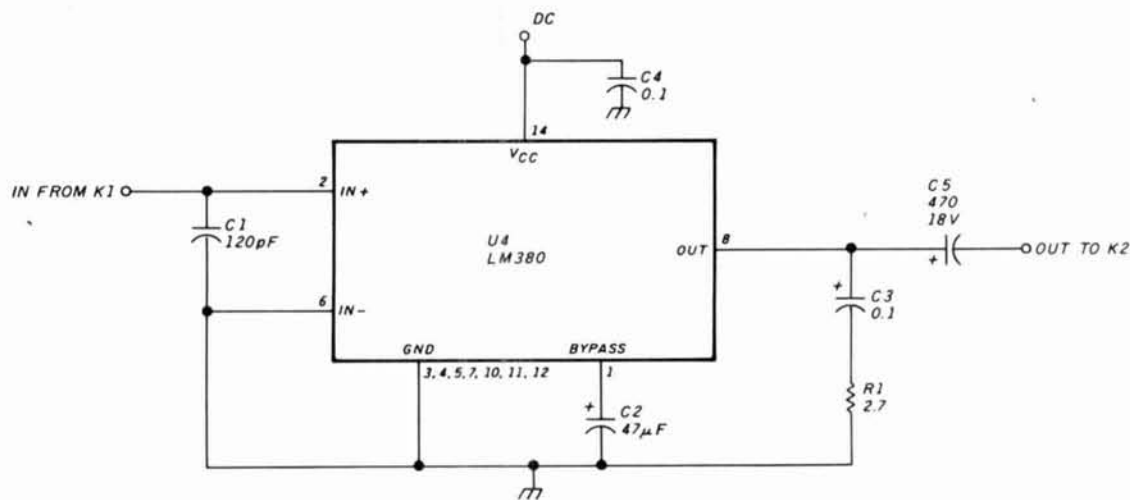
PC board 3, 2-watt amplifier, etch side.

FIGURE 4B



PC board 3, 2-watt amplifier, component layout.

FIGURE 4C



Schematic of 2-watt amplifier.

× 5-1/4" × 5-7/8". If you use this case you must modify the drilling guide. I like the latter because it allows more space for ease of construction. Each board is mounted with a single small L-bracket to the case bottom (Figure 6B). The 8-watt audio amplifier (if used) is mounted with two holes and 1/4-inch standoffs. You won't need the holes marked with an X

(Figure 6A) if you exclude the PA feature. Apply the panel markings after drilling all the holes; use a dry transfer kit. Don't try to do this after mounting the components to the panels. It's a good idea to let the lettering dry overnight before putting on any kind of protective coating. If you don't, the liquid coating may cause the lettering to float off the panel surface. You can use Datacoat™ or clear nail varnish to cover the lettering; it should be *flowed on* gently with a small brush. I prefer two light spray coats of Varathene™ plastic lacquer applied evenly to the whole panel. This dries to a glass-hard finish, which resists scratching.

Cable assembly

Your final step is to make the cable which connects the Mobile Companion to your HT. Figure 7 shows this cable. It's made of two lengths of miniature single-conductor shielded wire and one length of two-conductor shielded wire.

The connectors for the Companion end are determined by the jacks mounted on the Companion, as shown in the drawing. Select the connectors for the other end based on your radio's connectors. Some HTs have coaxial jacks for external power input with the center pin grounded and V+ on the outer ring. Others have V+ on the center pin and the outer ring is grounded. The Companion power jack has the center pin at V+. If your HT input is center-pin grounded, you'll have to be careful when making a polarity reversal in this cable. Connect the shield to the center pin and wire to the sleeve, as shown in Figure 7.

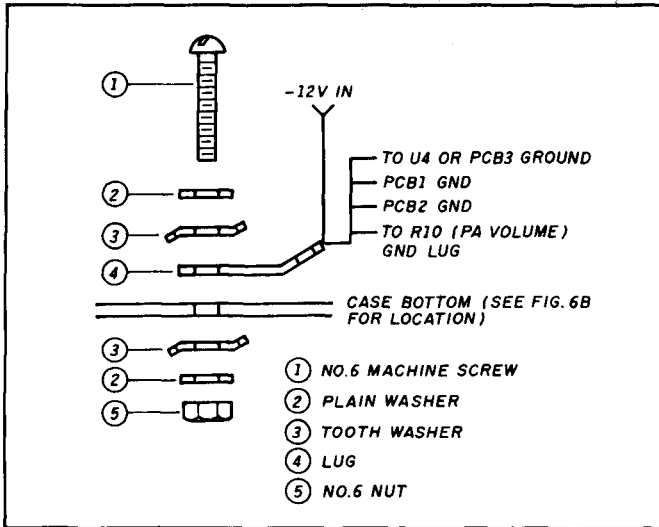
Cable assembly length is determined by the distance between the Companion and the HT inside the car. You can bundle the finished cables into a single assembly with plas-

tic cable ties, dressing the length of each to reach the jacks on the HT.

Checkout and adjustment

I use a small 0 to 20-volt 200-mA power supply for my initial checkout of any project. This ensures that any shorts or

FIGURE 5



"Star ground" arrangement.

miswirings have minimal chance of causing damage. Apply 13.6 volts to the input terminals. Check that the U1 output is 5 volts DC and the U4 output is 10.5 volts. If they check out, it's safe to assume that there are no power shorts.

Adjusting HT power

If the voltage measured at U4 OUT is more or less than 10.6 volts DC, vary the value of R21 until the voltage is 10.6 volts DC. If your HT is normally 12 volts DC, you can change R21 to obtain 12 volts regulated output. However, to stay in regulation, the input from the car battery will have to be at least 14 volts (2 volts more than the regulator output). When the engine is idling or stopped, the battery voltage falls to about 12 volts. If your HT has a 6-cell battery pack (7.5 volts), adjust R21 to about 120 ohms to measure 7.5 volts from U4.

Adjusting the battery charger

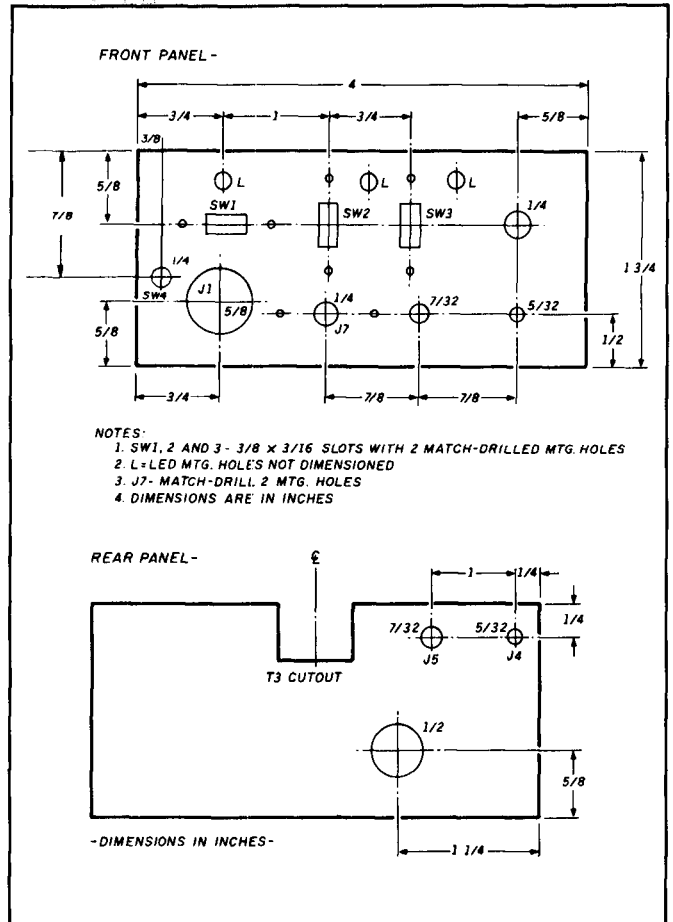
Note that CR4, a 1-A diode, is connected in series with the charger output and J8. CR4 is wired directly from SW3A to the connector. The purpose of this diode is to prevent the battery from shorting when the line is plugged into the HT. If your HT has an internal diode for the same reason, omit CR4 and run a wire between SW3A and J8.

Measure the voltage at SW3A (Q2 collector) with a digital voltmeter, making sure there is no battery connected to J8 and that SW3A is switched to CHARGE. Adjust R18 until the voltmeter reads 12 volts. This is equal to eight cells fully charged to 1.43 volts, plus the 0.7-volt drop through CR4. If you don't use CR4, adjust for a DVM reading of 11.4 volts.

If your HT has a six-cell, 7.5-volt battery, adjust R18 for a DVM reading of 9.3 volts with CR4 in the circuit, or 8.6 volts if CR4 is not used.

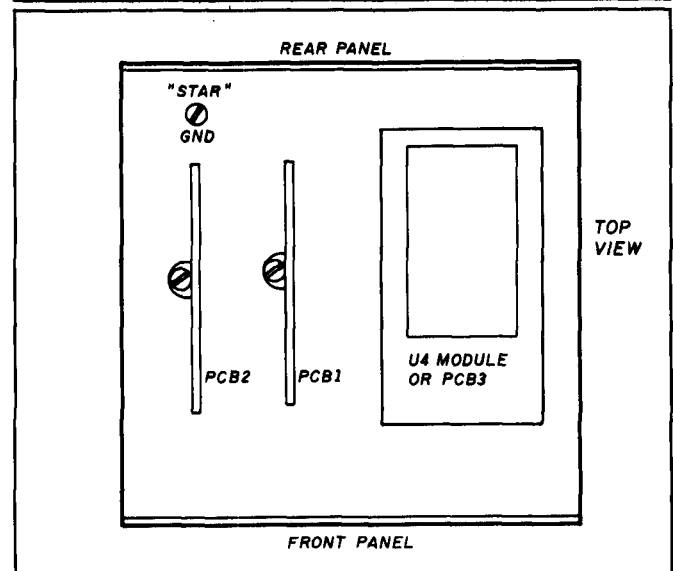
Once the unit is set up for your battery voltage, you don't need to make any further adjustment of R18. Charge taper is automatic. You can check current limiting by connecting a 10-ohm, 10-watt wire-wound resistor across J8 in series with a 1-A current meter. You should get a reading of 650 to 850 mA. For a final check, charge the battery fully with its regular charger.

FIGURE 6A



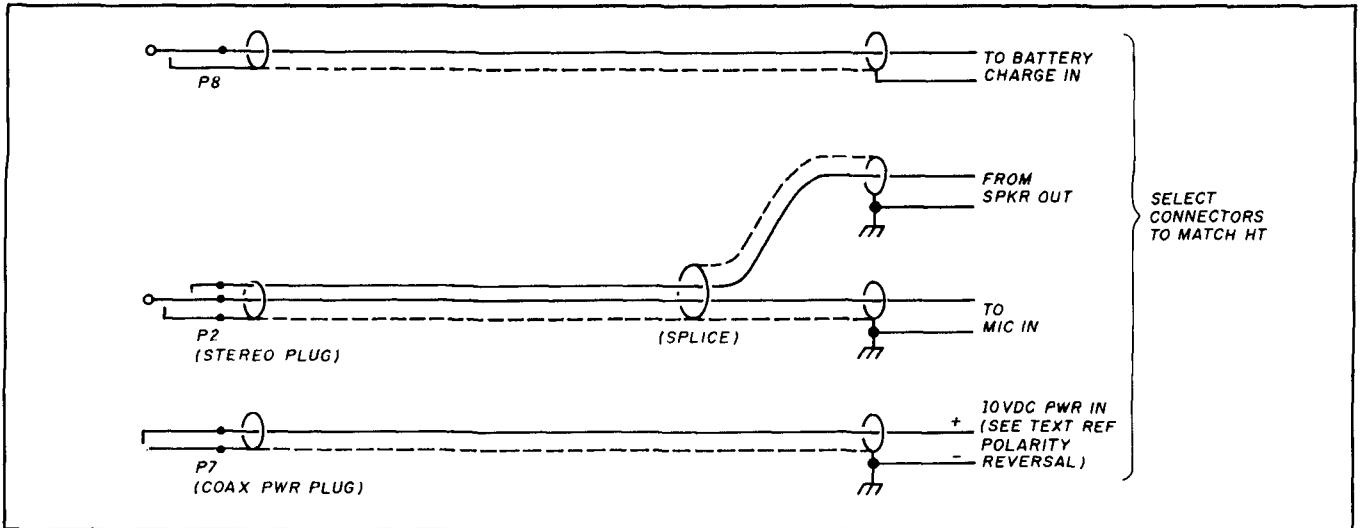
Case drilling guide.

FIGURE 6B



Case component locations.

FIGURE 7



Interconnecting cable assembly.

Connect it, in series with a 100-mA current meter, to J8. It should read approximately 25 mA. Switch SW3A to TRICKLE; this should give a current reading of 10 mA. If the reading is higher, increase the value of R15 (100 ohms) to get a reading of 10 mA or less.

One last note: If your HT doesn't have separate inputs for POWER and CHARGE, you obviously can't use the setup as described. In this case you may omit the U4 regulated power circuit and cable, but have the charger connected during operation. The HT will operate from its battery during transmit mode, and the battery will receive a "boost" charge during receive cycles. This is an acceptable alternative that I used for several years without having a battery run down on me.

Adjusting the receiver audio amplifier level

R10, the trimmer potentiometer on pc board 1, is a set-and-forget control for adjusting the audio level to the external speaker. Connect the HT speaker output to J4. Make sure SW1 is set to RADIO. Adjust R10 fully counterclockwise. Turn on both units. Set the HT to an active channel. Turn the HT volume fully clockwise. Adjust R10 until audio level from the external speaker is a little too high for comfort. A comfortable audio level is always regulated by the HT volume control; R10 needs no further adjustment.

Checking the PA system (optional)

Plug the microphone into J1 and PA, or a test speaker into J5. Set SW1 and SW2 to PA. Set R11 (PA VOLUME) counterclockwise. Turn on Companion. Advance PA volume slowly clockwise while talking into a microphone. **Caution:** If a test speaker is close by, you will get loud audio feedback. Run the speaker in another room, if possible. Disconnect the mic. Advance PA volume fully clockwise; an increasing hiss should be heard.

If an audible howl occurs near full volume, or if the hiss level drops suddenly near full volume-control setting, instability is present. The drop in the hiss level is caused by high-frequency oscillation above the audible level. If this happens, check U2 output with an oscilloscope or AC voltmeter for obvious oscillation. If it is present, it's almost certainly due to poor grounding. Check to see that your grounding conforms with that shown in Figure 5. Do not let the 8-watt amplifier continue to run in an oscillating mode. Severe overheating and damage will occur. *hr*

BIBLIOGRAPHY

1. Joe Moell, K0OV, "A Fast Charger and Regulator for the Tempo S," QST, June 1981, page 35.

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WANTED: Schematic or any other information concerning the AN/JRR58 receiver from Squires-Sanders. Wanted also SS-1R, SS-1VRS in good condition: Marion Arthur Ambrose Jr, Contel Page Inc, APO New York, NY 09794-007.

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IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14,280 MHz, 1-3 PM Eastern. Nine hundred Amateurs in 40 countries. Rev. Thomas Sable, S.J., University of Scranton, Scranton, PA 18510.

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ONTARIO: February 3. The Niagara Peninsula ARC's 12th annual Big Event, hamfest and dinner dance, C.A.W. Hall, 124 Bunting Road, St. Catharines. Admission \$3. Talk in on 147.24/84. For information write NPARC, PO Box 692, St. Catharines, Ont. L2R 6Y3 or phone (416) 682-4844.

NEW YORK: February 11. Melville. Hamfest sponsored by the Long Island Mobile ARC, Electricians Hall, 41 Pinelawn Rd. 9 AM to 4 PM. \$5 at door. No advance. Talk in on 146.25/85. Contact Neil Hartman, WE2V (516) 462-5549 or Mark Nadel, NK2T (516) 799-2366.

OHIO: February 11. The Mansfield Mid-Winter Hamfest/Computer Show, Richland County Fairgrounds, Mansfield. Doors open 7 AM. Tickets \$4/advance; \$5/door. Tables \$7/advance; \$10/door. Talk in on 146.34/94. For information of tickets/tables SASE to Dean Wrashe, KB8MG, 1094 Beal Road, Mansfield, Ohio 44905. (419) 589-2415 after 4 PM EST.

MASSACHUSETTS: February 17. Electronic Flea Market sponsored by the Algonquin ARC. Marlboro Middle School Cafeteria, Union St, Marlboro. 10 AM to 3 PM. Sellers 8 AM. Admission \$2.00. Tables \$10/advance; \$12/door. Wheelchair accessible. For information Ann, KA1PON (508) 481-4988 or write AARC, Box 258, Marlboro, MA 01752.

FLORIDA: February 17-18. Sarasota Hamfest, South Florida Section ARRL Convention, Computer Show, Roberts Arena, 3000 Rigling Blvd, Sarasota. 9 AM to 4 PM. Admission \$5/advance; \$7/door. Talk in on 146.31/-91, 147.90/-30 and 449.425/-925. For information contact Hadley Carrigan, N4ODK, 101 N. Adams Drive, Sarasota, FL (813) 388-2868.

INDIANA: Saturday, February 24. The LaPorte ARC's Winter Hamfest, LaPorte Civic Auditorium. Donation \$3.50. Talk in on 146.52 simplex. For table reservation (\$3.50) SASE to LPARC, POB 30, LaPorte, IN 46350.

MINNESOTA: February 24. The 9th annual Midwinter Madness Ham Radio and Computer Show, Medina Ballroom, Highway 55. Sponsored by the Robbinsdale ARC. 7 AM to 2 PM. Admission \$3/advance; \$5/door. Talk in on 147.60/00. For information Bob Arel (612) 545-1059 evenings and (612) 425-2424 days.

FLORIDA: February 24. Brooksville. The Hernando County ARA's 8th annual Hamfest, Hernando County Fairgrounds auditorium. Tickets \$3/advance; \$4/door. Swap tables \$8. For tickets/table reservations SASE with check to Hernando County Amateur Radio Association, PO Box 34605-1721, Brooksville, FL 34605.

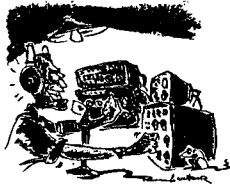
VERMONT: February 24. Milton. The Northern Vermont Mid-Winter Hamfest, Milton High School, Rt. 7. 9 AM to 3 PM. Admission \$2. Free tables FCS. Talk in on 145.47/-600. For information Mitch Stern, WB2USJ (802) 879-6589 or Tom Taylor, N1EXY (802) 893-4834.

TEXAS: February 24. The Orange ARC's fifth annual Hamfest-Flea Market, National Guard Armory, Meeks Drive, Orange. 8 AM to 4 PM. For information Sherwood Buckalew, KA5VOT (409) 883-6111. Dan Killough, WB4GYS (409) 769-9603 or Rick Lewis, KF5GV.

MICHIGAN: February 25. The 20th annual Livonia ARC's Swap 'n Shop, Dearborn Civic Center, Dearborn. 8 AM to 4 PM. Talk in on LARC Repeater 144.75/5.35 and 146.52 simplex. For information SASE to Neil Coffin, WA8GWL, Livonia ARC, POB 2111, Livonia, MI 48151.

VIRGINIA: February 25. Winterfest sponsored by the Vienna Wireless Society, Vienna Community Center, 120 Cherry Street, Vienna. Setup 6 AM. Public 7:30 AM. Admission \$5/door. \$10/altgate. No advance. For information Harry Kalklikian, WAACN, 4941 Andrea Avenue, Annandale, VA 22003. (703) 978-4402.

OHIO: February 25. The Cuyahoga Falls ARC's 36th annual Hamfest, Akron North High School. 8 AM to 3 PM. Tickets \$3/advance; \$4/door. Tables \$5/advance; \$6/door. Wheelchair accessible. For information Bill Sovinsky, KB3SL, 2305 24th St, Cuyahoga Falls, OH 44223. (216) 923-3830.



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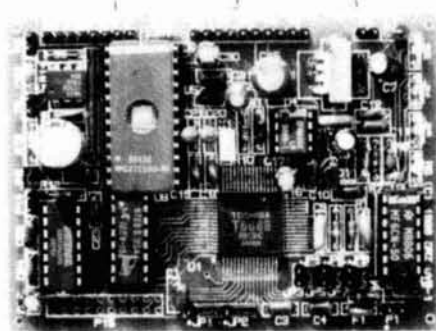
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(continued from page 71)

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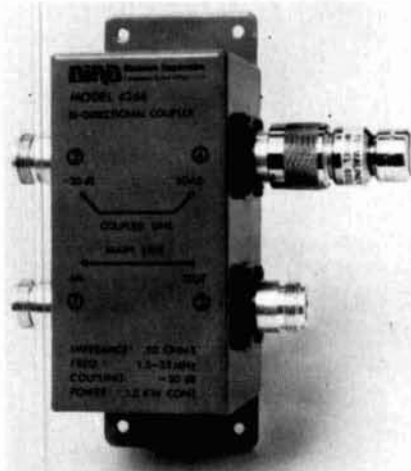


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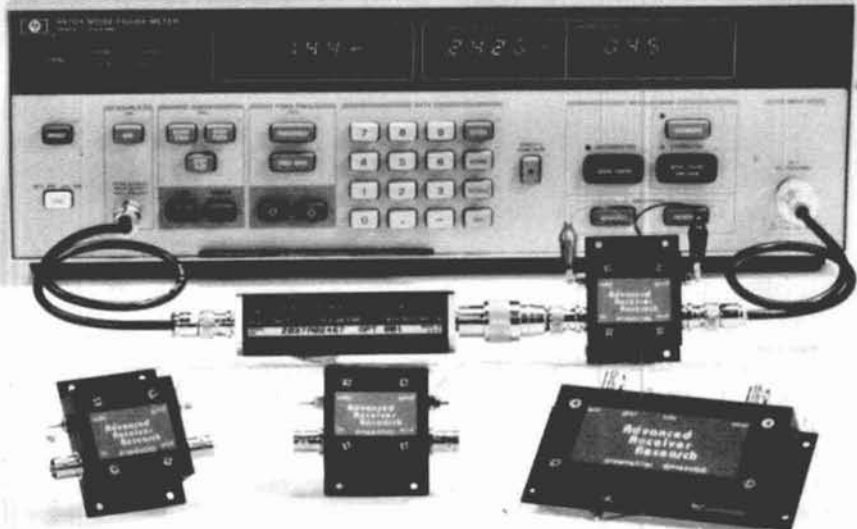
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SP220VDA	220-225	<1.3	15	0	DGFET	\$67.95
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Elmer's Notebook

Tom McMullen, W1SL

ELEMENTARY ELECTRONICS: CAPACITORS

In previous columns I've discussed magnetic fields and resistance. This month I'd like to explore another important element in electronics — capacitors and how they work.

In simple terms, a capacitor is an energy storage device; it stores electrons. The number of electrons a capacitor can store, and how long they remain stored, depends on such things as the size of the electrodes (more surface area allows more electrons to be stored) and the type of dielectric (material between the electrodes) used. For long term storage, the dielectric must not act like a resistor. This means it can't let the electrons "leak" from one plate to the other.

What's in a capacitor?

There are several ways to make a capacitor. If you've ever looked through one of the "grab bag" assortments at a supply house or hamfest flea market, you know that capacitors come in a variety of shapes and sizes. A very basic capacitor is simply two metal plates or disks parallel to each other, with air between (see Figure 1). In this type of capacitor either plate can be of either polarity (+ or -), and the air between is called the dielectric. The dielectric material between the plates acts as insulation to keep the plates apart and concentrates the electrostatic lines of force between them. The standard dielectric constant for air is approximately 1.006 (vacuum is rated at 1.0). Other materials have higher numbers. This means that if you have a pair of plates with air between and a capacitance of 100 pF, and you replace the air with some material that has a dielectric constant of 2.0, you end up with a capacitance of approximately 200 pF for the same size. The capacitance is directly proportional to



the area of the plates and the dielectric constant of the material between them. It increases when you decrease the spacing between the electrodes or plates. Paper, glass, ceramic, and many plastics are types of dielectric materials used in capacitors. Each material has characteristics making it desirable for one use or another. Table 1 gives the constants of some of the most common materials.

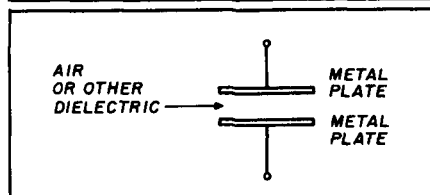
Some capacitors, called electrolytics, have a paper or plastic material between the electrodes that is impregnated with a chemical which increases the dielectric constant to many times that of air. This allows a much greater

"storage capacity" than is found in a capacitor of the same size using plain paper, glass, ceramic, or air dielectric. Such units are usually polarized, and any voltage applied must be of the correct polarity to prevent damage to the capacitor or power supply. If the applied voltage has the wrong polarity, the dielectric will usually overheat and short circuit. When too much voltage or reverse voltage was applied to the older "wet" electrolytic capacitors, they would "cook" and release a jet of steam with a startling bang. Modern electrolytic capacitors use a dry electrolytic which usually isn't prone to this type of destruction, but simply becomes a short circuit.

What's a pF?

The basic unit of capacitance is a farad, named after physicist Michael

FIGURE 1



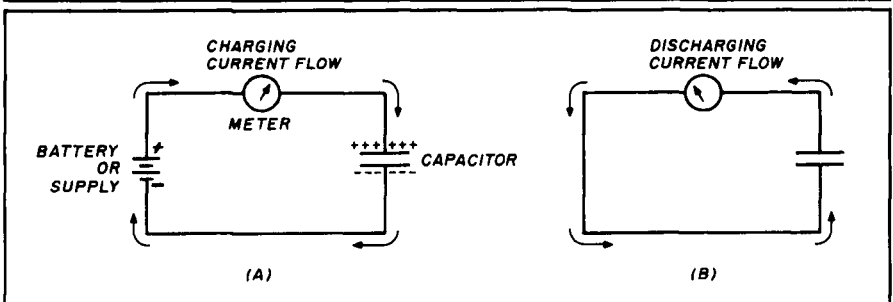
A simple representation of a capacitor — two metal plates with air or some other insulation (dielectric) between them.

TABLE 1

Dielectric constants for some common materials.

Vacuum	1.000
Air (dry)	1.006
Rubber	2 or greater
Paper	2 or greater
Ceramic	3 to 7
Glass	4 to 8
Mica	6 or 7
Water (pure)	80
Barium Titanate	7,000 or higher

FIGURE 2



When a capacitor is connected to a power source, electrons flow from one plate into the other, causing an instantaneous charging current (A). When the capacitor discharges through an external circuit, the current flow reverses (B).

Faraday. A farad is a very large unit. The "official" description says that a capacitor "has a capacitance of 1 farad when a change of 1 volt per second across the unit produces a current of 1 ampere through it." Another definition says that a 1-F capacitor can store a *difference* of 1 coulomb (6.25×10^{18} electrons) between the plates. These definitions tell us something about "changing" voltages and current, but what does this mean in practical terms? In the days before modern materials were developed for miniaturization, the saying was that a 1-F capacitor (they were sometimes called "condensers") would be about the size of a one-car garage. More recently, someone estimated that a medium-sized 12-volt automotive battery had the approximate storage capacity of a 1-F capacitor. A farad is also far larger than anything we work with in electronics, so rather than write out all the decimal fractions of a farad for our circuits and calculations, we use abbreviations. One common unit is a microfarad, abbreviated μF . Another is a picofarad, or pF. There is also a nanofarad, nF, which is not widely used in the United States.

Here's what those abbreviations mean in decimal terms:

1 μF = 0.000,001 farad

1 nF = 0.000,000,001 farad

1 pF = 0.000,000,000,001 farad

You can see why the abbreviations are so helpful! As a general rule, the larger units — like 1 μF to several hundred μF — are used to filter out noise, hum, or ripple in power supply circuits and to pass lower audio frequencies from one stage to another. The smaller values, from 1 pF to .01 μF , are often used in tuned circuits or in critical coupling between radio frequency circuits.

How capacitors work

As I noted before, a capacitor is a device that stores electrons. When two electrodes separated by a dielectric are connected to a source of power, there's an immediate rush of free electrons from the power supply to the plates (see **Figure 2A**). This is where the "current flow" part of the definition comes into play. Because the electrons are flowing from the supply to one plate, and from the other plate to the supply, current *appears* to flow through the capacitor. Because of the dielectric, the electrons can't cross from one

plate to the other. Instead, they build up on one plate, creating a surplus of electrons like that found on the terminal of the battery. The other plate, which is connected to the battery terminal that has a deficiency of electrons, takes on the same characteristic as that terminal. It too has a deficiency. When a circuit is completed between the plates, the electrons try to get to the other side, causing a brief current flow out of the (+) plate to the (-) one. Interestingly enough, the positive (+) terminal is the one that has a deficiency of electrons while the negative (-) terminal has a surplus of electrons. However, we still say that current flows from positive to negative — even though electrons are going the other way. (Old habits are hard to break.)

This means you have two plates with electrons trying to get across the gap, and in this condition the capacitor is said to be "charged." If the dielectric is of good quality (dry air, for example), the charge will remain for a long time — minutes, or even hours — after the power supply is disconnected. However, many dielectric materials aren't that good; the charge leaks off in a few minutes through the resistance caused by moisture and poor insulation used in manufacturing the capacitor.

Taking advantage of the rush

The tendency of electrons to rush from one plate to the other is put to good use in tuned circuits. When a capacitor is connected across an inductor, as in **Figure 3**, something interesting takes place. For purposes of this discussion, imagine that one plate of the capacitor has been given a charge from a power supply. The

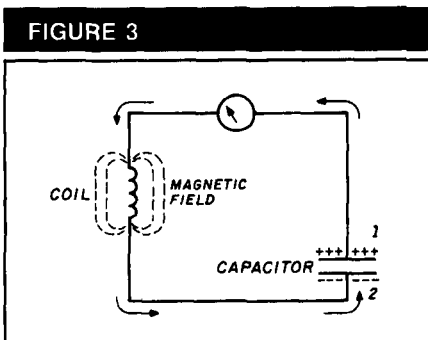
supply is then disconnected and the inductor is connected across the capacitor. The extra electrons on one plate try to rush to the other side through the wire in the coil. But as I mentioned in an earlier column about magnetic fields, a field is created whenever current flows in a wire, and if the wire is wound into a coil the field is more intense. In addition, the field from the first turn in the coil cuts across the second turn, and the field from the second turn cuts across the third turn, and so on. It so happens that when the magnetic field cuts across the second turn, it generates a small current flow opposite in direction to the current flow in the first turn. This opposes the current flow from the capacitor and increases the time required for the capacitor to discharge. The capacitor eventually wins out and the electrons do get to the other plate, but not as quickly as they would through a short circuit. When the electrons stop moving, the magnetic field around the coil collapses, releasing the stored energy back into the capacitor. As a result, plate two now has too many electrons. This starts a rush to get to the other side again, and the inductor tries to slow things down as before. This process continues until the resistance of the wire and the energy in the magnetic field have depleted the charge of electrons.

The amount of time that the electrons require to move from one plate to the other is an indication of the resonant frequency of the circuit. For instance, if the size of the capacitor (number of electrons stored) and the effect of the inductor (intensity of the magnetic field) allow the electrons to complete their journey 1,000,000 times in one second, it is resonant at one million cycles per second. Older hands will remember this as 1 megacycle (1,000,000 cps), but modern terminology uses hertz as a term meaning "cycles per second," so this circuit is resonant at 1 megahertz, or 1 MHz.

Obviously, changing the size of the capacitor (number of electrons it can store) or the inductor (larger or smaller magnetic field) will change the cycle time of the circuit, thus changing the resonant frequency.

Uses for the larger capacitors

Large value capacitors, especially those of one to several thousand μF ,



Current flow is impeded by a coil (inductance) and its magnetic field. The interaction between the coil and capacitor creates a resonant circuit — a vital part of all radio circuits.

are widely used in audio and power supply circuits to remove hum and noise and provide the smooth DC required for almost all circuits. The output from the rectifier diode in a simple power supply (point A in Figure 4) is simply half of a sine wave. But it's not really useful in that form. An audio amplifier with this type of power applied would have a loud hum. Here's where the capacitor comes to the rescue. In the period when this pulsating voltage is present at the output of the rectifier, the capacitor is storing some of the energy. When the voltage drops to zero between diode conduction cycles, the capacitor releases that stored energy to the output and whatever is hooked to it. The result is a somewhat smoother output, as at point B in Figure 4. The larger the capacitor, the more energy it stores and releases, and the smoother the output becomes.

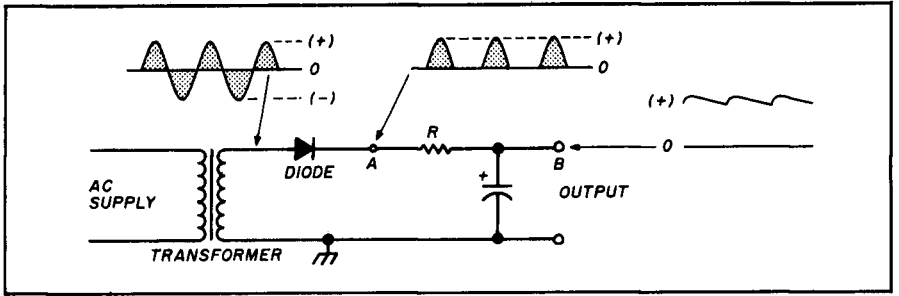
It usually requires a very large value to smooth the DC output completely, so some power supplies use two capacitors in a dual section filter (see Figure 5A). In this case, the first capacitor does a fair job of smoothing out the "ripples." The current then flows through the resistor to the second capacitor, further smoothing the DC until it reaches a nearly "pure" DC state. In some power supplies, the resistor is replaced by an iron core inductance called a "choke" (see Figure 5B). This choke helps by storing and releasing energy in its electromagnetic field. (You haven't heard the last of this magnetic field business. Just wait until I explore how alternating current, or AC, behaves in various circuits in a future issue!)

High voltage supplies, like those required by vacuum tube amplifiers, often use a capacitor-choke-capacitor filter to keep the physical size of the required capacitor to a minimum. A 500- μ F capacitor that works at 15 volts is often only an inch or so long and a half inch or less in diameter. The same capacitance that would work at 1000 volts or so would be several inches tall and probably several inches wide.

What is a "coupling" capacitor?

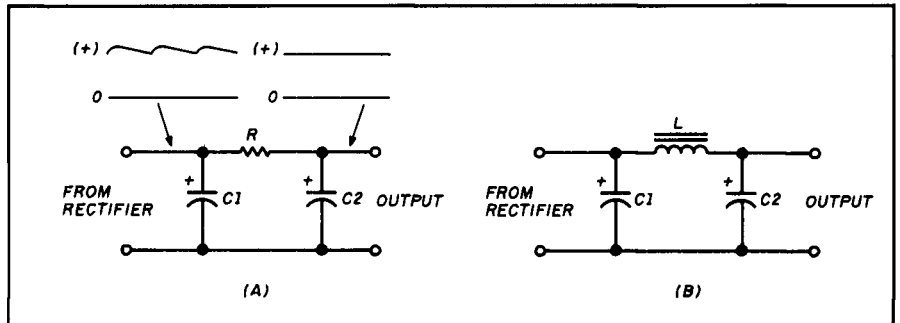
A "coupling" capacitor is obviously used to couple something. If a transformer is used between the first amplifier and the second one in a simple audio circuit, it is said to be "transformer coupled." If a capacitor is used,

FIGURE 4



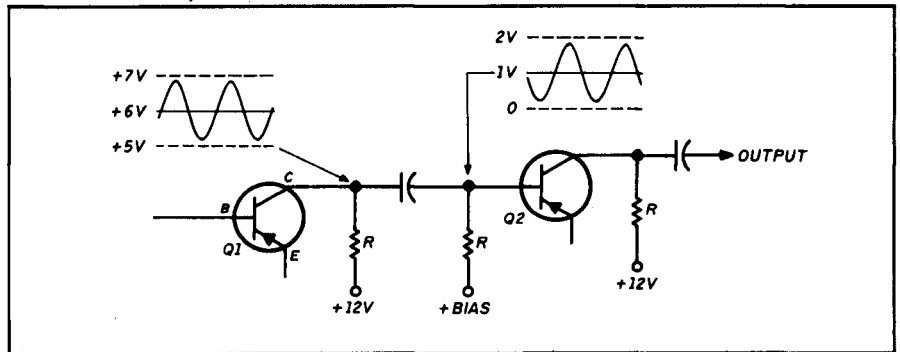
Capacitors also smooth out the rectified AC from a diode in a power supply, making it useful to power other circuitry.

FIGURE 5



Adding a second capacitor and a resistor to the power supply output further smooths the DC output (A). Placing an iron core inductance (filter choke) in the circuit helps smooth the DC output and allows smaller value capacitors to be used (B).

FIGURE 6



Capacitive coupling between two transistor amplifiers. A replica of the audio signal at Q1 is impressed on the bias voltage of Q2 while keeping the two DC voltages separated.

the circuit is "capacitance coupled" (see Figure 6).

When a voltage change appears on one plate, an equal and opposite change appears on the opposite plate. That describes capacitive coupling. Let's follow what happens in Figure 6 when a signal (audio sine wave) appears at the collector of the first transistor (Q1). As the voltage starts to go positive on the collector, the same thing is happening on one plate of C1 (the capacitor in the center of the drawing).

As the electrons increase on one plate, the electrons on the other plate flow back to the power supply through the ground or chassis return. This causes an inverted replica of the signal to appear on the base of Q2. The signals aren't exact opposites of each other because there's a slight lag, called phase shift, through a capacitor. For purposes of discussion, we'll ignore this lag.

When the voltage on C1 (Q1 collector) passes its peak, drops to zero, and

starts going in the opposite direction (negative), the voltage on the base of Q2 follows — producing a replica of the signal on the base of Q2. The transistor then amplifies the signal, and so on. There's an important secondary capacitor function here which you must not overlook. The DC operating voltage on the collector of Q1 is prevented from getting to the base of Q2. This is vital because Q2 needs only a small fraction of a volt on its base to work properly; too much voltage here would destroy the transistor. Because the capacitor passes the *changes* in voltage along, but not the *steady state* DC voltage, the DC bias on the base stays where it belongs and all is well. Thus, you can say that the signal is "coupled" from the first stage (Q1) to the second stage (Q2) by the coupling capacitor (C1). This principle works in the same way for audio, IF, and RF signals. The major difference is that C1 must be larger for audio (typically 0.05 to 5 μF) and smaller for RF (2 pF or less to perhaps 500 pF), depending upon the frequency being coupled. The capacitance required is related to the reactance and impedance in an AC circuit. (I'll discuss this in an upcoming column on alternating current.)

Series and parallel capacitors

Just as you can use resistors connected in series or parallel to obtain a needed value, you can work with capacitors. However, there is a slight difference. The formulas for capacitance are reversed from those for resistors. Capacitance values in series produce *smaller* values; those in parallel produce *larger* values. Figure 7 gives some examples.

The formula for series capacitors is:

$$C(\text{total}) = 1/((1/C1)+(1/C2)+(1/C3)),$$

and so on. Here's a tip. In a series hookup, the total capacitance will always be smaller than the value of the smallest capacitor. Use the formula to calculate how much.

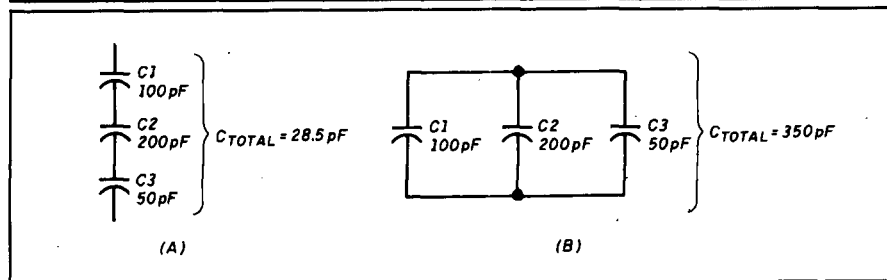
For parallel capacitors, the formula involves straight addition:

$$C(\text{total}) = C1 + C2 + C3, \text{ and so on.}$$

The experiment

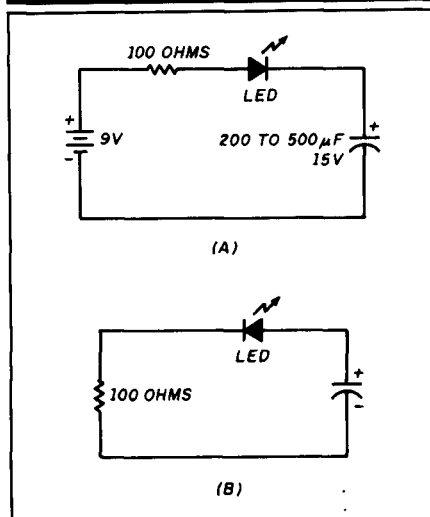
Here's a simple experiment you can perform to see how a difference in

FIGURE 7



Capacitors can be hooked in series or parallel to obtain unusual values. See text for the formulas used.

FIGURE 8



Note the time required to charge the capacitor through the resistor and LED at A, then reverse the diode and note the discharge time. You can try different values of capacitance to see how the time varies for charging and discharging.

capacitance value changes the amount of energy needed to charge a capacitor, and how long it will take to release that energy to an external circuit. The hookup is shown in Figure 8. It uses an LED, a resistor (to limit current through the LED to a safe value), and a 9-volt battery. Try some different capacitor values; many grab bags will contain several. Start with a large one — say 400 to 500 μF rated at 15 or 20 volts.

First connect the battery, resistor, and LED together and touch one leg of the LED to the opposite side of the battery. If it lights, you're all set; if it doesn't, reverse the LED wires and try again. After you've proved that the hookup works, connect the negative (–) lead of the capacitor to the negative side of the battery, and connect the resistor and LED between the battery

and the positive (+) lead of the capacitor as shown in Figure 8A. It should glow for 2 or 3 seconds, then gradually fade out. The capacitor is now charged — it won't accept any more electrons. Now, disconnect the battery, reverse the LED wires, and connect the resistor/LED combination across the capacitor while watching closely (Figure 8B). The LED will glow briefly as the energy stored in the capacitor flows through the LED and resistor until the charge is dissipated.

Do the same experiment with smaller values. Try a 5 or 10- μF value and note the difference in time. Connect several large ones in parallel and see how long they take to charge and discharge. To test the quality of the dielectric and insulation, charge a capacitor up and then wait several minutes before connecting the LED. If the capacitor makes the diode blink after 5 minutes, see if it still works after 10 or more minutes. Some capacitors have a very high quality dielectric and will store a charge for an hour or longer; others last only a couple of minutes.

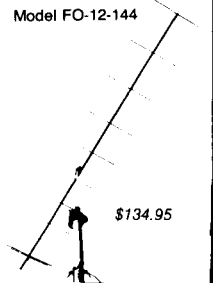
What does the voltage rating mean?

You'll note that electrolytic capacitors have a voltage listed along with the capacitance. Unless otherwise stated, this is the maximum working DC voltage for which that type of capacitor is designed. You'll see many electrolytic capacitors for transistorized circuits rated for 6, 12, 15, 20, or 35 volts. You can always use less than the rated voltage without a problem. It's bad practice to use a capacitor right at its rated voltage because a voltage surge will short the unit sooner or later. It's a definite "no no" to use a voltage higher than the rating.

Nonelectrolytic capacitors, like disk ceramic, dipped mica, or molded

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Model FO-12-144



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ELECTRICAL SPECIFICATIONS:

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E-Plane beamwidth 34 deg
H-Plane beamwidth 37 deg
Sidelobe attenuation
1st E-Plane -18 dB
1st H-Plane -15 dB
SWR 1.13 (typical)
F/B ratio 22 dB
Impedance 50 ohm

MECHANICAL SPECIFICATIONS:

Length 17ft. 4in
Boom 1.375" 6061 T-6 Aluminum
Elements 1/4" Aluminum rod
Wind survival 120 + MPH
Mast up to 2" diameter
Element Insulators Black Delrin
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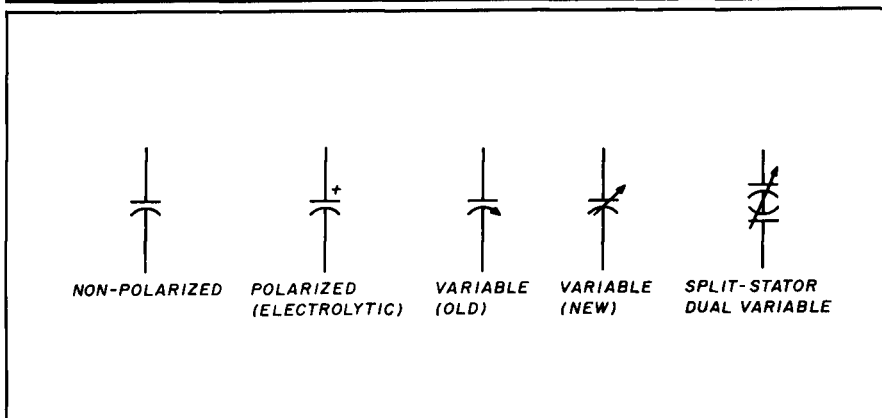
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FIGURE 9

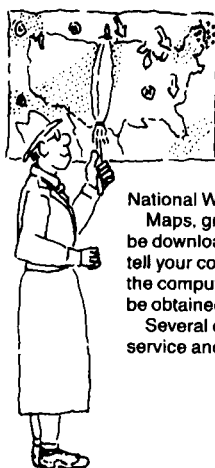


Schematic symbols for various types of capacitors.

paper have voltage ratings too. These units differ from electrolytic capacitors in that they can be reversed in the circuit; that is, they are not polarized.

Many nonpolarized capacitors have a black band around one end of the body, or have one end that is entirely darkened or colored to differentiate it from the other. This was called the "outside foil" in older molded paper capacitors. These capacitors were made by rolling up two strips of aluminum foil with a thin sheet of paper (often waxed paper) between. One foil was connected to one lead; the other was connected to the opposite lead. The layer of foil that ended up on the outside of the roll was designated as the layer to tie to ground because it helped shield

the circuit from unwanted signal or noise pickup. If the circuit didn't go to ground, the outside foil was connected to the lower impedance or lower voltage part of the circuit. In modern circuits, even though most capacitors aren't made of rolled foil, there are still reasons to follow convention and designate one side as the "ground" or low impedance connection. In schematic diagrams, you'll notice that a capacitor symbol has one straight bar and one curved bar (see Figure 9). The curved bar is the outside foil side in nonelectrolytic capacitors and the negative side in electrolytic ones. Electrolytic units are further identified by a plus (+) sign near the straight bar in a diagram. **ET**



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Transmitter kit LB-4 **\$8.95**
Receiver kit LB-5 **\$9.95**

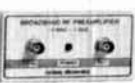
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CT-50	5 Hz-600 MHz	LESS THAN 25 mV < 25mV @ 50 MHz < 10mV @ 500 MHz < 100mV @ 800 MHz	1 PPM	8	1Hz, 10Hz	189.95
CT-125	10 Hz-1.25 GHz	< 50mV to 150 MHz < 10mV to 800 MHz	1 PPM	9	0.1Hz, 1Hz, 10Hz	189.95
CT-70 WITH OP-1 OPTION	10 Hz-600 MHz	< 10mV to 150 MHz < 150mV to 800 MHz	0.1 PPM	9	0.1Hz, 1Hz, 10Hz	229.95

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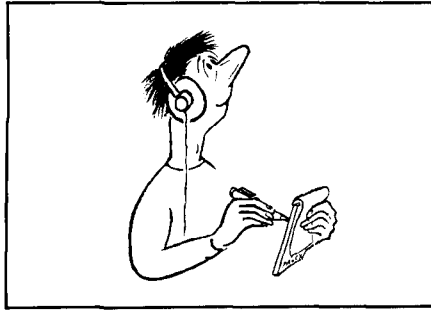
By Garth Stonehocker, KØRYW

RECEIVING DX BEACONS

Last month I presented beacon fundamentals, including beacon criteria and the frequency ranges of established beacons. As I mentioned, there are other radio stations that meet enough of the criteria to be useful as beacons. Among the stations that qualify are those owned and/or operated by Amateurs, broadcasters, researchers, and governments. Two examples of such stations are the Amateur network, PROPNET, on 14.1 MHz from worldwide locations and the standard frequency and time radio stations (also worldwide), mainly on 5, 10, and 15 MHz. There are also several useful stations on other frequencies. The best beacons would be from continuously on-the-air radio stations, identifiable not only by frequency but by voice or CW modulation, transmitting on several frequencies spaced across the HF frequency range (perhaps into the MF and low VHF ranges) and, of course, at locations known worldwide. Other information about the station like power and antenna (type and beam direction) would also be a plus.

Here's a method that will let you receive and use the ultimate beacon station, or those of lesser attributes. RadioShack carries a crystal-controlled receiver for 5, 10, and 15 MHz plus VHF for NOAA weather; the price is \$39.95. It doesn't have an RF input stage, so it's limited in sensitivity. Marring it with an active antenna fixes the sensitivity problem and provides an omnidirectional indoor antenna to boot, or a connection for an external antenna if necessary. One such active antenna, the MFJ-1020A, is available from MFJ Enterprises, Box 494, Mississippi State, Mississippi 39762 for \$79.95. Similar active antenna units have been featured in Amateur Radio magazines as homebrew weekend projects.^{1,2} Because no metering is involved, this unit is only used for listening.

A high-tech beacon receiving setup can use a programmable receiver like the Kenwood R-5000, then do an analog to digital (A/D) conversion of its auto-



matic gain control voltage with an A/D board in the controlling microcomputer/PC. Software in the PC can execute control of the R-5000 at selected frequencies and times, control interrupts for the A/D functions, and manage the digital A/D data in storage and graphs. The graph would show the stored data and the latest values for a running plot of signal strengths at the monitored frequencies versus time. An automated system like this leaves you with little to do but view the graph for the beacon information you need.

There are, of course, all sorts of intermediate receiving setups for single or multiple frequencies. A hands-on application of a general coverage receiver (digital frequency dial and good image rejection would be advantages) using its S-meter for signal strength data values is one compromise. Another way to create a single frequency dedicated beacon receiver is by homebrewing. You can make an HF receiver/converter using a \$3 pc board without the loopstick from a transistor AM radio with a three-transistor RF-mixer crystal oscillator. The converter circuit has been published in Amateur magazines and the *ARRL Handbook*. Receiving beacon signals for propagation information can be an interesting part of DXing. Why don't you give it a try?

Last-minute forecast

The first two weeks of February are expected to have lower solar flux levels so the ionospheric MUF may be down, favoring the lower frequency bands. Signal strengths should be high and distortion from multipath lower. Thunderstorm noise is almost nonexistent and geomagnetic disturbances should be limited to the third and fourth weeks, possibly around the 16th and 23rd. These weeks are expected to exhibit higher solar flux levels, causing


absorption and MUFs favoring the higher frequency bands with longer openings. No significant meteor showers are scheduled to appear in February. A full moon will occur on the 10th; perigee will be on the 7th.

Band-by-band summary

Ten and 12 meters, the highest day-only DX bands, are nearest (although somewhat below) the MUF on Southern Hemisphere paths. They'll be open most days when the solar flux is above 200 during the 10 to 14-hour period centered on local noon. These bands open on paths toward the east and close toward the west. The paths are up to 4000 km (2400 miles) in single hop length and, on occasion, double that during evening transequatorial openings.

Fifteen meters, a day-only DX band open most of each day, has lower signal strengths and greater multipath variability than 10 and 12 meters. It will be best when the MUF is just resting above this band and will remain so until it drops below the band — a transition period that occurs right after sunrise and just before sunset. Transequatorial openings will occur, with distances similar to 10 and 12 meters.

Seventeen, 20, and 30 meters are both day and nighttime DX bands. Seventeen is the maximum usable band for DX in the northern directions during daytime. In combination with 20/30 meters, it provides nighttime paths for the day-only bands. Thirty meters becomes the main over-the-pole daytime band, with some hours covered by 17 and 15 meters. This path may be affected by anomalous absorption on a few days of the month.

Forty, 80, and 160 meters, the night-only DX bands, exhibit short skip during daylight hours during the lowest solar flux, then lengthen at dusk. They are always far below the MUFs, except during disturbances on northern and east-west paths. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas for a few hours before dawn. 

REFERENCES

1. R.W. Burhans, "Active Antenna Preamplifiers," *Ham Radio*, May 1986, page 47.
2. Peter Bertini, K1ZJH, "Active Antenna covers 0.5-30 MHz" *Ham Radio*, May 1985, page 37.

WESTERN USA

GMT	PST	N	NE	E	SE	S	SW	W	NW
0000	4:00	17	30	12	10	10	10	10	12
0100	5:00	15	30	15	10	10	10	10	12
0200	6:00	15	30	15	10	10	10	10	12
0300	7:00	17	30	15	12*	10	10	10	15
0400	8:00	20	30	17	12	10	10	10	17
0500	9:00	20	30	17	15	10	10	12	20
0600	10:00	30	30	17	15	12	12	12	20
0700	11:00	30	30	17	15	12	12	15	20
0800	12:00	30	30	17	17	15	15	15	30
0900	1:00	30	30	17	17	15	15	15	30
1000	2:00	30	30	17	17	17	15	17	30
1100	3:00	30	30	20	17	17	17	17	30
1200	4:00	30	30	20	17	17	17	17	30
1300	5:00	30	20	10	12	20	17	17	30
1400	6:00	30	15	10	10	20	17	17	30
1500	7:00	30	15	10	10	10	17	15	30
1600	8:00	30	15	10	10	10	17	15	30
1700	9:00	30	15	10	10	10	15	15	30
1800	10:00	30	15	10	10	10	12	17	30
1900	11:00	30	17	10	10	10	10	15	30
2000	12:00	30	20	10	10	10	10	12	15
2100	1:00	30	20	10	10	10	10	10	12
2200	2:00	30	20	10	10	10	10	10	12
2300	3:00	30	30	12	10	10	10	10	12

FEBRUARY

MID USA

GMT	MST	N	NE	E	SE	S	SW	W	NW
0000	5:00	30	30	15	10	10	10	10	12
0100	6:00	30	30	15	10	10	10	10	15
0200	7:00	30	30	15	10	10	10	10	17
0300	8:00	30	30	15	12	10	10	10	20
0400	9:00	30	30	17	12	12	12	15	20
0500	10:00	30	30	17	15	12	12	15	20
0600	11:00	30	30	17	15	15	15	15	30
0700	12:00	17	30	17	17	15	15	17	30
0800	1:00	15	30	17	17	15	15	17	30
0900	2:00	15	30	17	17	17	17	17	30
1000	3:00	17	30	17	17	17	17	17	30
1100	4:00	20	30	20	17	17	17	17	30
1200	5:00	20	20	10	12	17	17	17	30
1300	6:00	20	15	10	10	20	17	17	30
1400	7:00	30	15	10	10	15	17	15	30
1500	8:00	30	15	10	10	10	15	15	30
1600	9:00	30	12	10	10	10	15	15	30
1700	10:00	30	12	10	10	10	12	17	30
1800	11:00	30	15	10	10	10	10	12	30
1900	12:00	30	17	10	10	10	10	10	17
2000	1:00	30	20	10	10	10	10	10	15
2100	2:00	30	30	10	10	10	10	10	15
2200	3:00	30	30	12	10	10	10	10	15
2300	4:00	30	30	12	10	10	10	10	15

EASTERN USA

GMT	EST	N	NE	E	SE	S	SW	W	NW
0000	7:00	30	30	15	10	10	10	10	17
0100	8:00	20	30	15	10	10	10	10	20
0200	9:00	30	30	15	12	10	10	10	20
0300	10:00	30	30	17	12	12	12	15	20
0400	11:00	30	30	17	12	12	12	15	30
0500	12:00	30	30	17	15	15	15	17	30
0600	1:00	30	30	17	15	15	15	17	30
0700	2:00	30	30	17	17	15	15	17	30
0800	3:00	30	30	17	17	15	17	17	30
0900	4:00	30	30	17	17	17	17	17	30
1000	5:00	30	17	10	17	20	17	17	30
1100	6:00	17	15	10	12	17	17	17	30
1200	7:00	15	15	10	10	12	17	17	30
1300	8:00	17	12	10	10	12	15	15	30
1400	9:00	20	12	10	10	10	15	15	30
1500	10:00	20	12	10	10	10	15	15	30
1600	11:00	20	12	10	10	10	17	15	30
1700	12:00	30	15	10	10	10	15	15	30
1800	1:00	30	15	10	10	10	12	15	30
1900	2:00	30	15	10	10	10	10	12	30
2000	3:00	30	17	10	10	10	10	10	17
2100	4:00	30	20	10	10	10	10	10	15
2200	5:00	30	30	10	10	10	10	10	12
2300	6:00	30	30	12	10	10	10	10	15

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We stock a full line of Transistors & Tubes for amateur, marine, and business radio servicing



Partial Listing of Popular Transistors in Stock

P/N	Net/Ea	P/N	Net/Ea	P/N	Net/Ea
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CD2654A	26.50	PT6619	19.75	MGF1402	\$17.95
ECG340	3.40	PT9847	22.75	MRF901	1.50
MRF134	16.00	RF120	21.00	MRF911 & 966	2.50
MRF136	21.00	SD1229	12.00	NE25537/2SK205	3.25
MRF136Y	47.00	SD1272	12.00	NE41137/3SK124	3.25
MRF137	24.00	SD1278-1	13.75	U309 & U310	1.75
MRF138	35.00	SD1405	16.00	2N4416 & J310	1.00
MRF141G	190.00	SD1407	25.00	3N204 & 3N211	2.00
MRF148	34.00	SD1428	29.50	OUTPUT MODULES	
MRF150	79.50	SD1429-3	16.00	(Partial listing only - call for numbers not listed)	
MRF151G	179.50	SFR2072	12.75	SAU4 440 LIN	49.50
MRF153	395.00	SFR3662	29.50	SAV6 158	42.50
MRF156	537.00	SFR3775	13.00	SAV7 144	42.50
MRF171	34.50	SFR3800	17.50	SAV12 144 HT	23.50
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MRF174	80.00	2N3553	3.00	SAV17 144 50W	66.50
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MRF238	16.00	2N5179	1.00	M57737 144	48.50
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MRF247	24.75	2N5642	16.50	M57762 1296	69.75
MRF248	35.00	2N5643	19.00	M57764 606	74.00
MRF260	10.00	2N5944	11.00	M57712, M57733 use	
MRF261	10.50	2N5945	10.00	M57737, SC1019 SAV7	
MRF262	10.50	2N5946	12.50	SC1027 use SAU4	
MRF264	12.50	2N6080	9.00	MHW710-1,2,3	61.00
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MRF314A	33.00	2N6082, 3,4	14.75	MHW820-2	82.00
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MRF485MP	19.75	25C2290MP	39.50	7581/KT66	16.95
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MRF629	3.25	25C2630	23.00	8874	359.50
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MRF644	23.00	25C2642	28.25	3CX1200A7	469.00
MRF646	26.00	25C2694	46.75	3CX1500A7	699.50
MRF648	31.00	25C2695	31.75	3CX3000A7	719.95
MRF660	13.25	25C2782	32.75	4CX250B	99.95
MRF643/F	21.00	25C2879	21.00	4CX350A	199.50
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 - Dual Frequency Receive
 - Digital AF Filter • 100 Memories

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THE BEST OF THE BEST

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

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

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- 45W on 2M/35W on 70cm
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- Extended Receiver Range
- More Features for the Money Than Anyone Else

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TS-140S AFFORDABLE DX-ing!

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- Compact, Lots of Features

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- Optional 50 MHz, 220 MHz or 1.2 GHz
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- 10 Watts Output on 6 Meters and 1.2 GHz • 100 Memories

ICOM

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- USB/LSB/CW, AM Receive
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- Receive 30 kHz to 33 MHz
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- TS-140, Affordable HF
- TS-680, HF Plus 6 Meter
- TM-231, 2 Meter Mobile
- TM-731, Dualband, FM
- TM-701, 25W, 2M/440 MHz
- TH-75, 2M/70cm HT

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- RS50A \$199
- RS50M \$219
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- RS35A \$139
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MODEL	2210	1300H/A	2400H	CCA	CCB
RANGE: FROM TO	10 Hz 2.2 GHz	1 MHz 1.3 GHz	10 MHz 2.4 GHz	10 MHz 550 MHz	10 MHz 1.8 GHz
APPLICATIONS	General Purpose Audio-Microwave	RF	Microwave	Security	Security
PRICE	\$219	\$169	\$189	\$299	\$99
SENSITIVITY					
1 KHz	< 5 mv	NA	NA	NA	NA
100 MHz	< 3 mv	< 1 mv	< 3 mv	< .5 mv	< 5 mv
450 MHz	< 3 mv	< 5 mv	< 3 mv	< 1 mv	< 5 mv
850 MHz	< 3 mv	< 20 mv	< 5 mv	NA	< 5 mv
1.3 GHz	< 7 mv	< 100 mv	< 7 mv	NA	< 10 mv
2.2 GHz	< 30 mv	NA	< 30 mv	NA	< 30 mv

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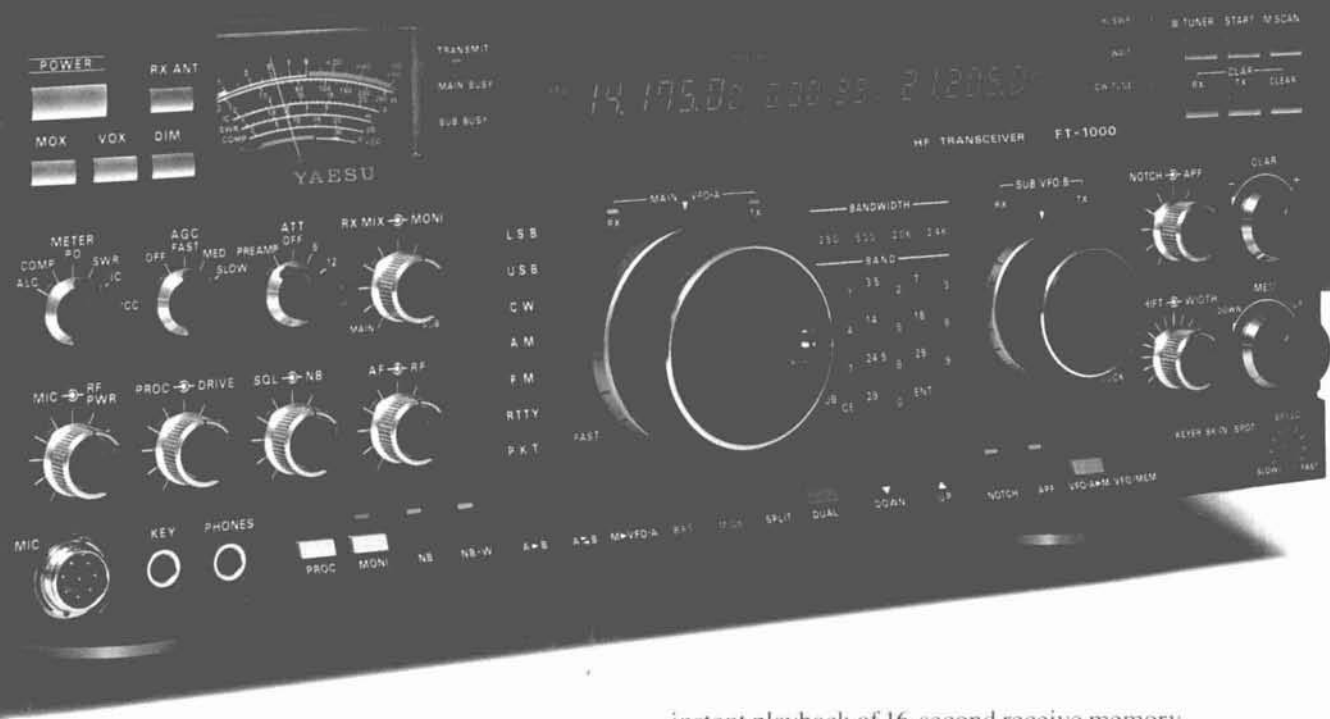
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- **Digital Voice Storage (DVS-2)** option provides

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- **Automatic Antenna Tuner** built-in with fast action and 39 memories for quick band changes.
- **QRM Rejection Systems**, including a variety of cascaded filter selections, width control, IF shift, IF notch filter, all-mode squelch, dual-mode noise blanker and a CW audio peaking filter.

Additional Features: 108dB dynamic range • front panel RX antenna selector • built-in electronic keyer module • stereo dual receive • flywheel effect on main and sub VFO tuning dials • twin frequency displays • CW spot.

A product of three years of intensive research and design. This HF rig will allow you to achieve a position of competitive dominance.

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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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