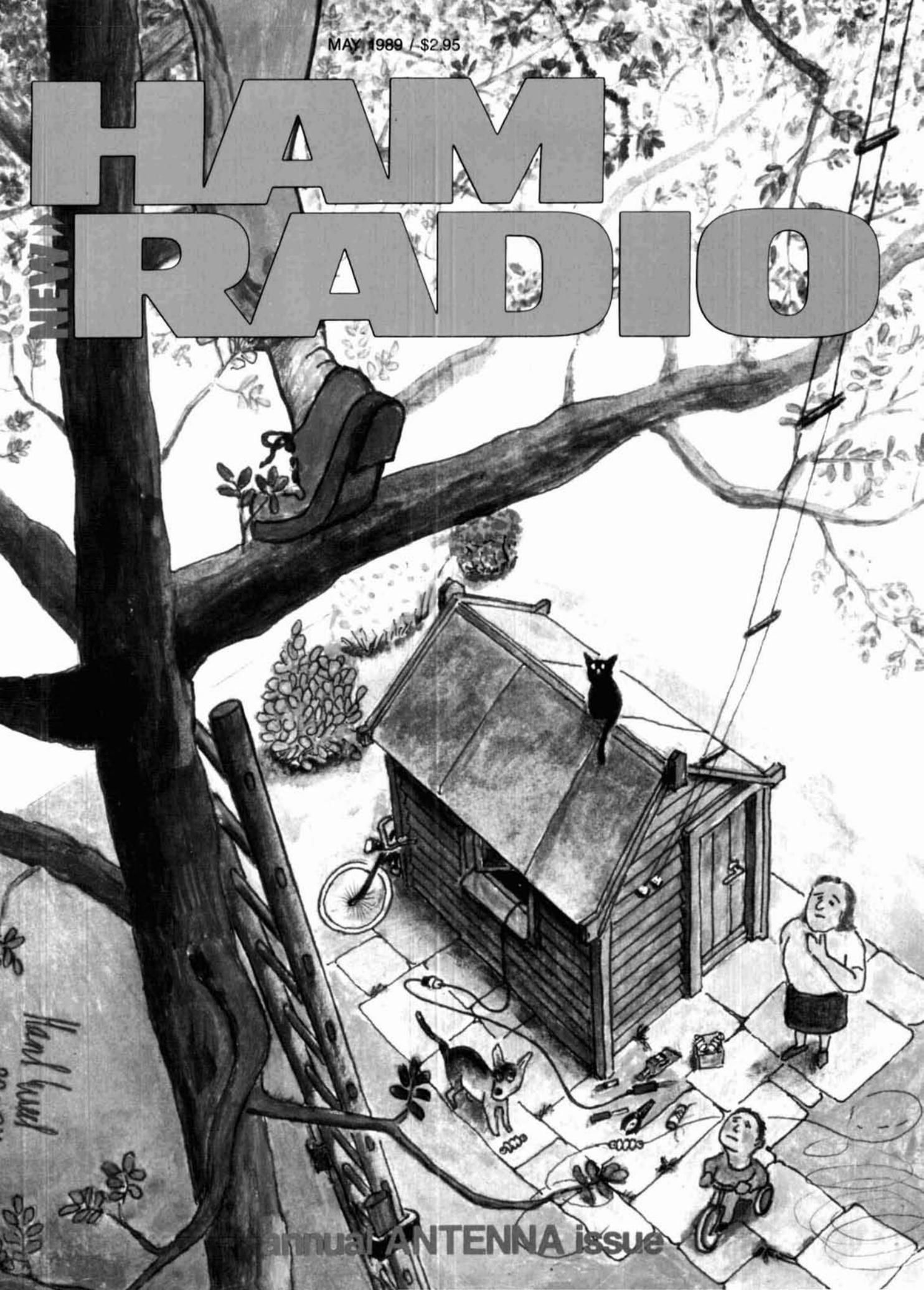


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Annual ANTENNA issue

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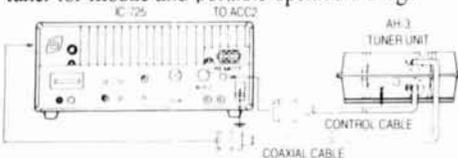
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Extraordinary Performance! Includes: 160 through 10 meter operation • 100 watts output • Shortwave reception from 100kHz to 33MHz • SSB, CW and AM modes (FM optional) • Sensitive 105db dynamic range receiver • Low noise DDS switching • Panel-selectable RF preamp and attenuator • Dual VFO's • Selectable AGC • Rugged full duty cycle finals.

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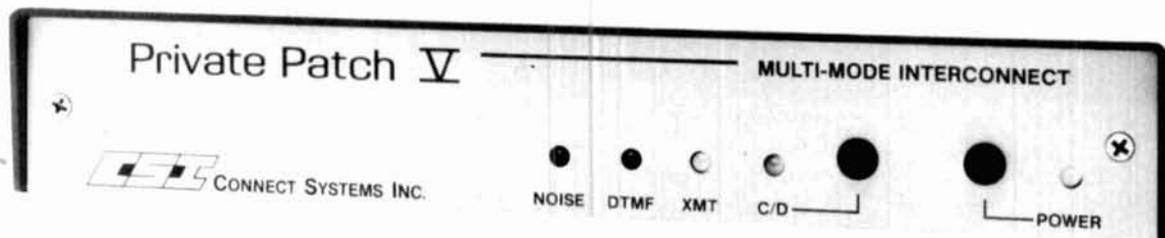
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113



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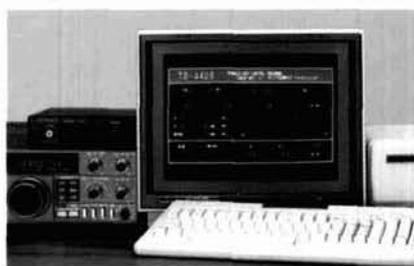
TS-440S Compact high performance HF transceiver with general coverage receiver

Kenwood's advanced digital know-how brings Amateurs world-wide "big-rig" performance in a compact package. We call it "Digital DX-citement"—that special feeling you get every time you turn the power on!

- **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.
- **Built-in automatic antenna tuner (optional)**
Covers 80-10 meters.
- **VS-1 voice synthesizer (optional)**



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- **Adjustable dial torque**
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Frequency and mode may be stored in 10 groups of 10 channels each. Split frequencies may be stored in 10 channels for repeater operation.
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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/PS-30 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-5/6/7 headphones
- SP-40/50B 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
- VS-1 voice synthesizer
- SW-100A/200A/2000 SWR/power meters
- TU-8 CTCSS tone unit
- PG-2S extra DC cable.

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

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HAM RADIO

MAY 1989

volume 22, number 5



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W6SAI, page 74

See page 25 for the winners of March's Weekender contest.

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William W. Eitel, W6UF
1908 - 1989

W6UF: Amateur Radio frontiersman

To say "It can't be done" was a challenge to William W. Eitel, W6UF. In 1927 it was thought that the new 10-meter band was workable for line-of-sight transmissions only, as the 5-meter band seemed to be. W6UF was one of the first stations on the band, determined to prove that 10-meters was good for long distance communication. In a few weeks he was running daily skeds with W1CCZ and ZL2AC, establishing beyond a doubt the DX qualities of the new band. His homemade transmitter used a crystal he had ground from a chunk of quartz found in a mountain stream. He saved money for weeks to buy an 852 tube for the 150-watt amplifier stage. He built a two-tube regenerative receiver. His efforts were summarized in *QST* in January 1929. Ten meters was not like the "ultra-high" frequencies! It was a DX band!

This was an auspicious beginning for a 20-year-old lad with a consuming curiosity about radio communications. His biggest problem was the cranky 852 tube; it required enormous plate voltage to work properly at 10 meters. He decided he could build a better tube, which would work at a reasonable plate potential.

His chance came in 1933 when he went to work for a vacuum tube manufacturer in the San Francisco area. Bill, along with Jack McCullough, W6CHE, developed a low-voltage, high-current tube that proved superior

to the 852. Unfortunately, the company (a marine communication business) wasn't interested in selling tubes to Amateurs, so tube sales languished. Bill and Jack soon left the company, and in 1934 started a new enterprise — Eitel-McCullough, Inc. Their goal was to build reliable "EIMAC" tubes that would operate at higher frequencies than anything available. They borrowed \$5,000 and designed a revolutionary new triode tube, the 150T.

This was the start of something big. EIMAC tubes were quickly adapted for commercial and military use, and the little company prospered. Within a decade it became the United States' leading producer of power electron tubes and related devices. The combination of Jack (the planner and designer) and Bill (the hands-on activist) was fortuitous. The right guys with the right products at the right time!

Bill's interest in Amateur Radio, although curtailed by the effort of building a company and the demands of war production, never flagged. Bill and Jack went out of their way to enlist Radio Amateurs in the company, to encourage them in the new communication industry, and to develop new tubes for Amateur Radio.

Over the years, Bill never lost his inventor's curiosity. His interest in advancing the frontiers of Amateur Radio continued. In 1961 the EIMAC Radio Club, under the leadership of Bill, Jack, and Hank Brown (W6HB), established the first Amateur two-way "moonbounce" contact on 1296 MHz with W1FZJ.

Bill was a member of the Northern California DX Club and Project OSCAR. He donated time, equipment, and money to make the early OSCAR satellites successful. If there was a job to be done, he'd do it. His enthusiasm and support were often all that kept the early OSCAR satellite program from foundering. He was a remarkable, enthusiastic leader in the best sense of the word. He was an excellent operator, CW or phone, and set the pace for Amateurs many years his junior.

Upon retirement, Bill moved to Dayton, Nevada and set up his own experimental laboratory, continuing to work on ideas that interested him. He was a life member of the ARRL, Project OSCAR, and the 5-Star Operator's Club. He was elected to a fellowship in the Radio Club of America.

He passed away in February 1989 at the age of 81. His accomplishments were many. In addition to being an active Radio Amateur, inventor, company founder, and executive, he was a good companion. He left behind a multitude of friends who mourned his passing. He left his mark in electronics and the Amateur world. His discontent with the status quo drove him to succeed when others dropped by the wayside. His inquisitive mind made him an alert problem solver. Along the way he helped others. He was an American original: a self-educated small-town boy who grew up in a turbulent era of rapid and productive scientific growth — and mastered his world.

We will all miss him. 73, Bill, and SK.

Bill Orr, W6SAI

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All New!

Stacked in Your Favor!

TM-231A/431A/531A

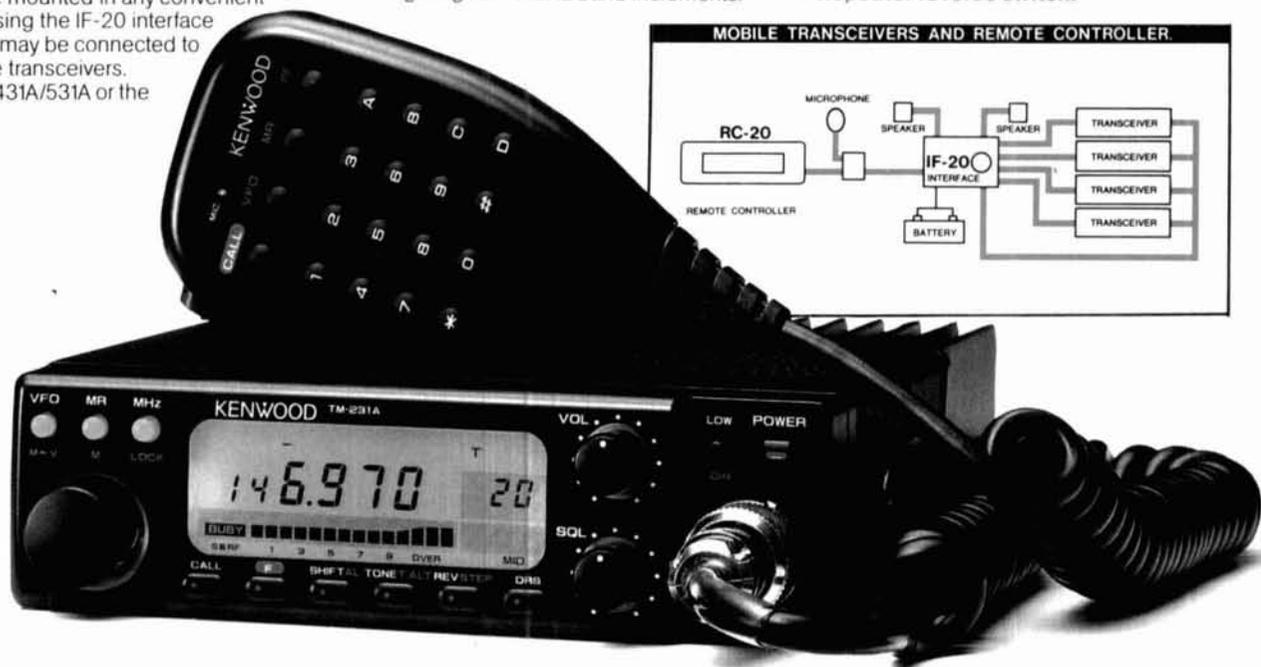
FM Mobile Transceiver

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

- 20 multi-function memory channels. 20 memory channels allow storage of frequency, repeater offset, CTCSS frequency, frequency step, Tone On/Off status, CTCSS and REV.
- High performance—high power! 50W (TM-231A), 35W (TM-431A) with a 3 position power switch (high, medium, low).
- Optional full-function remote controller (RC-20). A full-function remote controller using the Kenwood bus line, model RC-20, may be easily connected to the TM-231A/431A/531A and can be mounted in any convenient location. Using the IF-20 interface the RC-20 may be connected to four mobile transceivers. (TM-231A/431A/531A or the TM-701A)

- Multi-function DTMF mic. supplied. Controls are provided on the microphone for CALL (Call Channel), VFO, MR (Memory Call or to change the memory channel) and a programmable function key. The programmable key can be used to control one of the following on the radio: MHz, T. ALT. TONE, REV, DRS, LOW or MONITOR.
- Easy-to-operate illuminated keys. A functionally designed control panel with backlit keys increases the convenience and ease of operation during night-time use.
- Auto repeater offset on 144 and 220 MHz.
- Built-in digital VFO.
 - a) Selection of the frequency step (5, 10, 15, 20, 12.5, 25kHz)
 - *TM-531A: 10, 20, 12.5 25kHz
 - b) Programmable VFOThe user friendly programmable VFO allows the operator to select and program variable tuning ranges in 1 MHz band increments.

- Programmable call channel function. The call channel key allows instant recall of your most commonly used frequency data.
- Selectable CTCSS tone built-in.
- Tone alert system—for true "quiet monitoring"! When activated this function will cause a distinct beeper tone to be emitted from the transceiver for approximately 10 seconds to signal the presence of an incoming signal.
- Easy-to-operate multi-mode scanning. Band scan, Program band scan, Memory scan plus programmable memory channel lock-out, with time operated or carrier operated stop.
- Priority alert.
- DRS (Digital recording system). The optional DRU-1 can store received and transmitted messages for up to 32 seconds, allowing the operator to quickly check or return any call using the tone alert system.
- Automatic lock tuning function (TM-531A).
- Repeater reverse switch.



Optional Accessories

- RC-20 Full-function remote controller
- RC-10 Multi-function remote controller
- IF-20 Interface unit handset
- DRU-1 Digital recording unit
- MC-44 Multi-function hand mic.
- MC-44DM Multi-function hand mic. with auto-patch
- MC-48B 16-key DTMF hand mic.
- MC-55 8-pin mobile mic.
- MC-60A/80/85 Desk-top mics.
- MA-700

- Dual band (2m/70cm) mobile antenna (mount not supplied)
- SP-41 Compact mobile speaker
- SP-50B Mobile speaker
- PS-430 Power supply
- PS-50 Heavy-duty power supply
- MB-201 Mobile mount
- PG-2N Power cable
- PG-3B DC line noise filter
- PG-4H Interface connecting cable
- PG-4J Extension cable kit
- TSU-6 CTCSS unit

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

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PUBLISHER'S LOG

You are now looking at the final step in our redesign program here at *Ham Radio*. It represents well over a year of very hard work on the part of a large number of people including many of you, our readers.

We started with a new editorial staff. Next, we interviewed a sizable sample of both subscribers and non-subscribing Amateurs to find out where you felt our strengths and weaknesses were. What should be done to make the best Amateur Radio magazine even better?

By September of last year you saw phase one, with our new logo and a redirected emphasis on practical construction articles and shorter technical and tutorial pieces. We also instituted the feedback cards to find out just what you liked and didn't like about our many changes.

We listened, we fine tuned, and we listened again. All the while, those responsible for the design of the magazine were hard at work coming up with a product which would be more eye pleasing than ever before. One that would, at the same time, be very efficient in helping the reader to get the most from each page they read.

I'm very happy with this finished product, but I'm even more proud of the many people who have gone the extra distance to bring all of this together. Thanks must go to the whole *Ham Radio* staff; to those at Wallace Press, our typesetter; and to Anne Desmarais, our design consultant.

The formal program is now over, but the striving for improvement will never stop. If you have any good ideas or suggestions, we're always open to them. Let us know what you think.

Skip Tenney, W1NLB

Comments



A good laugh

Dear HR

Your February issue arrived in the mail yesterday and, after a look at the cover, I had one of the best laughs in a long time!

It's hard to tell what the piece of gear is that has your "cover ham" looking so apprehensive, but I immediately thought of the old SB-104 that is sitting on my table; the schematic is just about the same size, so big that I finally pinned it on the wall for convenience.

So congratulations to you and your artist for identifying with us readers who enjoy (?) digging into gear!

**Ray Burke, VE1BFG, Bathurst,
New Brunswick, Canada**

A good recipe— blending old with new

Dear HR

My good friend, Bert Cliff, W2QN, gave me a one year subscription to HR for Christmas. Today, for the first time since a lapse of almost ten years, I received the January issue in top condition.

I could not help but sit down and work through the magazine as soon as I could do so. If I am not mistaken, you are endeavoring to create a blending between the former *Ham Radio Horizons* and *Ham Radio*, which I followed for many years. My first impression is that you have achieved an excellent mixture. I, therefore, would suggest that you continue along this line, and I am eagerly looking forward to the following issues.

I have included my magazine evaluation card. I found the article written by Joe Reiser, W1JR, "VHF/UHF World," of special interest since it may well have put quite a dent into my HF one-sidedness.

Congratulations to Joe Carr, K4IPV, for his splendid article on "Writing the Technical Article." I was the last editor of the pre-war German DASD Amateur Magazine CQ in 1941/43. I do wish I could have read Joe's advise then, or at least in 1947 when I started my sideline career of writing pieces for our first ham mag QRV after the war.

**Albert Heine, DK7CN,
D-8990 Lindau, W. Germany**

Food for thought

Dear HR

I have just read a letter by AA6FW in your magazine of February 1989.

I am surprised. Perhaps the radio operators in San Diego are different from those here.

As an Amateur who is a relative newcomer (about 6 years), currently 49 years old, a V.E., and a volunteer operator at a museum demonstration station, I can make the following personal observations:

- I was stimulated to get my license by a man years my junior, not a grandfather.
- The growth of Amateur Radio, in this area, is positive and being aided by classes conducted by radio clubs in Delaware, Pennsylvania, and New Jersey.
- The newcomers are of all ages, and it's often a family affair.
- 5 WPM code can be passed by anyone who makes even a minimal effort. Of course code is an entrance requirement, but, so is a lot of other knowledge — frequencies, rules, regulations, electronics, etc.

Now, if the code requirement is eliminated under the guise of "too hard, scares them away," "not necessary," or some other feeble excuse, what do you suppose might be the next requirement to go — perhaps questions about...oh shucks, why even have an examination! "All an Amateur does is push a bunch of buttons." I can hear it now.

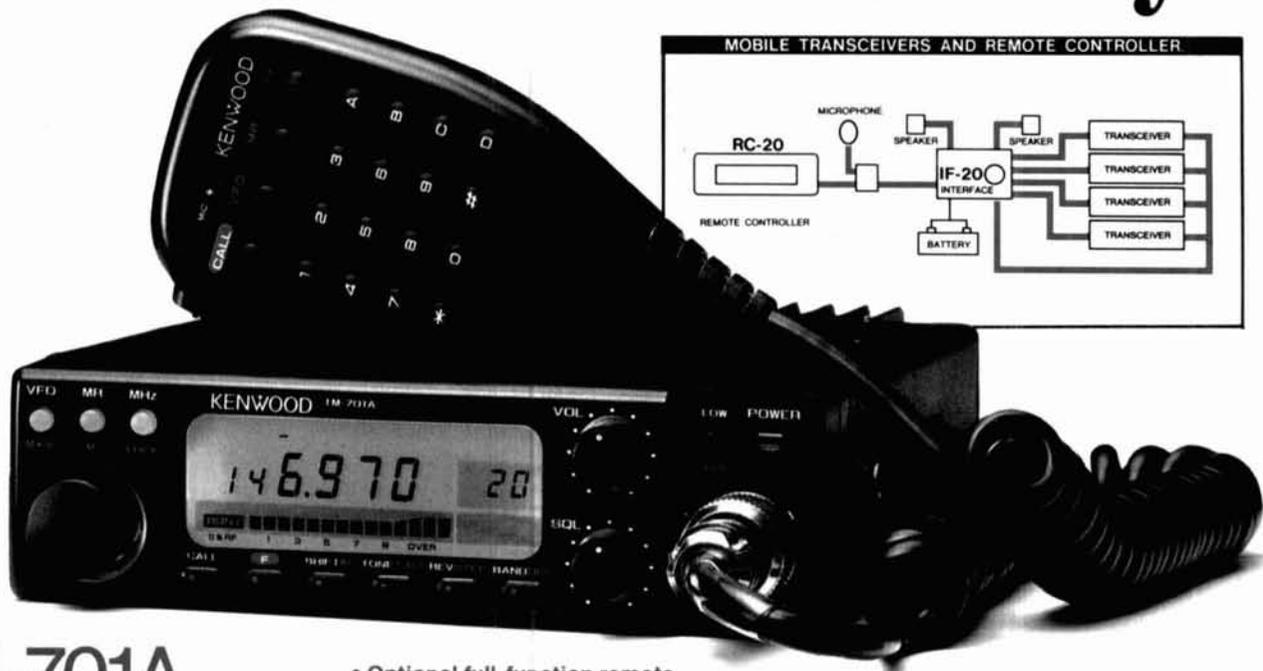
**Merrill Jay Mirman, D.O., KT3Z,
Springfield, Pennsylvania**

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Affordable
Breakthrough!

Dual Band Afford-ability!



TM-701A

Dual Bander

The TM-701A combines two radios into one compact package. You get 25 watts on 2 meters and 70cm, 20 memory channels, tone encoder built-in, multiple scanning, auto repeater offset selection on 2 meters, and a host of additional features!

- **20 multi-function memory channels.** 20 memory channels allow storage of frequency, repeater offset, CTCSS frequency, frequency step, and Tone On/Off status. CTCSS and REV, providing quick and easy access during mobile operation.
- **25W on 2m and 70cm.**
- **Selectable full duplex-cross band (Telephone style) operation.**
- **Easy-to-operate front panel layout.**
- **Multi-function DTMF mic. supplied.** Controls are provided on the microphone for CALL (Call Channel), VFO, MR (Memory Call or to change the memory channel) and a programmable function key. The programmable key can be used to control one of the following functions on the radio: MHz, T, ALT, TONE, REV, BAND, or LOW power.
- **Easy-to-operate illuminated keys.** A functionally designed control panel with individually backlit keys increases the convenience and ease of operation during night-time use.

- **Optional full-function remote controller (RC-20).**

A full-function remote controller using the Kenwood bus line may be easily connected to the TM-701A and mounted in any convenient location. The new controller is capable of operating all front panel functions.

- **Built-in dual digital VFO's.**

a) Frequency step selection (5, 10, 15, 20, 12.5, 25kHz)

b) Programmable VFO

The user friendly programmable VFOs allow the operator to select and program variable tuning ranges in 1MHz band increments.

- **Programmable call channel function.**

The call channel key allows instant recall of your most commonly used frequency data.

- **Programmable tone encoder built-in.**

- **Tone alert system - for true quiet monitoring.**

When activated this function will cause a distinct beeper tone to be emitted from the transceiver for approximately 10 seconds to signal the presence of an incoming signal.

- **Easy-to-operate multi-mode scanning.**

a) VFO scan

Band scan, Programmable band scan.

b) Memory scan plus programmable memory channel lock-out

c) Dual scan

Dual call channel scan
Dual memory scan
Dual VFO scan

d) Scan stop modes

Time operated scan (TO)
Carrier operated scan (CO)

e) Scan direction

f) Alert

When the AL switch is depressed memory channel 1 is scanned for activity at approximately 5 second intervals.

- MHz switch.
- Lock function.
- Repeater reverse switch.

Optional Accessories

- RC-20 Full-function remote controller
- RC-10 Multi-function remote controller
- IF-20 Interface unit handset
- MC-44 Multi-function hand mic.
- MC-44DM Multi-function hand mic. with auto-patch
- MC-48B 16-key DTMF hand mic.
- MC-55 8-pin mobile mic.
- MC-60A/80/85 Desk-top mics.
- MA-700 Dual band (2m/70cm) mobile antenna (mount not supplied)
- SP-41 Compact mobile speaker
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IMPROVED HIGH- PERFORMANCE YAGIS FOR 432 MHZ

Obtaining the most from a design

*By Steve Powlishen, K1FO, 816 Summer Hill Road,
Madison, Connecticut 06443*

In the July 1987 issue of *Ham Radio Magazine*,¹ I described a high-performance 432-MHz Yagi design which I built from a Cushcraft 424B. This design improved the radiation pattern and wet weather performance of the original and offered a substantial increase in forward gain.

I gave two versions of the design. The first used 24 elements on a 17' boom. K2OS, WA3FFC, and W7HAH now use this Yagi on EME. All of them have reported on-air EME performance improvements. A number of tropo operators have also been pleased with the results of my modification. But tropo performance is much harder to quantify and prove than EME performance. The Yagi works so well that several operators have chosen to build it from scratch. As it stands, the 24-element Mark 3 Yagi is still the final version for this boom length. I examined the possibilities of further optimization and found that I wouldn't achieve more than an additional 0.1 dB in theoretical gain. Unfortunately, I could obtain this gain increase only at the expense of pattern deterioration and increased resistive losses.

The second version of my design used a 24' extended boom and had 32 elements. (See Table 1 for element length and spacing dimensions.) This extended version has also been used successfully on EME, both in NC1I's 16-Yagi array and my 4-Yagi portable EME array. NC1I used the 4-

Yagi array in his Rhode Island and Vermont EME DXpeditions. WA9FWD has used a similar 4 × 32 element Yagi array on EME. (He's now replaced it with an even longer 36-element model.) In the July 1987 *Ham Radio* article,¹ I mentioned that while the 24-element design was a third-generation effort, the 32-element model I described was a second-generation effort which still had room for improvement. I also gave alternative, but untested, director lengths for a potentially improved 32-element Yagi and an even longer 38-element model (both third-generation designs).

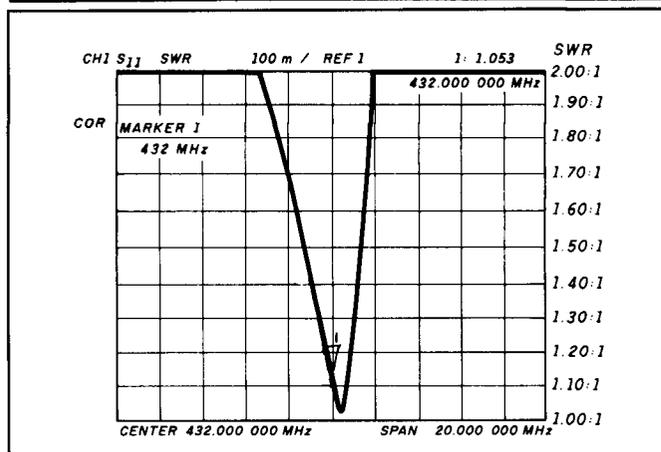
The revised design

Here are the final improved and tested dimensions for the 32-element Yagi. It's called the 32-element Mark 4 Yagi and is the fourth major revision of this design. The improvements to this 32-element Mark 4 Yagi over the earlier published version are:

- Improved radiation pattern, primarily in the rear lobes.
- Greatly improved VSWR bandwidth and wet weather performance.
- Reduced element resistive losses.
- Higher forward gain (~0.2 dB).
- Higher center frequency tuning for improved array performance.

I've also detailed a fully tested 27'6" long model.

FIGURE 1



Resonant frequency and VSWR bandwidth of the original 32-element Mark 2 Yagi.

Why improve it?

I got the impetus for these changes while helping to assemble NC11's 16-Yagi EME array. I had determined the driven element T match dimensions by testing a Yagi mounted on a pole in my back yard. When we started checking the driven element matches of the individual Yagis, mounted in place with the other Yagis in the 16-Yagi array, the dimensions for an acceptable VSWR didn't agree with my earlier work. A similar but lesser match problem arose in the 4-Yagi portable EME array. And, although wet weather performance of NC11's 16-Yagi array was greatly improved over the unmodified Yagis, the wet weather VSWR performance wasn't as good as we'd hoped.

I examined a sample 32-element Yagi on a Hewlett-Packard 8753A network analyzer; it revealed a very narrow match bandwidth. As you can see in **Figure 1** (a printout of the network analyzer measurement), the under 1.2:1 VSWR bandwidth is approximately 1 MHz. The driven element is under 2:1 VSWR over a 5-MHz span. This was substantially narrower than that displayed by the 24-element Mark 3 Yagi. My attempts to improve the match bandwidth on the 32-element Mark 2 Yagi by driven element adjustments alone were unsuccessful. With the help of the network analyzer I found a "natural" match frequency with acceptable bandwidth centered at 423 MHz. Previous work on the 32-element Yagi had shown that shortening all of the elements to raise the center frequency 9 MHz would lower the gain by several tenths of a dB.

I continued my computer analysis. As described in the *Ham Radio* article, the original computer analysis for the version 2 Yagi was done using a variation of the WB3BGU program. Because of the limited accuracy of this program, I had to make element adjustments to control the radiation pattern. Investigations done with MININEC showed excessive currents in the first few directors. These directors were also quite long when compared with some other designs. In fact, the director string could be divided into three parts: the first few directors which were tuned too low in frequency; the middle set of directors which were tuned too high in frequency; and the last directors which were tuned close to the correct frequency, but slightly low. Further analysis and work on other long Yagi designs² showed that a smooth

TABLE 1

Dimensions K1FO 24', 32-element Mark 4 Yagi.

Element Spacing (inches)	Element Length (mm)	Boom	Element
1.000	348		REF
5.250	336		DE
7.875	323		D1
11.563	314		D2
16.813	309	1"	D3
23.563	305		D4
31.875	301		D5
42.125	297		D6
52.375	294		D7
62.625	292		D8
72.875	290		D9
83.125	288		D10
93.375	286	1 1/8"	D11
103.625	285		D12
113.875	284		D13
124.125	283		D14
134.375	283	1 1/4"	D15
144.625	282		D16
154.875	280		D17
165.125	279		D18
175.375	278	1 1/8"	D19
185.625	277		D20
195.875	276		D21
206.125	276		D22
216.375	275		D23
226.625	274		D24
236.875	273		D25
247.125	273	1"	D26
257.375	272		D27
267.625	272		D28
277.875	271		D29
288.125	271		D30

minor lobe pattern coincided with a good current distribution. I used this information in the 24-element Mark 3 Yagi design.

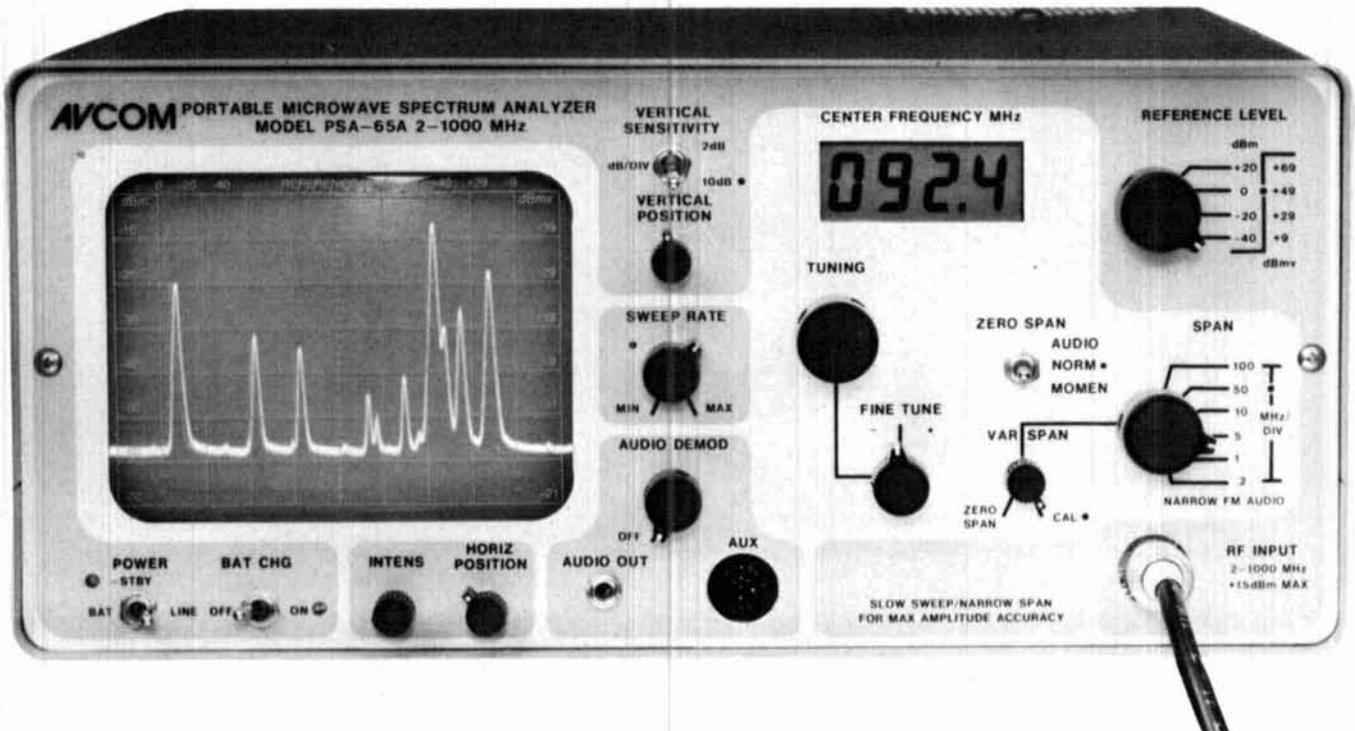
Design details

Up to this point, I'd been doing all my work on the Yagis in English dimensions. I needed an easier measurement method because I was spending considerable time building test Yagis. Metric dimensions for the element lengths were the answer. Not only is the millimeter an easier unit for working with 432-MHz Yagis, but the size of a millimeter (.039") allows for a smoother element taper — without the confusing fractional units.

All element spacings are the same as they were in the earlier versions; they are given in inches. I spent a fair amount of time looking at other spacing arrangements. It was possible to obtain slight performance improvements, but only with extensive spacing changes. These changes would make additional modification and upgrading difficult, defeating my purpose.

To make this improved design, I first converted the English lengths used in the 24-element Mark 3 Yagi to metric dimensions. Next I rounded these millimeter-sized directors to whole millimeters. Then I smoothed out the large changes in director lengths. The computer analysis on MININEC looked promising, but there was a substantial frequency

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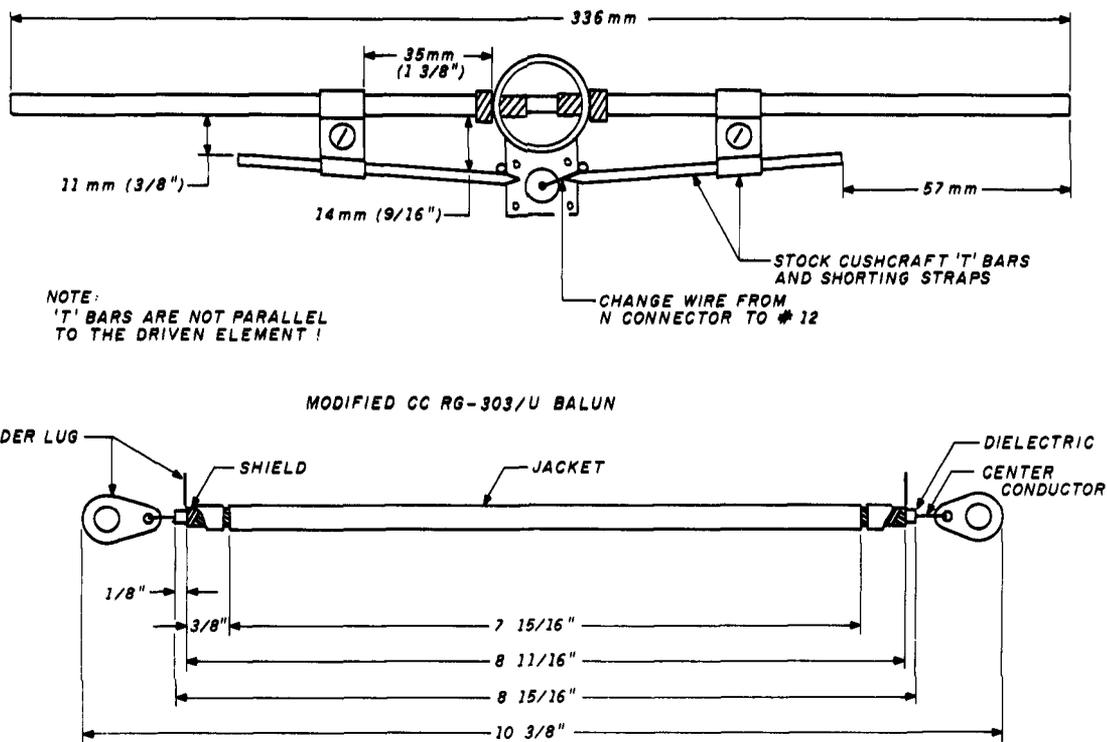
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FIGURE 2



Construction detail of the 32-element Mark 4 driven element using 424B parts.

shift as a result of these manual length changes. I shortened the lengths to center the Yagi at the desired frequency.

Now I had to perform the final length optimization using MININEC. This was a painfully slow process because of the computation time required by the computer available to me at that time. A quick look at the Yagi modeling with four segments required over 2 hours; a more accurate examination using eight segments took almost 8 hours. Consequently, I was able to make only a few element adjustments each day. I'd make an overnight high-accuracy run to ensure that I was still on the right track. Fortunately, the machine I use now solves the problem over six times faster.

When I started this project 4 years ago, I was using a simple program. The original design objective was to increase the Yagi's gain while creating a clean radiation pattern. My current design process adds to the original goals with the following requirements:

- Very high forward gain per boom length.
- Very clean radiation pattern.
- Wide gain bandwidth.
- Acceptable dry and wet weather performance.
- Good driven element match bandwidth.
- Reasonably high natural driven element impedance.
- Good director current distribution.
- Low resistive losses.

Knowledgeable use of the original program can get you 80 percent of the way to a good Yagi design. But the new requirements have rendered my first program obsolete, as it is unable to get all the way to an optimum solution. More complex programs, along with the all-important post-

computer optimization steps, now take you 95 percent of the way to the perfect Yagi.

After finalizing the computer-generated dimensions, I began to build and test the project. As I had already built and tested a number of Yagis, I needed to make adjustments only to the driven element and director 1.

Construction

The Yagi's mechanical layout is the same as originally described in the July 1987 *Ham Radio* article. Elements are mounted on plastic bushings which insulate them from the boom sections they extend through. The element ends are chamfered like those of the earlier Yagis. Supports keep the boom from sagging unacceptably. I suggest you review my earlier article before attempting to build these Yagis.

I compared driven element T matches constructed from the original Cushcraft parts used on the 24-element Yagi, with T matches using no. 12 T wires and a UT-141 balun like those of the 32-element Mark 2 Yagi. I obtained similar dry weather matches and match bandwidths with both driven element arrangements. Wet weather performance was slightly better with the no. 12 T wires. A slight adjustment to the balun length made it correspond to an electrical half wavelength. This improved the Yagi's pattern balance.

Figure 2 details the driven element construction using the original Cushcraft parts. Note that the balun must be shortened by 1 inch. As with the 24-element Yagi, I didn't use the original rectangular black spacers and I changed the jumper from the N connector to the T match bar to no. 12 to get a proper match. For the best match don't place the T bars parallel to the driven element.



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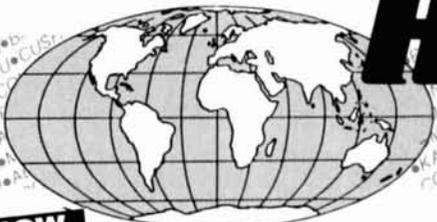
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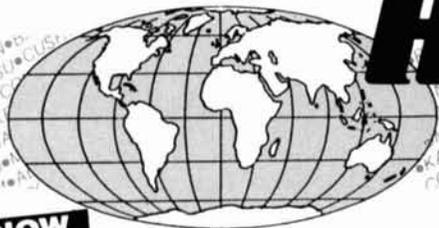
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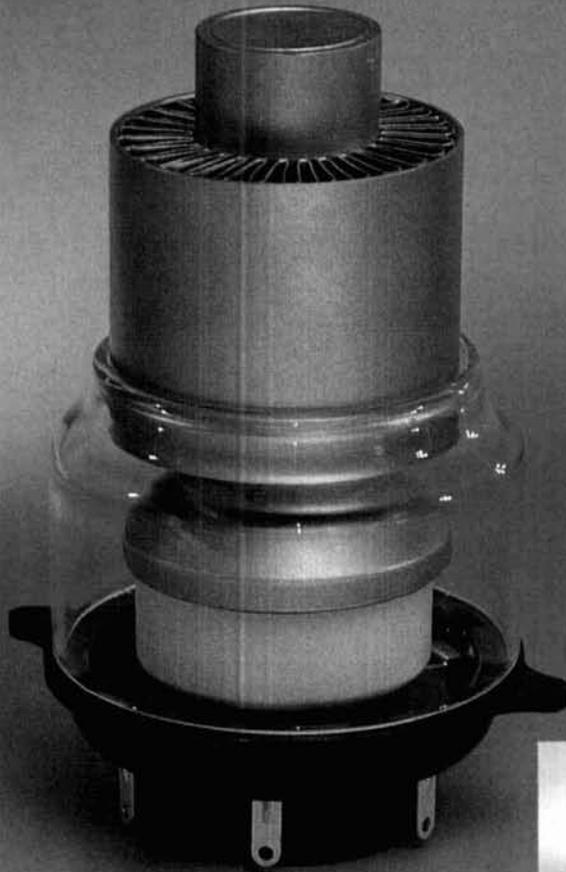
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The second version of the driven element is shown in **Figure 3**. It uses no. 12 T wires in place of the original 3/16" diameter T bars. This allows for a greater natural impedance setup. I prefer this arrangement because the shorting straps are farther out on the driven element, away from the high-current point. I tried baluns made from UT-141 solid copper shield coax and RG-303 (like the original 424B) with similar results. If you make the balun from RG-303 you'll need to use a set of dimensions different from those used with the first driven element arrangement. I thought it was desirable to eliminate the original solder lugs and solder the center conductor directly to the T match wires instead.

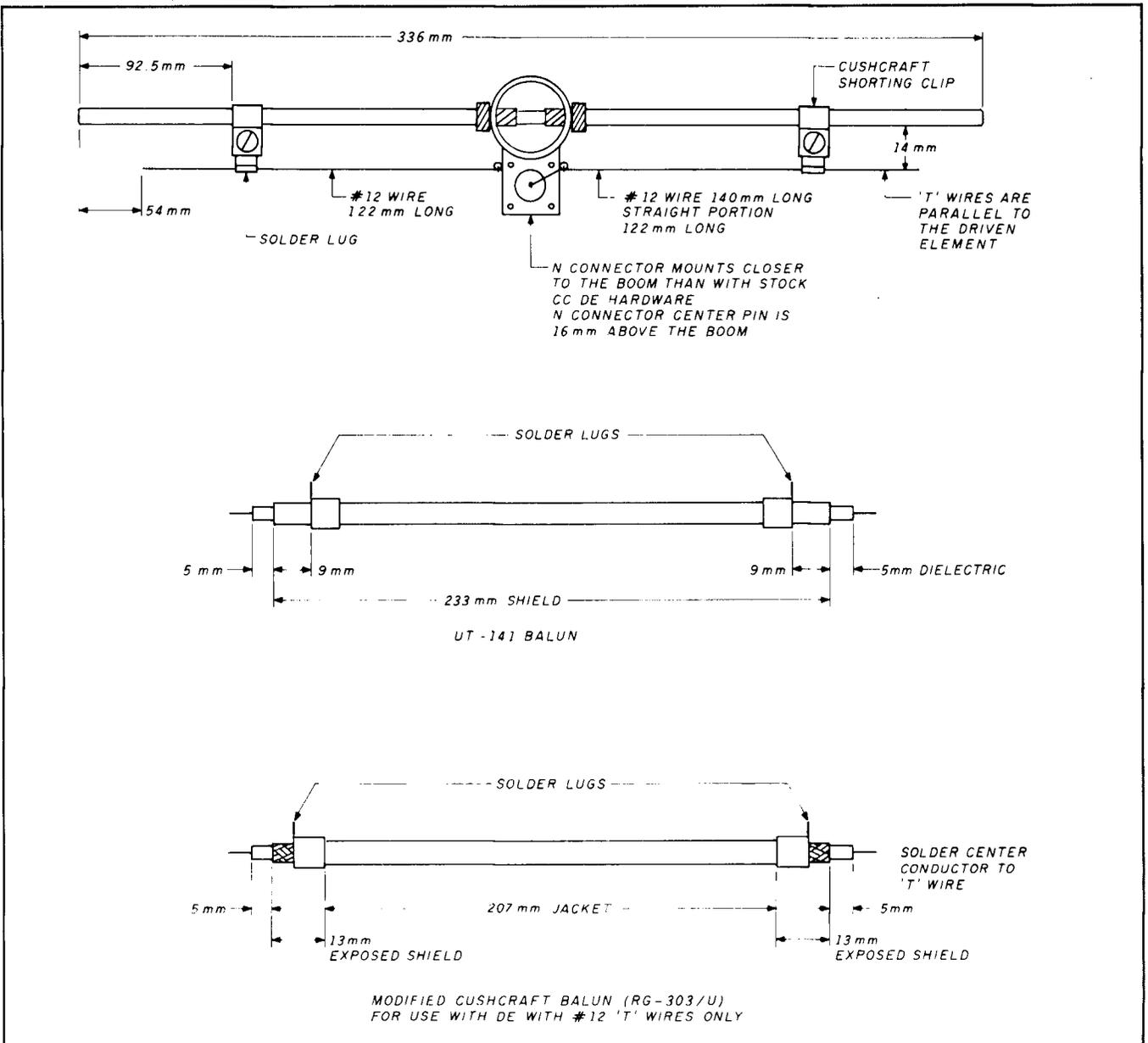
Performance

The computed E and H plane patterns for the 32-element Mark 4 Yagi in **Figure 4** show a very smooth lobe struc-

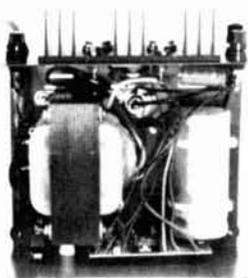
ture. The first sidelobes are 1 dB stronger than those in the original version. This seemed an acceptable tradeoff for a smoother overall lobe structure and significantly lower rear and mid-H plane lobes, in combination with higher overall gain. Calculated gain on MININEC is 17.9 dBd (20.1 dBi). Because of program inaccuracies and resistive losses, the real gain of the Yagi is closer to 17.8 dBd (19.9 dBi) — still an excellent figure for the boom length. DJ9BV examined the Yagi design using the more sophisticated NEC program. His results gave an excellent pattern correlation. The NEC-calculated gain figure of 17.8 dBd (19.9 dBi) also agrees closely with antenna measurements.

On the antenna range, the new model consistently measures about 0.2 dB higher than the earlier 24' long version. It also measures about 0.4 dB higher than my "high-gain" reference Yagi, the KLM 432-30 LBX. This places the real-

FIGURE 3



Construction detail of the 32-element Mark 4 driven element using no. 12 T wires. Further details for modifying the Cushcraft Balun for use with the no. 12 T wire match.



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RM-35A	25	35	5 1/4 × 19 × 12 1/2	38
RM-50A	37	50	5 1/4 × 19 × 12 1/2	50
• Separate Volt and Amp Meters				
RM-12M	9	12	5 1/4 × 19 × 8 1/4	16
RM-35M	25	35	5 1/4 × 19 × 12 1/2	38
RM-50M	37	50	5 1/4 × 19 × 12 1/2	50

RS-A SERIES



MODEL RS-7A

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A	2.5	3	3 × 4 1/4 × 5 1/4	4
RS-4A	3	4	3 1/4 × 6 1/2 × 9	5
RS-5A	4	5	3 1/2 × 6 1/4 × 7 1/4	7
RS-7A	5	7	3 3/4 × 6 1/2 × 9	9
RS-7B	5	7	4 × 7 1/2 × 10 3/4	10
RS-10A	7.5	10	4 × 7 1/2 × 10 3/4	11
RS-12A	9	12	4 1/2 × 8 × 9	13
RS-12B	9	12	4 × 7 1/2 × 10 3/4	13
RS-20A	16	20	5 × 9 × 10 1/2	18
RS-35A	25	35	5 × 11 × 11	27
RS-50A	37	50	6 × 13 1/4 × 11	46

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MODEL RS-35M

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RS-12M	9	12	4 1/2 × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10 1/2	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13 1/4 × 11	46

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MODEL VS-35M

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MODEL	Continuous Duty (Amps)			ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
	@13.8VDC	@10VDC	@5VDC			
VS-12M	9	5	2	12	4 1/2 × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10 1/2	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13 1/4 × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5 1/4 × 19 × 12 1/2	38
VRM-50M	37	22	10	50	5 1/4 × 19 × 12 1/2	50

RS-S SERIES



MODEL RS-12S

- Built in speaker

MODEL	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	5	7	4 × 7 1/2 × 10 3/4	10
RS-10S	7.5	10	4 × 7 1/2 × 10 3/4	12
RS-12S	9	12	4 1/2 × 8 × 9	13
RS-20S	16	20	5 × 9 × 10 1/2	18

world gain of the new 32-element Mark 4 Yagi at about 17.8 dBd (19.9 dBi). Earlier measurements with the Mark 2 Yagi were slightly optimistic; the real-world gain for this version was about 17.5 dBd. (This has also been confirmed by NEC analysis.)

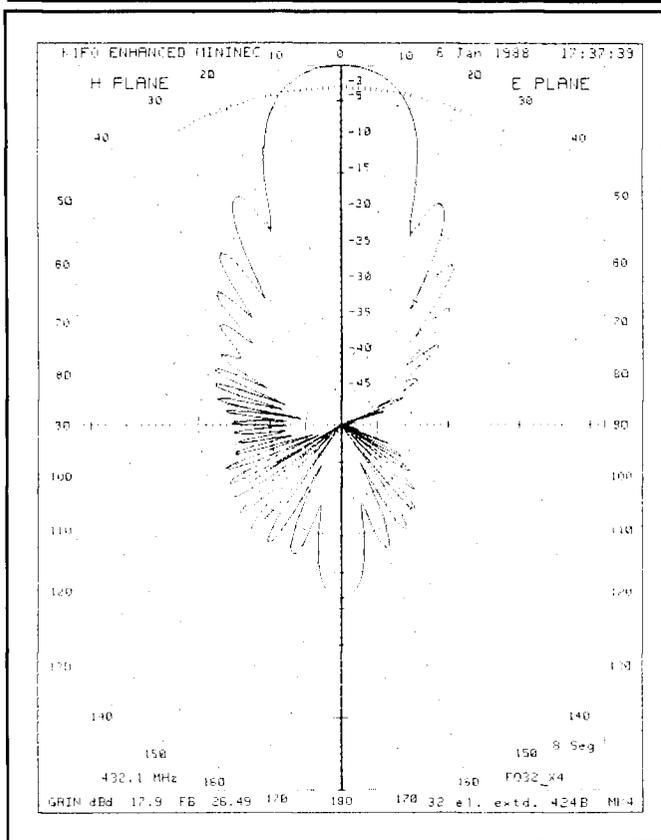
Array temperature is an important parameter on EME and for high-performance tropo stations. Array noise is the combination of noise received by the array (manmade or natural) and the noise generated from resistive losses in the

material used to make it. DJ9BV calculated that an array of four of the 32-element Mark 4 Yagis pointed at cold sky has a noise temperature of 25K. This figure is 5K lower than the original Yagi design — a significant number on EME. Array noise measurements, using earth to cold sky and stellar sources to cold sky, place the array temperature somewhat higher than the calculations. Measured array temperatures for four Yagis arranged 2 x 2 are about 30K for the new Mark 4 Yagi, and about 37K for the old version.

To calculate the overall system temperature, you have to add the phasing line, relay, and balun losses, and the receive system noise temperature to the array noise. For a high-performance EME system with very low loss phasing lines (like the Andrew HeliAx™ and a 25K preamplifier), this reduced array temperature would provide an additional 0.5-dB signal-to-noise improvement on receive over the original 32-element Yagis. Including the additional gain of the improved design, you can expect to hear your own moon echoes almost 1 dB stronger — a significant improvement for no array size increase.

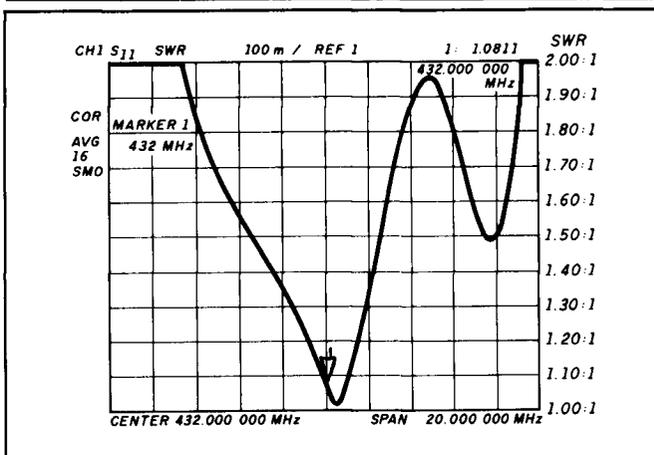
Figure 5 is a network analyzer plot of the driven element match. When you compare this with Figure 1 (the same plot for the original 32-element Yagi) you can see that the

FIGURE 4



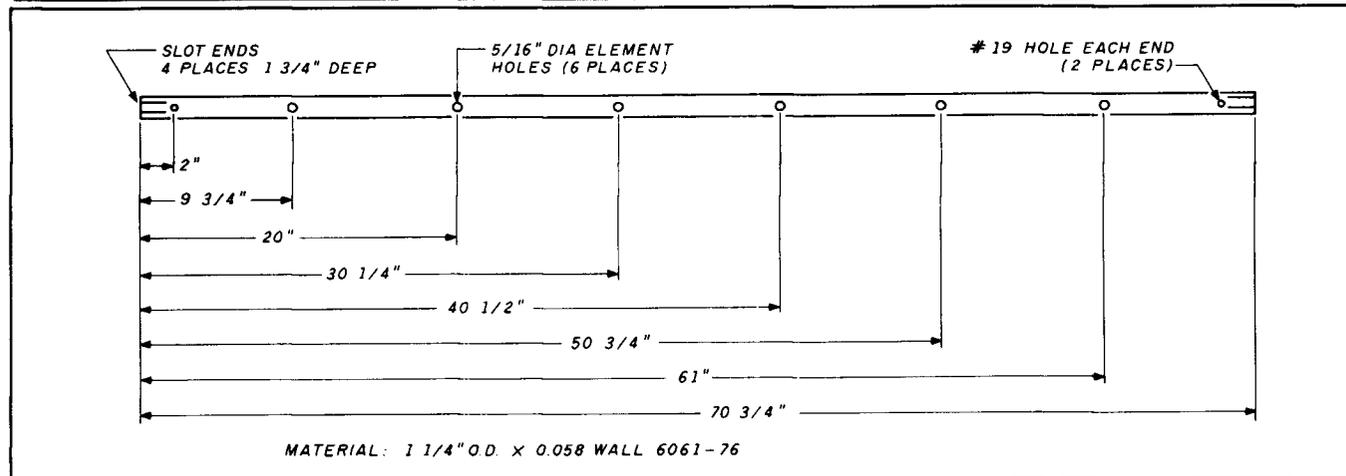
Calculated E and H plane patterns for the K1FO 32-element Mark 4 Yagi.

FIGURE 5



Resonant frequency and VSWR bandwidth of the improved 32-element Mark 4 Yagi.

FIGURE 6



Element mounting dimensions for the center section of the boom for the 36-element extended modified 424B Mark 4 Yagi.

match bandwidth is much broader. On the 32-element Mark 4 Yagi, the SWR is now less than 1.2:1 for almost 2.5 MHz. This is 2-1/2 times wider than the original. The SWR is less than 2:1 over 16 MHz, or more than 3 times greater than the earlier version.

With the revised dimensions, the 32-element Yagi behaves well when wet. The VSWR curve shifts down in frequency approximately 2 MHz under simulated heavy rain conditions. This raises the SWR to a still acceptable 1.35:1 when the antenna is very wet. This is a significant improvement over the 32-element Mark 2 Yagi, which would show about a 2.2:1 VSWR in heavy rain.

Proper stacking distances for the 32-element Mark 4 Yagi are 82 inches in the E plane and 78 inches in the H. At these distances, the stacking gain (before phasing line losses and mechanical errors are factored in) is over 2.9 dB in each plane. A 4-Yagi EME array using low-loss phasing lines would have 23.3 dBd (25.5 dBi) array gain. This is more than adequate to work a number of different 432-MHz EME stations. An 8-bay 32-element Yagi array has enough gain to give you a standout EME signal.

A longer 36-element version

NC1I did his portable EME operations in the summer, usually the worst time of year for EME. Although the original 4 x 32 element Yagi array performed well, I wanted a little extra performance without having to add more Yagis. I chose the 27-1/2' length because it was the minimum size increase which would make a significant performance improvement. (See Table 2 for element length and spacing details.)

Electrically, the design is virtually identical to the 32-element Mark 4 Yagi. Mechanically, the changes are a little more detailed. I built a 6' long 1-1/4" diameter center boom section from scratch for the long Yagi. I reinforced this new center section with a 1-3/8" outer diameter, 0.058" wall thickness, 24" long piece of 6061-T6 aluminum tubing. The 1-3/8" diameter tube is centered at the mast mounting point. This arrangement gives you a very rugged (though slightly heavy) boom section.

TABLE 2

Dimensions K1FO 36-element Yagi 27'-4-1/8" boom.

Element Spacing (inches)	Element Length (mm)	Boom Diameter	Element
1.000	347	1"	REF
5.250	327		DE
7.875	322		D1
11.563	313		D2
16.813	308		D3
23.563	304		D4
31.875	300		D5
42.125	296		D6
52.375	293	D7	
62.625	291	D8	
72.875	289	D9	
83.125	287	D10	
93.375	285	1 1/8"	D11
103.625	284		D12
113.875	283	D13	
124.125	282	1 1/4"	D14
134.375	282		D15
144.625	282	1 3/8"	D16
154.875	281		D17
165.125	279	1 1/4"	D18
175.375	278		D19
185.625	277	D20	
195.875	275	D21	
206.125	275	1 1/8"	D22
216.375	274		D23
226.625	274	D24	
236.875	273	D25	
247.125	273	D26	
257.375	272	D27	
267.625	271	D28	
277.875	270	D29	
288.125	270	D30	
298.375	269	1"	D31
308.625	269		D32
318.875	268		D33
329.125	268		D34

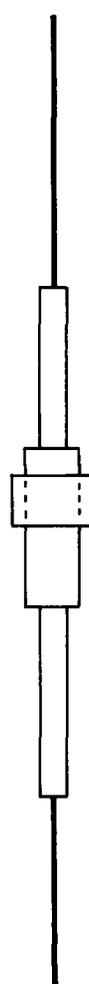
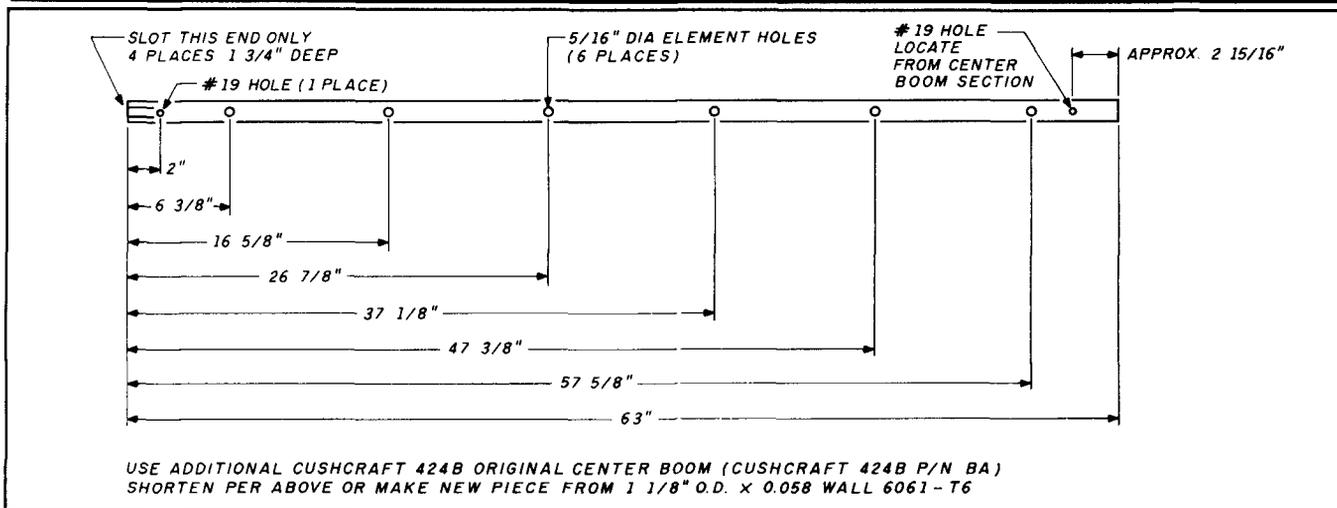


FIGURE 7



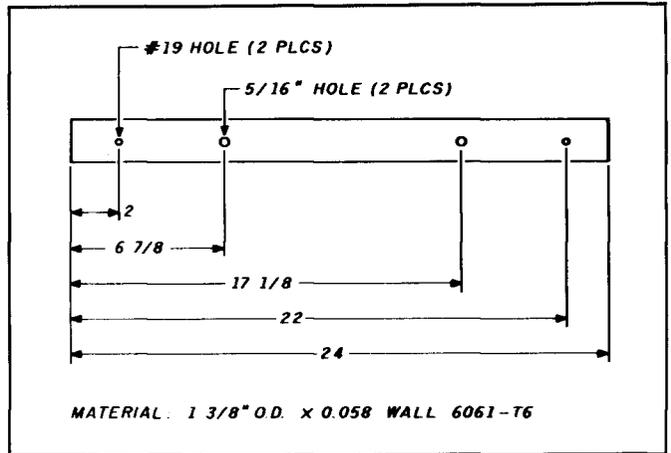
Detail for the new no. 2 rear boom section for the 36-element extended Mark 4 Yagi.

Though the doubled-up center piece may seem like overkill, it's very easy to bow a 0.058" wall tube when you tighten the mast bracket U bolts. A few degrees of bend in the boom may not be noticeable on a short Yagi, but when you translate this bend to a 27-1/2' long boom it becomes a significant curvature.

The new center boom section is detailed in **Figure 6**. **Figure 7** describes boom section no. 2 (between the rear and the middle section). You can make this second boom section from a spare 424B original center section, or from scratch. Just follow the drawing and use 1-1/8" diameter x 0.058" wall aluminum tube. The center boom reinforcing piece is shown in **Figure 8**.

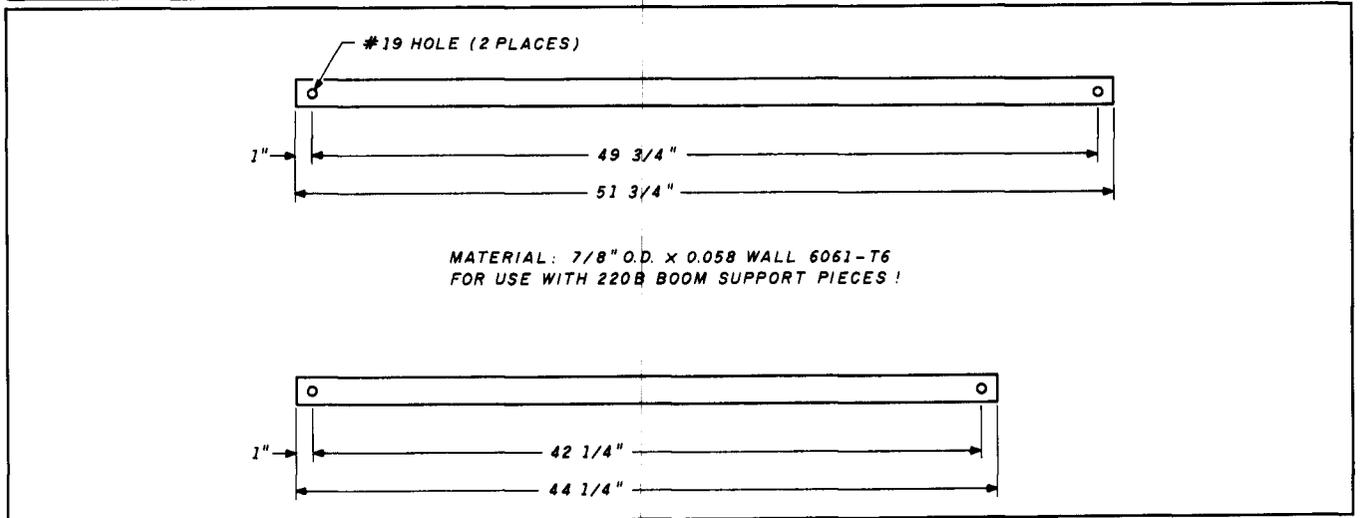
I made new pieces for the boom supports so I could extend them. I positioned the mast mounting point for the boom supports 30" from the boom. This creates a large enough angle and prevents overstressing the supports. Like the earlier Yagis, I used old-style Cushcraft 220B bent support pieces with longer homemade center sections. **Fig-**

FIGURE 8



Detail for reinforcing the center boom section of the 36-element extended Mark 4 Yagi.

FIGURE 9



Detail for the boom support pieces on the 36-element extended Mark 4 Yagi.

Figure 9 shows these new boom support splice pieces. Cushcraft has since changed the design of their boom supports. If you don't want to make your own supports from scratch, Cushcraft 4218-XL boom supports will do the job.

Electrical changes

You'll notice that the element lengths start out 1 mm shorter than on the 32-element Mark 4 Yagi. I changed the length to keep the gain center frequency in the right place. Remember that the Yagi's center frequency oscillates up and down as you add directors. Tapered designs like this one minimize the effect, but the trait still exists.

The driven element was easy to set up for both a good SWR at 432 MHz and a minimum centered above 432 MHz. This is the best way to ensure good wet weather performance. The match bandwidth on the 36-element model is actually better than on the 32. The driven element match on the 36-element Yagi also was relatively insensitive to the balun length — another good sign. The driven element for the 36-element Yagi is outlined in **Figure 10**.

Stacking the 36-element Yagi

At a 12-wavelength boom length a good Yagi will have a nearly symmetrical pattern. You can see from the calculated pattern in **Figure 11** how the H plane is starting to show nulls at 90 degrees in the pattern, similar to the E plane. The -3 dB beamwidth is still slightly wider in the H plane, even at this long boom length. This indicates that optimum spacings will be close but not quite equal in both planes.

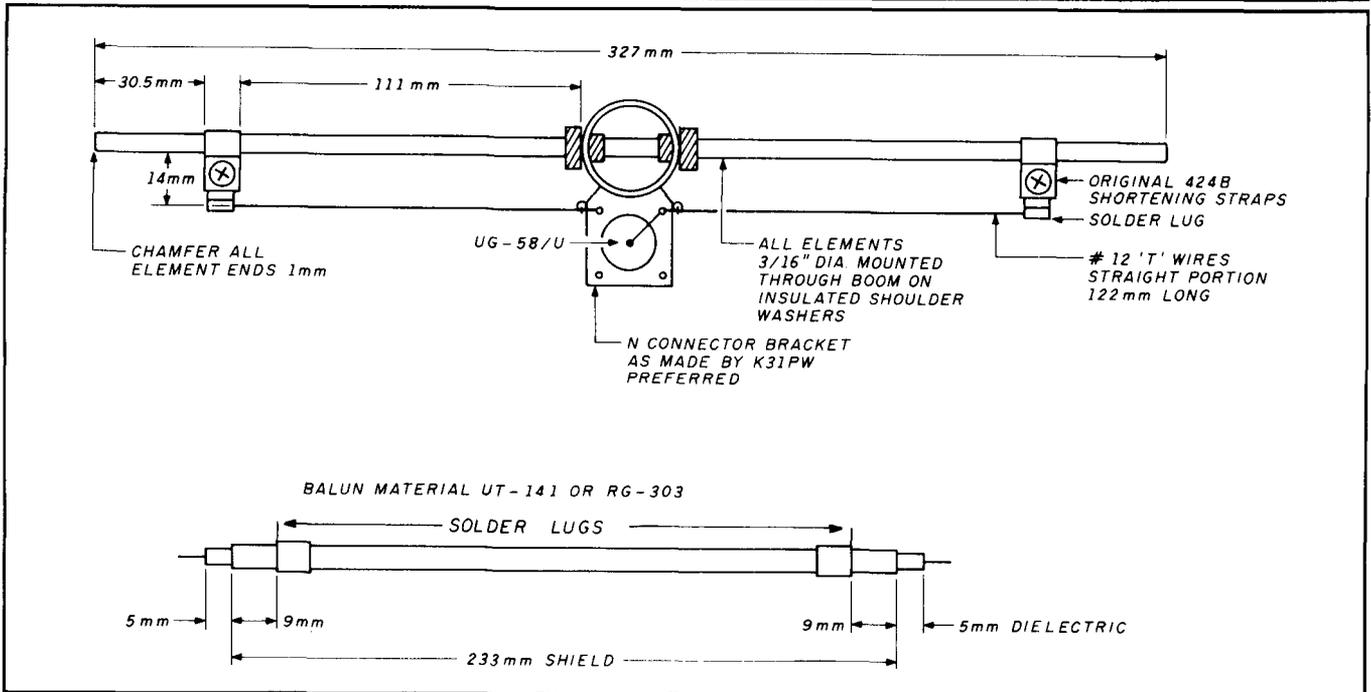
The optimum spacings for the 36-element Yagi are 87" in the E plane and 85" in the H plane. At these spacings the theoretical stacking gain in each plane is 2.9 dB.

Performance of the 36-element Yagi

The calculated pattern for the 36-element Yagi (**Figure 11**) is quite similar to the 32-element Mark 4 Yagi. The side-lobe structure is almost identical. The main lobe E and H plane beamwidths are about 1 degree narrower than the 32-element Yagi at 18 x 18.5 degrees.

Measured gain of the 36-element Yagi is approximately

FIGURE 10



Detail of the driven element of the 36-element Yagi using the no. 12 T wire match. Total boom length is 27' 4-1/8".

0.6 dB higher than the 32-element Mark 4 Yagi at 18.3 dBd or 20.5 dBi. Array temperature is even better than the shorter Yagis at a calculated 24K. Measurements indicate an array temperature under 30K.

Of course, on-the-air performance is what counts. WA9FWD reported a significant improvement when he upgraded from four of the 32-element Mark 2 Yagis to four of the 36-element model.

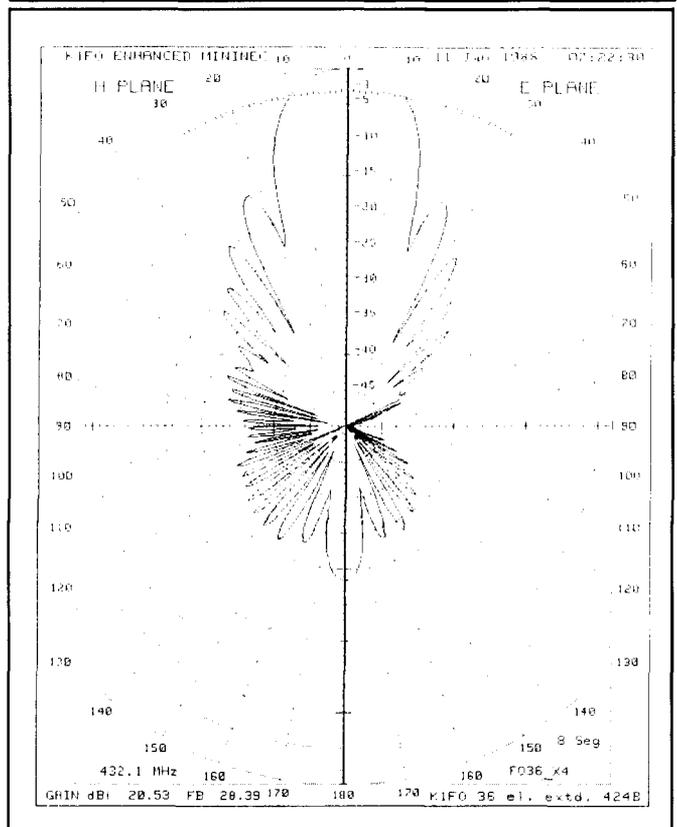
NC11 and I recently rebuilt our portable EME array to use four of the new 36-element models. The old array was 4 x 32 element Mark 2 Yagi. The new array seems to follow the predicted improvements. Measured sun noise is up 1.5 dB from the best we measured with the old array. The new array uses the same phasing lines, power divider, and preamp as the old one. The sun noise improvement is in the expected range. Gain of the 36-element Yagi is 0.8 dB higher than the 32-element Mark 2. Signal-to-noise improvement due to noise pickup and resistive losses is calculated to be over 0.5 dB. The sum, 1.3 dB, is close to the measured 1.5-dB improvement.

We first tested the new array at W1NY during the January VHF contest. We made a total of 15 EME QSOs in only 5 hours of EME operation, all with a bad antenna relay! After the contest we fixed the relay and activated the array the following weekend. We had 16 hours of EME operation spaced over the two weekends. We made 34 EME QSOs with 26 different stations, all on random. Echoes were noticeably better with the new array.

Conclusion

A top performing 432-MHz Yagi must have a proper balance of several desirable characteristics:

FIGURE 11



Calculated E and H plane patterns for the K1FO 36-element Mark 4 Yagi.



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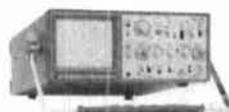
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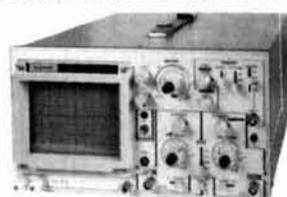
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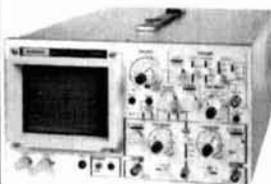
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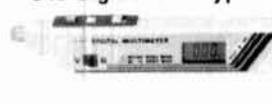
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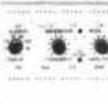
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Once you've defined these electrical traits, you must construct your Yagi so that it will not only work in the real world, but stay up and retain that performance for many years. The K1FO 32 and 36-element Yagis have an excellent balance of these design goals, especially when you consider that they can be made easily from an existing commercial Yagi and its spare parts.

The 4-element (41" long) Yagi extension appears to be worth the effort. During initial operation with the 36-element Yagi arrays, it seemed we finally had an array that was better than a 4-Yagi array was supposed to be. As any given Yagi design is extended, its driven element impedance and rear lobe structure oscillate up and down. At a length of 27-1/2', the pattern and driven element are in optimal combination. If you plan to build a long 432-MHz Yagi from scratch, I suggest that you take a serious look at the 36-element model.

The results of the computer analysis suggest that the design could be extended still further with good results. But keep in mind that the boom would have to be extended by more than the same percent of boom change when going from 32 to 36 elements for another 0.5-dB gain. You'd need to add at least five more elements, possibly six, to see an equivalent improvement. This would make the boom almost 32 feet long. Since the 36-element Yagi weighs over 12 pounds and has a wind area of over 3 square feet, an even longer Yagi may quickly become unmanageable. A very long object also develops quite a momentum when it's moved. This added inertia requires a large increase in the mechanical strength of the array's stacking frame.

Acknowledgment

I'd like to thank Rainer Bertelsmeier, DJ9BV, for his NEC analysis and array temperature calculations of my Yagi designs. 

REFERENCES

1. Steve Powlishen, K1FO, "High-Performance Yagis for 432 MHz," *Ham Radio*, July 1987, page 8
2. Steve Powlishen, K1FO, "An Optimum Design for 432-MHz Yagis, Part 1," *QST*, December 1987, page 20

MARCH

WINNERS

Congratulations to George Gorsline, VE3FIU, our March sweeps winner and Rick Littlefield, K1BQT, author of March's most popular WEEKENDER — "Solo-16 Acoustic CW Speaker." Both will receive a copy of *The Radio Handbook* by Bill Orr, W6SAI.

Our WEEKENDER sweepstakes ends with the April issue. You still have a chance to send your April evaluation cards and win. The winners of the April sweeps will be announced in the June issue of *Ham Radio*.

Thanks to all of you who sent in your cards. Your comments have been invaluable to us!

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CONVERTER TUNES 4 TO 18 MHz WITH NO BANDSWITCH

By Jack Najork, W5FG, 723 Flamingo, Duncanville, Texas 75116

If you buy a new SSB rig today, chances are it includes a "general coverage" receiver that tunes from 100 kHz to 30 MHz. This range includes all the international short-wave bands and many other radio services — lots of interesting listening!

Those of us with older rigs don't have this feature. If we want to eavesdrop on these frequencies we can trade in our rigs, invest several hundred dollars (or more) in a general coverage receiver, or dig into the junkbox and create some form of compromise. My junkbox is much larger than my bank account (thanks to Dallas' famous monthly electronic sidewalk sale), so I chose the compromise approach and built a converter.

My converter covers 4 to 18 MHz in two bands: 4 to 11 MHz and 11 to 18 MHz. This span includes most of the popular short-wave bands as shown in **Table 1**. To use this converter your SSB receiver needs only to tune to 3.5 MHz, and be able to receive in the AM mode (since most broadcasters still use this form of modulation).

How it works

If you look at **Figure 1**, you'll see that the unit consists of RF amplifier Q1, local oscillator Q2, and mixer Q3. The two bands are covered without a bandswitch by using an IF of 3.5 MHz. By selecting the appropriate oscillator frequency range, you can tune this range above or below the desired incoming signal. A bit of math is all you need to understand this approach.

The oscillator range is 7.5 to 14.5 MHz. Incoming signals from 4 to 11 MHz are mixed with the oscillator to produce the 3.5 MHz IF. Signals from 11 to 18 MHz mixed with the oscillator will also produce an IF of 3.5 MHz. All that remains is to make sure you can separate the two incoming signals. Because at any one oscillator frequency the two incoming signals are 7 MHz apart, this isn't a great problem.

How does this converter compare with a new general coverage receiver? First of all, it doesn't have the same extensive frequency coverage. Secondly, the tuning rate is coarser

(comparatively speaking). Each band covers 7 MHz, so the kilohertz zip by at an astounding rate as you tune. Tune very slowly and use your SSB receiver for fine tuning. On the plus side, the converter's sensitivity is excellent — 10 feet of antenna will tune in the world! And...you've saved lots of money. Despite its limitations, the converter fits the bill for casual short-wave listening.

Circuit details

RF amplifier input C1-L1 comprises a high-Q, lightly loaded, tuned circuit. This is essential for good band separation. A tapped toroid coil, along with light coupling to Q1 and loose antenna coupling, help keep the Q high. (If you don't have a suitable toroid, I've included specs for a solenoid-type substitute.) Space it at least 1 inch from metal surfaces on all sides to maintain high Q. C1 is a junkbox broadcast variable, with sections in parallel for a capacity range of 15 to 500 pF. This tunes the 4 to 18 MHz span without bandswitching.

Most SSB receivers have excellent sensitivity at 3.5 MHz, so the converter doesn't need high gain. Including an RF stage lets you use a small, indoor antenna (unless you live in a shielded building). This, in turn, aids front end selectivity and avoids possible overload. RFC1 in the drain of Q1 peaks up gain at the higher frequencies.

TABLE 1

International short-wave broadcast bands.

5.95 — 6.2 MHz — 49 meters
7.10 — 7.5 MHz — 41 meters
9.50 — 9.98 MHz — 31 meters
11.70 — 12.08 MHz — 25 meters
15.10 — 15.45 MHz — 19 meters
17.70 — 17.90 MHz — 16 meters
21.45 — 21.95 MHz — 21 meters

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Because of the large frequency spread of each band, you'll find a good-quality slow-tuning vernier dial is essential for oscillator tuning. It's best to do your dial calibration with a signal generator. You need to calibrate only one band; the second is automatically 7 MHz different from the first.

You should also do a rough calibration of C1 so you'll know which band you're tuning. To check the quality (selectivity) of your C1-L1 combination, tune to the 7 to 14-MHz calibration mark. Tuning C1 slowly from minimum capacity should first bring in 20-meter CW; as C1 is increased, 20 should drop out and 40-meter CW should appear. The C1 tuning peak should be very sharp and you can judge it best by watching your receiver's S-meter.

As I mentioned earlier, you should use a short antenna to minimize the possibility of overloading and decreased front end selectivity. 

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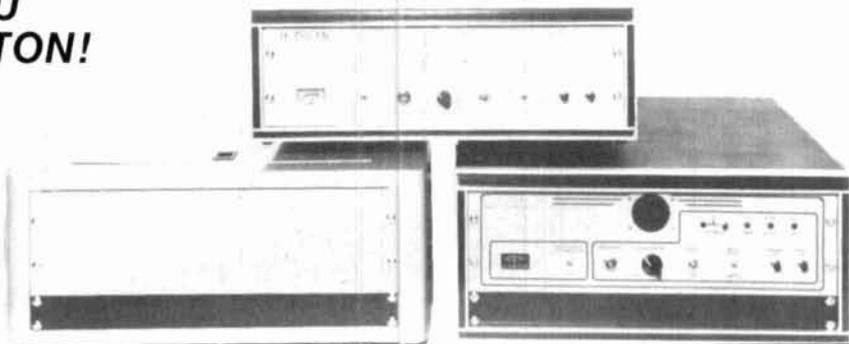
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THE FOLD WIRE FED TOP-LOADED GROUNDED VERTICAL

By Walter J. Schulz, Jr., K3OQF, 15225 Wayside Road, Philadelphia, Pennsylvania 19116

It's common practice to use the folded wire tapped to a grounded tower to feed the tower as a vertical radiator with top-loaded Yagi or quad antennas. You can use these installations as top-loaded verticals on 40, 75, and 160 meters. John True, W4OQ, did an empirical study of this type of system and wrote about the results in *Ham Radio Magazine*.^{1,2} After reading his study, I decided that I needed to be able to design the system on paper — independent of sets of interpolative electrical measurements. I had to answer a number of questions to achieve my design goals; there were two that concerned me most. What's the actual grounded tower electrical height with Yagi or quad antennas, and what's the impedance to be expected at a tap point on the tower? To resolve these questions, I looked at the physical aspects in a different way. I assumed that the whole antenna system was a transmission line UHF tank circuit (Figure 1).

It's well known that the quarter-wave transmission line will act as an impedance transformer and match two different impedances. This phenomenon is an advantage at ultra-high frequencies where lump-constant components of capacitance and inductance are too large to be used in a tank circuit. A quarter-wave transmission line can be used as a tank circuit;³ it will display a very high quality factor (Q). Keep the transmission line length at a quarter wave and place a tap along the length for matching impedances if you wish. This tap will transform impedances between the tap point and an active device like a tube or transistor.

Sometimes it's desirable to have a shorter length of transmission line. You can reduce the length of the line by using a loading capacitance at its open end. The transmission line is shorted at one end and capacitively loaded at the other. The tap point along the transmission line will always be a pure resistance with no reactive component, as long as the line is equivalent to a quarter wave which is parallel resonant.

The transmission line tank circuit can be compared with a tower capacitively loaded by a Yagi or quad antenna. The impedance is transformed by placement of a tap along the tower. It's difficult to figure out where to place the tap on

the tower or transmission line, as every Amateur location is unique. You can make a fairly accurate first approximation; this will place the tap near the optimum position on the tower for an impedance match. Once you've made your calculation, a little experimentation will show exactly where the tap should go.

Here's an example. A Rohn Model 25 tower has 90-degree electrical height on 3.8 MHz. This serves as a reference parameter for further calculations. Use the algorithm that follows to determine the tap point on the tower that gives you a 52-ohm match.

Establish the feedpoint impedance for the height and diameter of the tower using Schelkunoff's equation.⁴ In this instance his equations derive shape factor as $Z_0 = 316$, and a complex feedpoint impedance as $R_f = 37.7$ ohms + $j21.8$ ohms. Next, find the quality factor:

$$Q = \frac{X_f}{R_f} = \frac{+j21.8\Omega}{37.7\Omega} = 0.5782 \quad (1)$$

Treating the tower as a transmission line, find the loop resistance:

$$R = \frac{6.28(Z_0)}{Q(\lambda)} = \frac{6.28(316)}{(0.5782)(78.96)} = 43.4635 \quad (2)$$

$\lambda = 78.95$ meters

$$\lambda = \frac{3 \times 10^8}{f} = \frac{3 \times 10^8}{3.8 \times 10^6} = 78.9475 \text{ meters} \quad (3)$$

Find R1 (loop resistance) value on the tower:

$$R1 = \frac{8(Z_0)^2}{R(\lambda)} = \frac{8(316^2)}{(43.4635)(78.96)} = 232.773 \quad (4)$$

Find the ratio of R_f to R1:

$$\alpha = \frac{R_f}{R1} = \frac{52}{232.773} = 0.2234 \quad (5)$$

FIGURE 1

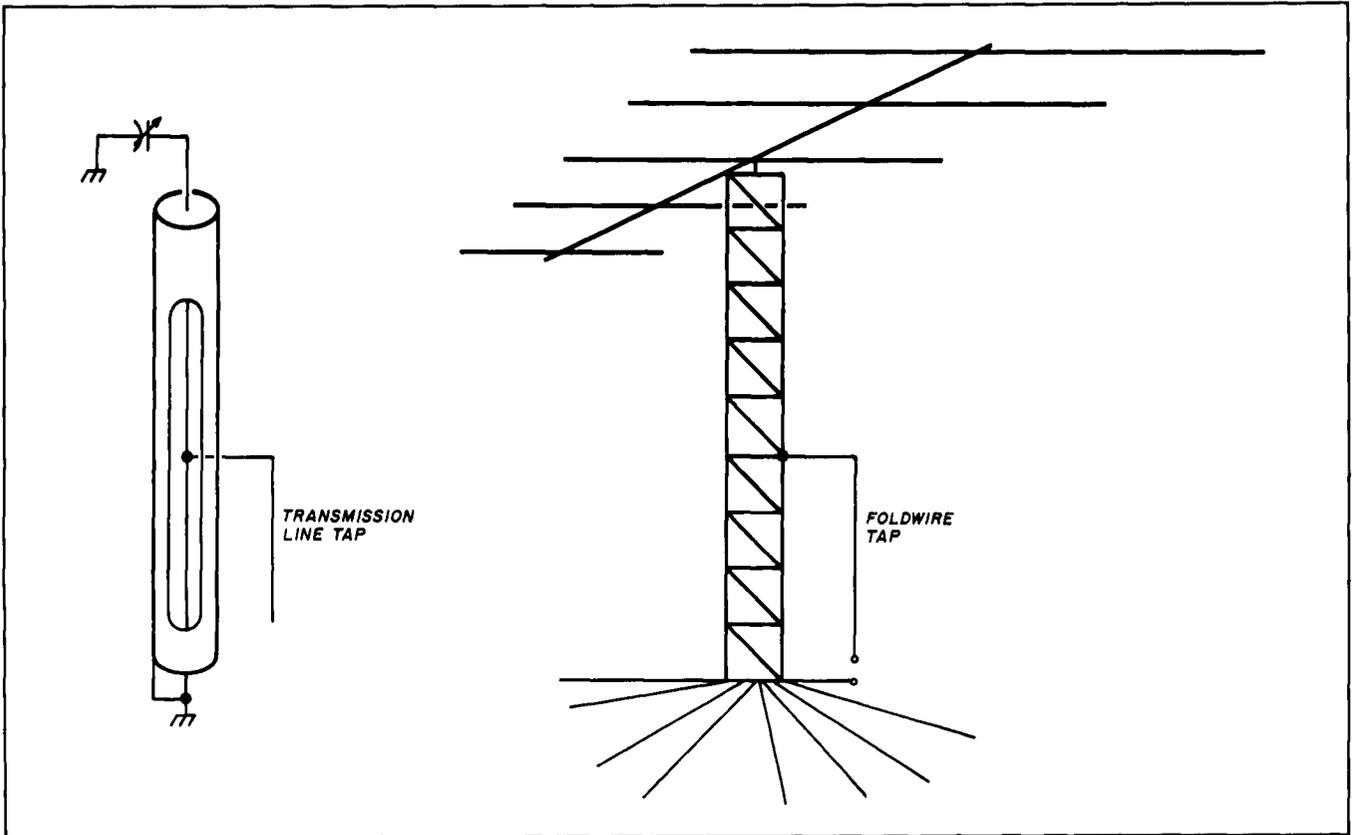


Illustration of a shunt fed tower and its electrical equivalent.

Note: R_t is the tap point impedance. In this case it's 52 ohms.

θ tap point on tower

$$\theta \text{ tap point} = \arcsin (\alpha \sin^2 \theta_{Gt})^{1/2}$$

$$\theta \text{ tap point} = \arcsin [0.2234 \sin^2 (90)]^{1/2}$$

$$\theta \text{ tap point} = \arcsin 0.4727$$

$$\theta \text{ tap point} = 28.2^\circ$$

(6)

Convert tap point height from degrees to feet above ground:

$$\frac{28^\circ}{360^\circ} (259) = 20 \text{ feet} \quad (7)$$

If you want to find other tap points above the tower's base and along its length, you need to know the total electrical height of the whole antenna system. Determine the tap point position on the tower length by using proportions to get the first approximation. The first approximation gives you a starting point from which to calculate tap placement on the tower. You may find you have to move the tap up or down. This adjustment, which increases or decreases resistance transformation at the tap point, is necessary for a match to the transmission line.

Electrical antenna height or line length

I'm sure you've seen how electrical antenna height differs from actual physical antenna height. Literature on the subject shows how this phenomenon has been put to use changing antenna current distribution and raising the radiation resistance. You might find it helpful to ask the follow-

ing questions about an antenna structure already supporting a Yagi or quad antenna, especially if you want to fold wire feed a grounded tower as a vertical radiator.

- What is the total antenna electrical height?
- What is the equivalent electrical height represented by top loading?
- What is the proper tower tap point for shunt feeding?

Consider a Rohn Model 25 tower 55 feet high. This tower supports a TH6DXX triband Yagi which, for our purposes, can be compared to a flat-top antenna with parallel wires. Assume the elements are parallel wires forming a flat top, but each is of a different length. You must average all element lengths (six in the example) to get a uniform length for use in a capacity equation by Grover.⁵ You must also average each element's different spacing along the boom. After averaging these boom lengths and spacings, you're ready to use these values to find the flat-top capacitance loading.

The average element spacing is 57"; the average element length is 276".

Find the capacitance:

$$C \text{ in pF} = 7.36 \frac{\lambda}{F} \quad (8)$$

$$Q = \log \frac{2h}{D} - K = \log \frac{1320}{57} - 0.874 = 0.4907 \quad (9)$$

$$P = \log \frac{4h}{dia} - K = \log \frac{2640}{1} - 0.874 = 2.5476 \quad (10)$$

$$F = \frac{P + (n - 1)Q}{n} - K_n = \quad (11)$$

$$\frac{2.5476 + ((6 - 1)(0.4907))}{6} - 0.252 = 0.5815$$

$$C = \frac{7.36 (23 \text{ ft})}{0.5815} = 291 \text{ pF} \quad (12)$$

The capacitive value represented by the Yagi is 291 pF. Find the capacitive reactance from the previous value:

$$X_c = \frac{1}{6.28 (3.8 \times 10^6) (291 \times 10^{-12})} = 144\Omega \quad (13)$$

Find the shape factor for the antenna tower:

$$Z_o = 60 \ln \frac{(2h)}{(a)} - 1 = 60 \ln (2) \frac{(660'')}{(2.95'')} - 1 = 306 \quad (14)$$

Find the equivalent electrical height of the whole structure:

$$\theta H_b = \text{arccotan} \frac{X_c}{Z_o} = \text{arccotan} \frac{144}{306} = 65^\circ \quad (15)$$

$$\theta H_a = 76^\circ$$

$$\theta H_b = 65^\circ$$

$$\theta Gt = 141^\circ \text{ total height } \epsilon \text{ electrical degrees}$$

Finding the tap point on the tower

You know that the tap point on the reference tower structure is 28 electrical degrees above the tower's base for a tower of 90 electrical degrees in length. This means you'll use proportions to find the tap point for antenna heights other than 90 degrees. Say you want to find the tap point for an antenna height of 141 electrical degrees. Let X be the unknown value for the tap point on the tower. Once you determine the tap point in electrical degrees, convert this value to feet.

$$\frac{90}{28} = \frac{141}{X} \quad (16)$$

$$3.2143 = \frac{141}{X} \quad (17)$$

$$X = \frac{141}{3.2143} = 44^\circ \quad (18)$$

Converting to feet:

$$\left(\frac{44}{360} \right) (258.9474) = 32 \text{ feet} \quad (19)$$

(See Figure 2.)

Preparing the tower for shunt feeding

One of the first things you must do when shunt feeding a grounded tower is insulate the upper portions of the tower from ground. Do this by placing egg insulators where the guy wires connect to the tower. It's usually advisable to break up the guy wires with insulators every 1/10 of a wavelength to prevent reradiation by the wires and radiation pattern distortion.

Once you've located the tap point, you can feed it in one of two ways. The first is to extend a folded wire straight out from the tap point and drop it down parallel to the tower's base. RF excitation will occur at the earth's surface. Another method (which gives slightly better bandwidth and results) is to build a cage about the same diameter as the tower, instead of using a single wire. You can hang the cage on an outrigger extending out from the tap point and running parallel to the tower down towards the ground. The cage works better than an aluminum pipe, and offers less wind resistance. It also decreases the amount of weight that's hanging off the tower.

Once the fold wire or cage is in place, there are a number of things to do before exciting the tower with RF power. To start, hang the outrigger arm at the first approximation tap point. (The arm can extend 1 to 3 feet out from the tower's side. The outrigger arm material is usually a light metal, like aluminum pipe, with a 1-inch outside diameter, that's clamped to the tower legs with U bolts.) Suspend the fold wire or cage from the first outrigger to a second one at the tower's base. Make sure the bottom outrigger has an insulator, so that the cage is insulated above the ground.

The impedance noise bridge is a valuable tool for determining the feedpoint impedance at the fold wire or cage feedpoint on the bottom outrigger. You can move the wire in towards the tower or away from it to determine which spacing gives the best impedance match, then try raising or lowering the tap point for a better match. Observe the reactance value on the bridge once you've found the tap point with the best match.

It may be a good idea to cancel out the inductive reactance before moving the tap point position, or the fold wire in and out from the tower. I've given approximate values for maximum capacitance to cancel out the inductive reactance for the following bands: 1000 pF for 160 meters, 500

FIGURE 2

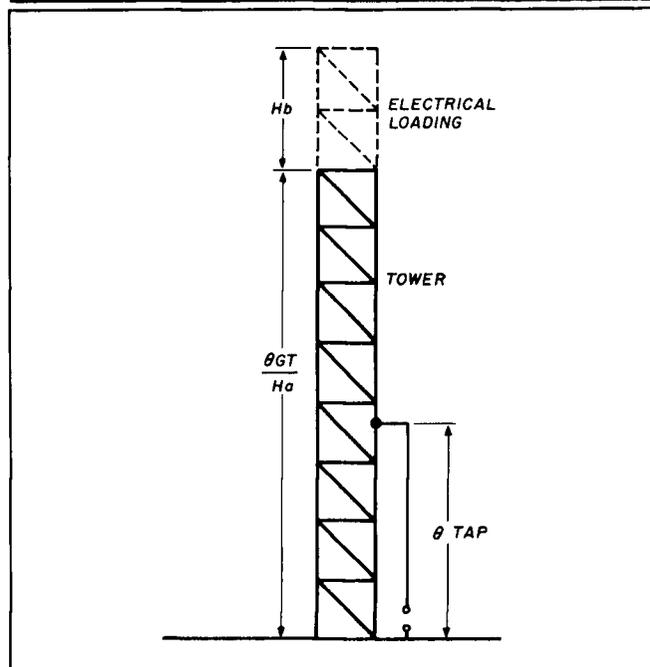


Illustration of the electrical height of the tower in degrees and the approximate placement of the tap in degrees.

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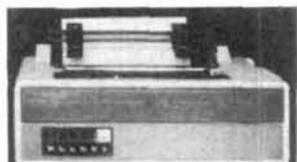
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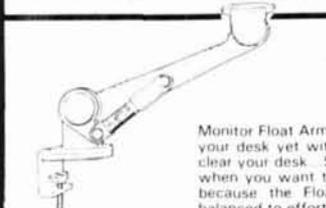
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Use it with any monitor. The arm adjusts vertically, and will accommodate monitors weighing up to 35 lbs. It rotates 90 degrees to the left or right along a 14" radius and arm floats. The unit has a universal 4" diameter mounting plate with a 3" dia bolt circle of 4 holes. We recommend you make a mounting plate of 1/2" thick plywood of a size to fit the

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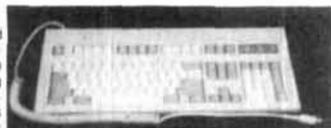
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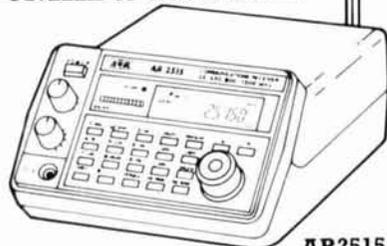
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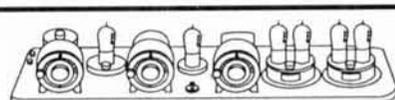
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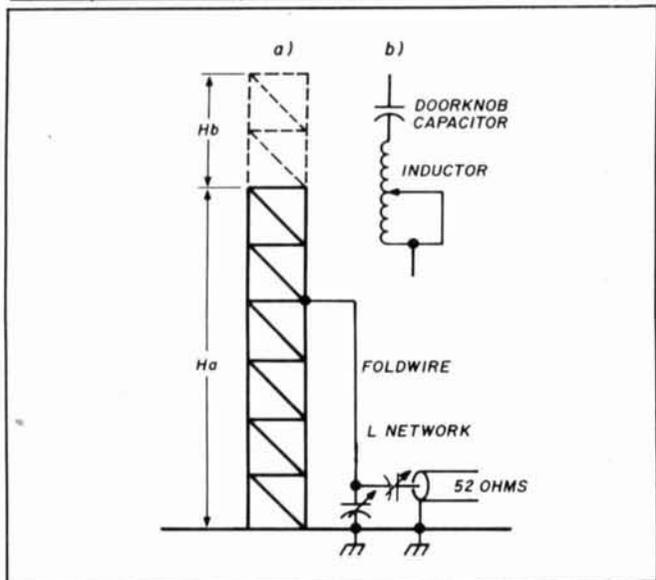
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FIGURE 3



Two methods of tuning out the inductive reactance at the feedpoint on a shunt-fed tower. (A) Variable capacitors are used to tune out inductive reactance. (B) Fixed capacitors are used with a series variable inductor to tune out any additional capacitance in the circuit.

pF for 80 meters, and 300 pF for 40 meters (Figure 3A). You can obtain these capacitance values by using a variable capacitor or a fixed capacitance (doorknob capacitor),

along with a series inductance to negate the fixed capacitance not needed to cancel out the antenna inductive reactance (Figure 3B). When you've canceled out the reactance component, match the resistance value in the usual way. Use an L or T-network, or toroid transformer, if the feedpoint isn't an exact 52-ohm match. Remember the fold wire will always present inductive reactance and this reactance must be canceled out with capacitive reactance. By using the fold wire you eliminate the need for a loading coil, increasing radiation efficiency.

It's also important to remember that you should have a good ground system with many radials when shunt feeding top-loaded towers. This doesn't mean just a few long radials, but a minimum of 60 radials 1/8 wavelength long. If your ground system is inadequate, it will be hard to find the tap point placement and total system adjustment will be very difficult. The calculations are made assuming the ground system will have little loss resistance. **73**

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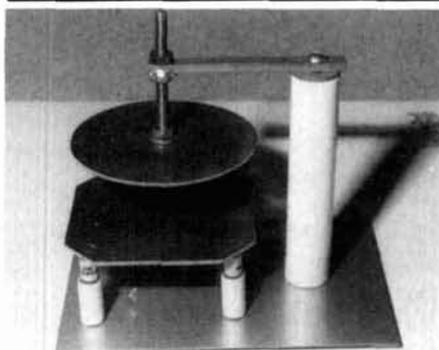
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Ham Notebook

Homebrew Neutralizing Capacitor

Most of us now have transceivers with an output of 100 watts. This is ideal for driving a high-power triode amplifier. Such an amplifier isn't difficult to build and can be quite economical if you have some 250Ts, 833As, etc. stored away somewhere. Perhaps you have a ham friend who works at a broadcast station. Stations often remove tubes from service that are perfectly good for Amateur use, especially for CW. Some

PHOTO A



Example of a completed Neutralizing Capacitor.

of the other components, however, may be difficult to find or too expensive. The neutralizing capacitor need not fall into this category.

My list of materials isn't cast in concrete. Use what you have on hand and calculate the capacity range from the formula in *The ARRL Handbook*. The items specified will result in a capacity of 1 pF at 3.0" spacing to 5.6 pF at 0.5" spacing.

Construction

Use pc board material with copper on one or both sides. Make the top plate using a circle cutter. Cut the bottom plate 4" square and then trim off the corners. Round the edges with a file and polish them with fine sand-



paper or crocus cloth. You can mount the bottom plate on just one center insulator, but I used four to make it sturdier. Countersink the top plate center hole and fasten on the 3" brass bolt with a lock washer and nut. Solder the bolt head to the pc board copper. This is the only tricky part of the entire project. Be careful not to make the countersink too deep. If you do, there will be a gap between the bolt head and the copper foil. It will be very difficult (I might even say nearly impossible) to bridge that gap with solder. Saw a slot in the other end of the 3" bolt for screwdriver adjustment.

PARTS LIST

- 1 - 4" diameter piece of pc board
- 1 - 4" square piece of pc board
- 1 - 1/4-20 3" long flat-head brass bolt
- 3 - 1/4-20 brass nuts
- 1 - 1/4" lock washer
- 1 - 3/16" x 3/8" x 4" brass bar
- 1 - 1" x 5" ceramic insulator
- 4 - 1/2" x 1.5" ceramic insulators
- Fiber washers and brass machine screws for the insulators

The support for the top plate is a piece of brass 1/8" x 5/16" x 4". If you use hollow brass, solder on a 1/4-20 brass nut after you drill a clearance hole for the 3" brass bolt. If the brass is solid, you can drill and tap it for the 3" brass bolt. You also need to countersink the four flat-head machine screws holding the bottom plate flush with the pc board copper.

Ken Leiner, N4LC

Adding 10 MHz to the HyGain HyTower

Although it's an older antenna, the HyGain HyTower is still in use in the United States, Canada, and around the world. The HyTower is basically a quarter-wave vertical. The antenna incorporates various tuning stubs; these decouple it for bands other than 40 and 80 meters. For 160 meters, the antenna uses a base-loading coil.

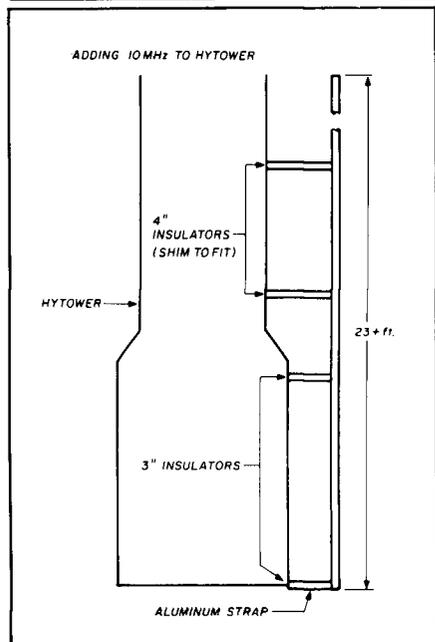
I've operated with the HyTower at W5UOJ for almost 16 years, and it was used when I got it! It's never failed on 160, 80, or 40 meters. Because I have beams for 20, 15, 10, etc., I never installed the stubs for 20, 15, and 10 meters on this particular antenna.

After the 30-meter band (10.100 to 10.150 MHz) opened, I tried the HyTower with an old Collins S-Line. The SWR was relatively low and it worked after a fashion. Because there's a coil for 160 installed in the 80-meter line (mostly shorted out for 80 meters), I could vary the SWR by moving the tap on the coil. Apparently the antenna was working as a three-quarter wavelength vertical.

I didn't get very good signal reports with the antenna this way. It worked, but...! I decided to add a matching section for the 30-meter band.

I garnered several pieces of aluminum from the antenna graveyard and pieced them together. Most of my matching section addition was 1-inch tubing. I inserted a shorter piece of 7/8-inch tubing into the end for tuning.

I started with a total length of just over 23 feet, and tuned the antenna for the best SWR by telescoping the 7/8-inch section up and down inside the 1-inch section. I installed two 3-inch stand-off insulators along the first section of tower and two 4-inch insulators on the second section (see Figure 1 for details). I shimmed them with small sections of plastic from margarine tubs to bring the 23+ foot section vertical. This took care of almost 16 feet; I left the remaining 7 feet freestanding.

FIGURE 1

Mechanical details for attaching the 23-foot matching stub to the Hygain Hytower.

Next I strapped the bottom of the newly added stub to the bottom of the tower leg with a scrap of aluminum. (You could probably use heavy wire.) Then I adjusted the antenna for minimum SWR. Because my antenna is mounted next to the house, I could make these adjustments with the stub mounted permanently. If your HyTower is mounted away from a building, put up the lowest and highest insulators and make the tuning adjustments by moving the antenna stub up and down. You can mount the stub permanently when the SWR is the lowest.

My stub was just short of 23 feet; the actual length of yours may differ. Because the antenna is mounted next to the house, and since the stubs for 20, 15, and 10 meters aren't installed on it, the final stub length could vary. Always check the SWR before finalizing your work!

The stub adds a third antenna in parallel with the existing 80 and 40-meter sections. With a good ground system this antenna can easily work the world. Stateside contacts are also much easier to make.

I spent less than an hour on this modification, and the time was well spent. It cost me nothing because I had all the materials on hand. Even if you have to purchase everything, it should cost less than \$20 — and prob-

ably less than \$10. Of course, the modified HyTower isn't a beam. But considering that many stations working the 30-meter band are using dipoles and even long wires, it does make for a better-than-average antenna! See you on 30!

Glen Zook, W5UQJ

The N5NBU "Nice but Ugly" \$1.29 Antenna

Living in an apartment presents special challenges for Amateur Radio operators. Limited space and landlord rules pose difficulties, particularly when it comes to antennas. When I was told that my ground plane had to come off the balcony, I decided enough was enough and the "Nice but Ugly" \$1.29 antenna was born. (See Figure 1.)

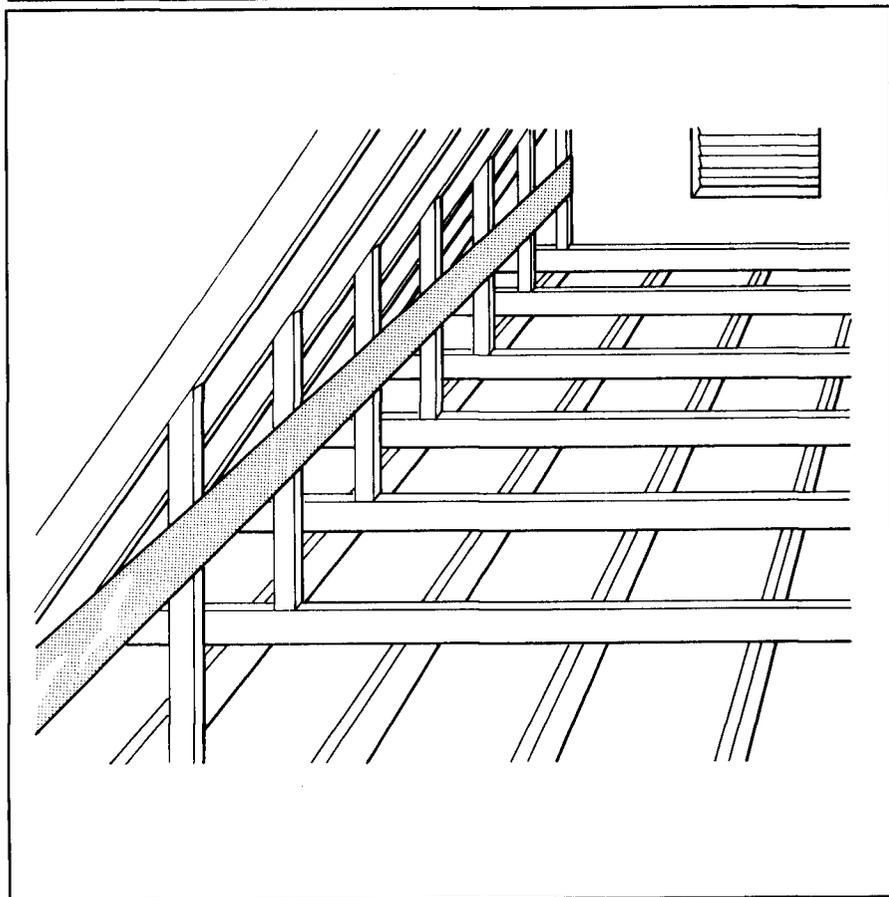
My requirements were simple. I needed to put up an antenna that was

as effective as possible within the space available. Then the XYL added another constraint: "Try not to spend any money on it."

I tried and discarded gems like "The Bedroom Wall Quad" (two square wire loops on opposite walls of the bedroom) and "The Guillotine" (a long wire strung throughout the place, which was a real adventure in the dark!).

Then, by sheerest chance, I saw some light at the end of the tunnel — or rather, in the ceiling. While replacing a light bulb in the closet, I noticed an access panel in the ceiling. I lifted it up and, lo and behold, space! A whole attic just waiting for an antenna installation! I immediately sat down and started contemplating antenna designs.

I made a couple of trips into the attic with measuring tape in hand and my plan began to take shape. Because I had severe budgetary restrictions, the antenna had to be simple and built with materials I had around the house. What's cheap, long, and conductive...?

FIGURE 1

Installed view.

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MRF137	24.00	MRF644	23.00	2SC2904	32.50
MRF138	35.00	MRF646	25.00	2SC2905	34.50
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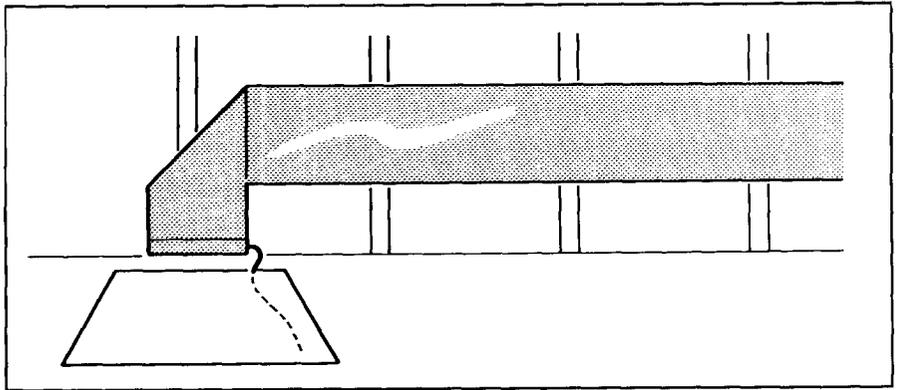
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FIGURE 2



End-feed view.

Eureka! Aluminum foil! Because the antenna would be installed in the attic there wouldn't be any wind loading, and there were convenient vertical supports (roof trusses) running the length of the attic. I went into the kitchen, got out the roll of foil, and measured off 66 feet (the length of my attic, and two full waves at 10 meters — how convenient) plus a little bit for anchoring purposes. I decided an end-fed antenna would be perfect as it involved just a single wire to my tuner.

After measuring out the foil and locating my box of tacks, I once again climbed into the attic. Starting at the far end, I rolled and flattened about a foot of foil for a good strong anchor, held it vertical, and thumbtacked it to the support on the far end of the attic. Working my way back to the access hatch, I unrolled the foil and tacked it to every fourth vertical. They're spaced about 18 inches apart. I continued this rather tedious process (tedious because the foil is quite delicate) until I came to the access panel. Then I folded the foil over to bring it down to the edge of the access. To connect the antenna to the tuner, I scraped off about 2 feet of insulation from an 18-foot piece of no. 16 enameled wire, then soldered it (sort of) to a piece of foil about a foot long using "the bigger the blob, the better the job" method. I rolled this up into the end of the antenna, flattened it, tacked it down to the edge of the access, and connected the wire to my MFJ Deluxe Versa Tuner II. (See Figure 2.)

Now it was time for the acid test. I tuned the transmitter up into the dummy load, switched to the antenna, and heard the (new and thrilling)

sounds of DX stations. After finding a clear spot on 10 meters to finish my tune-up, I called CQ and immediately got my own mini-pileup. The SWR meter said I was matched up with a 1:1 ratio, and the stations I was talking to were giving me S9+20 reports. It was the same story with CW on 15 and 40 meters. On 80 meters I have a problem with RF "bites," but a good ground should cure that problem. I've greatly alleviated and almost eliminated my TVI. My primary problem was fundamental overload. Now just turning the rabbit ears parallel with the plane of the antenna greatly reduces and just about gets rid of TVI on channel 2 (my problem channel).

I hope you'll try my Nice but Ugly antenna. Just remember basic safety precautions regarding power lines, etc. and you'll have a safe and trouble-free installation.

Now, I wonder if I can do something with these tin cans...

Don Lane, N5NBU

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SP144VD	144-148	<1.6	15	0	DGFET	\$59.95
SP144VDA	144-148	<1.1	15	0	DGFET	\$67.95
SP144VDG	144-148	<0.55	24	+12	GaAsFET	\$109.95
SP220VD	220-225	<1.9	15	0	DGFET	\$59.95
SP220VDA	220-225	<1.3	15	0	DGFET	\$67.95
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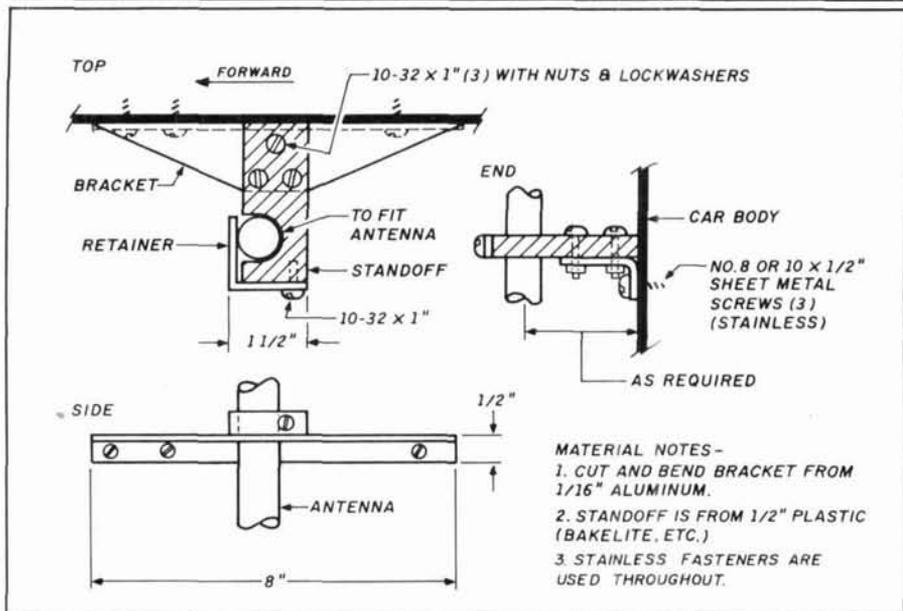
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FIGURE 1



Anti-rotate bracket for side-mounted mobile antennas.

I attached the aluminum bracket to my car with stainless steel sheet metal screws available from most hardware stores. My installation fits two different diameter antennas, so I cut the hole in the standoff for the larger one. I bushed out the smaller antenna with a 4-inch length of clear plastic hose, also from the hardware store. I split it lengthwise and attached it with a hose

clamp. I added the L-shaped retainer at the front of the mount to prevent rotation in the forward direction (less likely). You'll need to vary my dimensions to fit your car and antenna(s).

This device has stood the test of several thousand miles and many QSOs. Thanks to Bob, W1KSK, for his original suggestion.

George Wilson, W1OLP

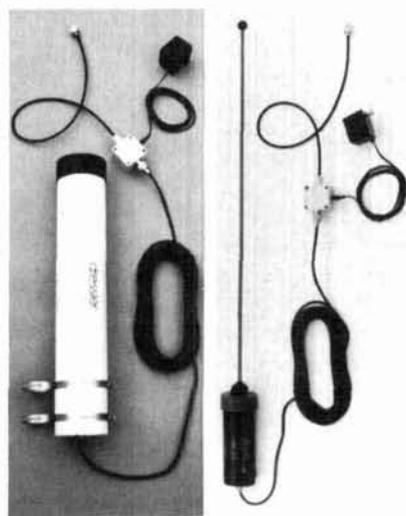
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Elmer's Notebook



Tom McMullen, W1SL

OSCILLATORS

Oscillators have been a part of radio since the use of spark transmitters. The waves spark transmitters generated weren't the clean waveforms available with modern circuitry, but they were radio waves. They weren't self-sustaining; they reached an energy peak when first started, and each successive peak was weaker (because of circuit losses) until the wave died out completely. These were known as "damped" waves.

The vacuum tube saved the day. It allowed for the addition of enough power to the circuit to overcome the losses and maintain the waveform as long as voltage was applied. Vacuum tubes were standard in many types of oscillator circuits for many years, and are still the mainstay when you need more than a few watts of RF power to drive the next stage.

Transistors are prevalent in modern equipment. These small, quiet, relatively cool devices perform well as oscillators from audio frequencies up into the microwave region. But whether the oscillator is a tube or transistor, the thing that makes it perform is feedback.

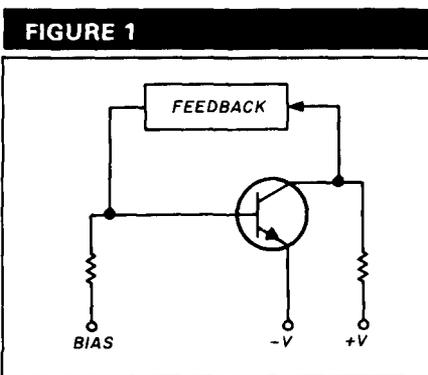
Feedback — the vital ingredient

Basically, an oscillator is an amplifier with some of its output signal fed back to the input in a proper phase relationship. If the waveform phase adds to the input waveform, the circuit will oscillate; if it tends to cancel the input waveform, it won't.

Feedback that tends to cancel the input waveform is called negative feedback, and this very characteristic has been used in many circuits to improve their operation. For example, negative feedback is used in high-quality audio

amplifiers to reduce distortion and prevent overload. In RF circuits, negative feedback is called *neutralization* and is used to *prevent* oscillation. Circuit designers can use phase-shifting networks and filters to tailor feedback for almost any situation.

Here's an example of how feedback gets things started. **Figure 1** shows a simple transistor circuit with a black



An oscillator is an amplifier with controlled feedback. Here, the "black-box" between the collector and base provides the voltage feedback and phase shift needed to keep oscillation going.

box called "feedback" connected between the output (collector) and input (base) of a transistor. (Don't worry about what's in the box right now.)

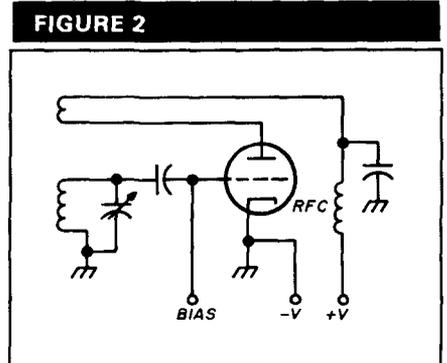
Let's say that the AC voltage on the collector reaches 10 volts, and that the transistor provides a voltage gain of 20 (that's simple voltage gain, not decibels). All the black box needs to do is supply 0.05 volts to the base. The transistor will amplify that by 20, and maintain the required 10 volts on the collector. A good design adds an extra fraction of a volt to the feedback to take transistor aging, low supply voltage, and other troubles into account — just to be safe. The phase of the voltage must be "aiding," that is, near 0 or 360 degrees. But applying too much feed-

back voltage isn't good. It can overdrive the transistor, causing the waveform to distort and create harmonics. The transistor will also draw more current than needed, and may overheat.

What's in the box?

You can obtain feedback in many ways, and electronic handbooks often show a dozen or more types of oscillators. **Figure 2** represents an old standby used for many years in various commercial and Amateur applications. It's a tube circuit with a small coil in the plate circuit located near the windings connected to the grid. This was once called a "tickler" coil.

The amount of feedback was adjusted by moving the coil closer to or further from the input circuit. The phase was determined by having the winding start and finish properly connected in the plate circuit. The old rule of thumb was: "If it doesn't oscillate, reverse the feedback windings." Some oscillators in early receivers had a feedback coil that could be rotated, letting you fine tune the strength of oscillation. (A few of these oscillators were known as "regenerative receivers.") This isn't a very elegant method of



An early "brute force" type of feedback used a plate coil located so that its magnetic field was coupled to the grid tuned circuit.

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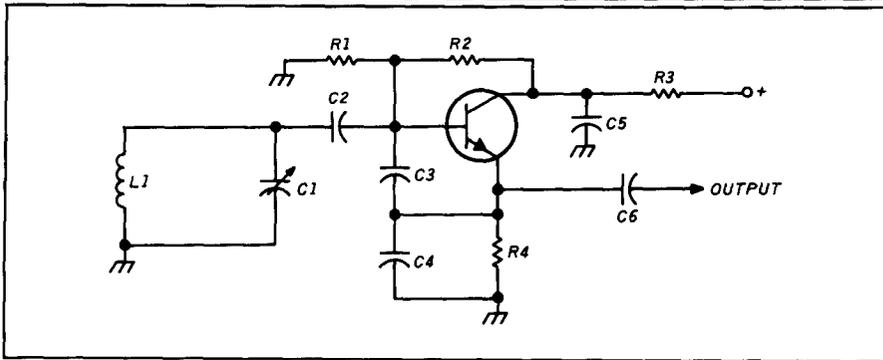
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FIGURE 3

A modern circuit uses feedback from the emitter to base of a transistor via C3. C3 and C4 form a voltage divider; R1 and R2 provide base bias. The collector is bypassed to ground for RF, and output is taken from the emitter via C6.

providing feedback, but it works. Because physical vibration and temperature cause problems in oscillators, it follows that a circuit with two coils prone to vibration and heat will be more unstable than a circuit with only one.

Figure 3 takes another approach to feedback. In a more modern transistorized circuit, a capacitive voltage divider couples some of the energy from the emitter circuit to the base circuit. The value of the capacitors from base to emitter and emitter to ground determines the strength of the feedback signal. The resonance of the tuned circuit sets the frequency of oscillation. The collector circuit is bypassed for RF, which completes the signal path back to the emitter.

You may wonder how the oscillation gets started in the first place. When power is first applied to an oscillator, the rush of current through the transistor (or tube) and the tuned circuit starts the "store energy/release energy" cycle. This cycle peaks at the circuit's resonant frequency. The action creates a small signal at the base (or grid, in a tube). Because the circuit is an amplifier, this small signal is amplified. A portion of it is then fed back to the input to further enhance the oscillation and be amplified again, and so it goes. The whole procedure requires only a few cycles to reach full strength.

Audio oscillators get started in a similar fashion, except that most depend upon a resistance/capacitance phase-shift network to determine the frequency of oscillation. The voltage change across the output load resistor creates a starting signal, which is fed back, amplified, and so on. It's possible to use a transformer to obtain

inductively coupled feedback for an audio oscillator, but why lug the heavy iron around if you can use something else?

Back to the RF circuits. Figure 4 shows two oscillators that use crystals as their frequency-determining element. In Figure 4A, the crystal is connected from base to ground. It serves the same purpose as the tuned circuit in Figure 3, by forming a parallel-

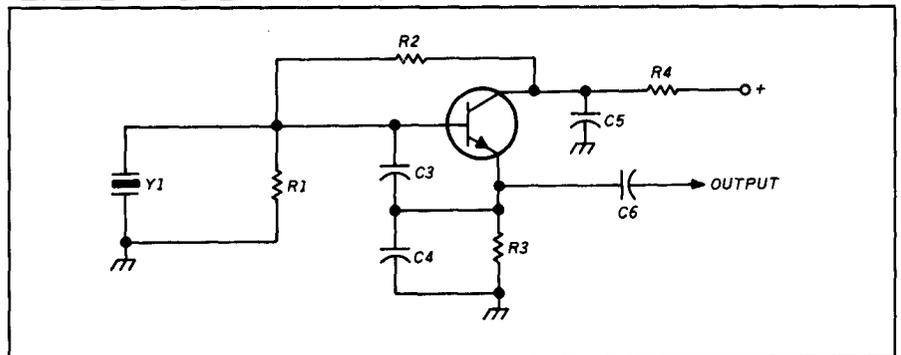
resonant circuit at the base. In Figure 4B, the crystal is a series-resonant type and provides feedback at the crystal frequency directly from collector to base.

There are many variations of these circuits; some have names associated with their inventors. Figures 3 and 4A, for example, are versions of what is often called a Colpitts oscillator; Figure 4B is a Pierce oscillator.

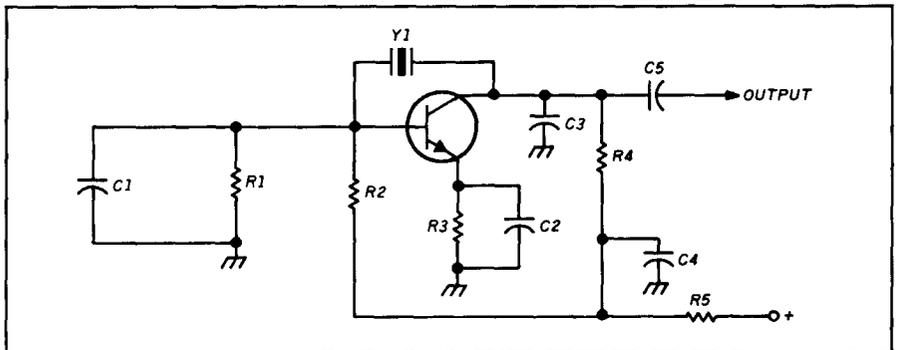
In Figure 5, you'll notice one of those voltage-variable capacitors I explained in last month's column. It looks like a cross between a capacitor and a diode, and changes its capacitance in proportion to the DC voltage applied. In the circuits used last month, I varied the negative (-) voltage applied to the diode to change the capacitance. Instead of using two supplies, one negative and one positive, this circuit simply reverses the diode and varies the positive (+) voltage to tune the circuit by means of R1.

What makes an oscillator unstable?

Frequency stability is one of the

FIGURE 4A

A Colpitts-type crystal oscillator.

FIGURE 4B

A Pierce-derivative crystal oscillator. This crystal provides feedback at its natural frequency. C1 and C3 are small-value capacitors used to stabilize the amount of feedback, preventing overdrive to the transistor and crystal.

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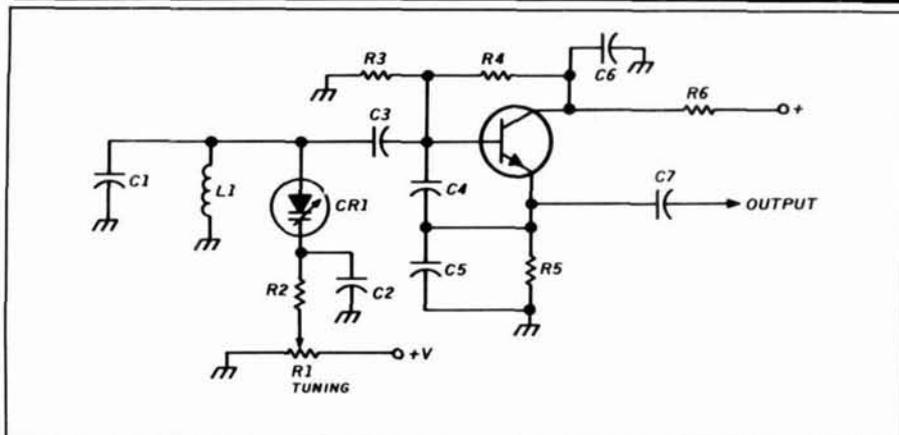
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FIGURE 5



A VFO circuit using a voltage-variable capacitor as the tuning element. This can also be called a voltage-controlled oscillator (VCO).

prime goals in designing and building an RF oscillator of any type. Mechanical stability is a close second.

It's almost impossible to design an oscillator that ignores temperature changes, but there are tricks that help you come close. Because many components cause frequency changes when the circuit gets warmer, it's possible to use temperature-compensating capacitors to provide an opposing characteristic when they warm up. If you select the proper values, the result will be an almost temperature-stable oscillator.

Of course, keeping the entire circuit away from other elements that produce heat makes your task much easier. Power supply rectifiers, transformers, and sometimes filter capacitors plus power-amplifier stages all heat up their surroundings, and you should avoid them when placing an oscillator on a chassis. Ventilate the enclosure to let heat escape or, if that isn't possible, use good heat-sink material to get the heat out of the box and into the free air where it won't harm anything. Good design also calls for keeping the current drawn by the oscillator device as low as possible. Current passing through resistors (and tubes or transistors) produces heat, so the less heat produced, the less heat there is to be dissipated.

Mechanical instability is evident in two forms. One form is called microphonics. If you tap on the enclosure and hear a "boing" when listening to the oscillator in a receiver, you have microphonics. The other form is a noise or a jump in frequency when you tune the oscillator. The

microphonic type is caused by something moving in the RF field of the oscillator-tuned circuit. It may be the walls of the enclosure, a nearby component with long leads, or even the tuned circuit components (loose windings and air-variable capacitors are prime suspects). The cure is to mount everything very rigidly, and allow space between the coil and other components so that the RF field doesn't intercept a loose or vibrating part. Heavy or rigid oscillator-enclosure walls and chassis help too.

Noise and frequency jumps are usually caused by dirty contacts in the variable capacitor of the tuned circuit. Even the voltage-variable capacitor isn't immune, because the potentiometer that changes the voltage can become noisy just as volume controls do in an audio circuit. Clean contacts are the answer for the variable capacitor. There are chemical solutions that you can use to cure the problem in both capacitors and variable resistors. If all else fails, replace the noisy part with a new one. It's worth noting that even oscillators that vary the inductance rather than the capacitance are not immune to this problem, because anything that is in the RF field of the coil can induce noise (and temperature instability) into the circuit. **EM**

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IMPEDANCE MATCHING TRANSFORMERS AND LADDER LINE

Simple schemes for matching 50-ohm coax to ladder line

By Peter H. Anderson, KZ3K, 915 Holland Road,
Bel Air, Maryland 21239

Hams enjoy experimenting with antennas. Simple wire types require a minimal investment, and the parts (consisting of nothing more than wire and insulators) are readily available. I've been working with 1:4 impedance transformers and ladder-line transmission line and would like to share what I've learned.

Introduction

A number of years ago I won my one and only door prize at a hamfest. It was an all-band "dipole-like" antenna consisting of 130 feet of wire center fed with 450-ohm transmission line. I couldn't wait to put up my new "toy" and give it a try. Imagine my disappointment when I found that it didn't work.

My preference for 450-ohm line used in conjunction with 1:4 ferrite transformers began with that antenna. I solved my impedance-matching problems using those transformers, and the antenna still works well today. I've since changed the feeds for all my other HF antennas to 450-ohm ladder line using the type of 1:4 impedance transformers shown in **Photo A**.

Ladder line has a lot to recommend it over coaxial cable. For instance, it's inexpensive and very light. Many of us depend on the smallest of twigs at the top of a tree to achieve maximum height when erecting a dipole. Anyone who has attempted to hoist the end of a dipole that's center fed with RG-8 can attest to the frustration of having those twigs break. The entire antenna falls, and you end up with a "V" shaped dipole.

Unlike coax, ladder line doesn't fill up with water, and I take a bit of comfort in being able to "see" whether the line is intact. Ladder line is also balanced and relatively loss-

less. Within hours of putting up a new antenna, I usually find myself wanting a bit more bandwidth than it can provide. Ladder line gives me flexibility in using a tuner that coaxial cable does not.

The difficulty with using ladder line seems to be in matching the 50-ohm rig to the 450-ohm transmission line and the transmission line to even the simplest of dipoles.* This drawback has become more apparent with the wide proliferation of solid-state rigs with a fixed 50-ohm output.

General analysis of 1:4 impedance transformer

Take a look at the general 1:4 impedance-matching autotransformer in **Figure 1**. It includes some practical implementations. Note that the output ("b" side) may be balanced or unbalanced depending on where ground is supplied on the input ("a" side).

A rigorous analysis of the network, which assumes only that the two coils are tightly coupled with one another, results in the following expression:**

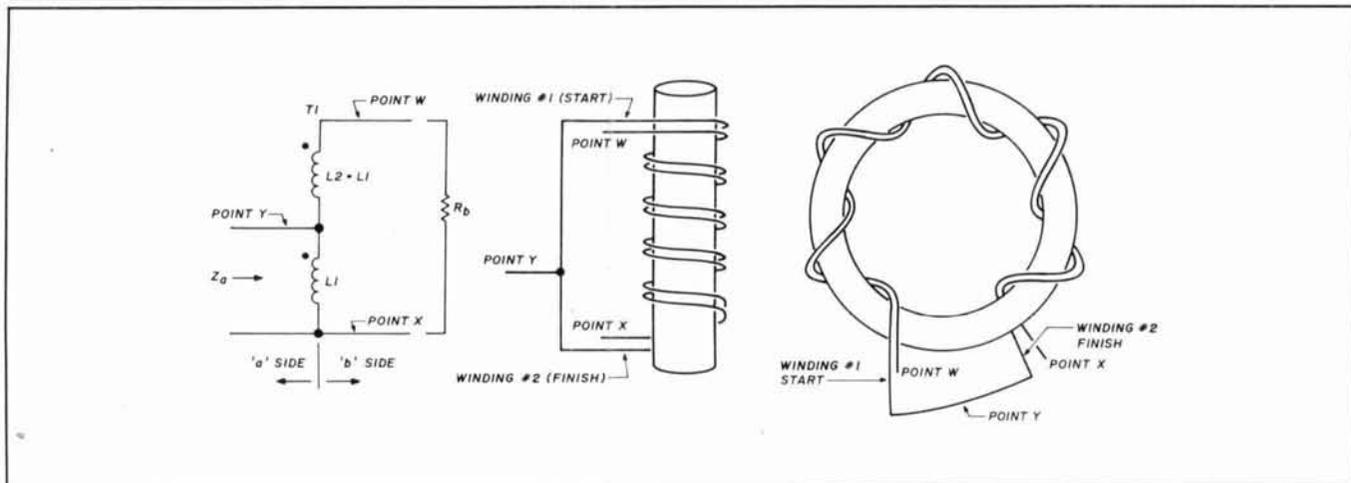
$$Z_a = \frac{R_b}{4.0 - j \frac{R_b}{X_L}} \quad (1)$$

The j operator in the denominator indicates that the second term adds to the first at right angles, as shown in **Figure 2**. Note that if $2\pi fL_1$ (the inductive reactance) is "large"

*Of course, the actual impedance ratio for transforming 50-ohm coax to ladder line is 1:9. I was unsuccessful in winding a 1:9 transformer that would provide satisfactory results. If you're interested in trying, you could use a pair of 1:3 transformers at either end to provide a closer 1:9 impedance transformation. Because of the low loss nature of 450-ohm ladder line, the additional loss due to impedance mismatch has negligible effects on the total power transformer.

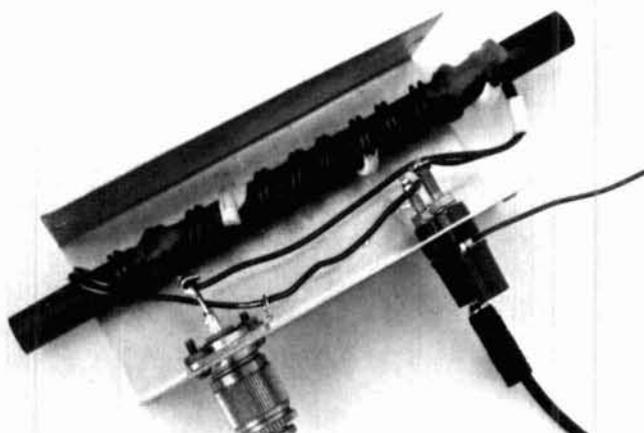
**Detailed derivations are available from the author for an SASE with one unit of postage. Computer analysis routines were written in Turbo Pascal, Version 4.0 for IBM compatibles; a complete library of source code is available on 5-1/4" diskette for \$5.00. You must have Turbo Pascal Version 4.0 (or later) and Turbo Graphix to support Version 4.0 to compile and execute these files. Hard copies of the source code listings are available for an SASE with three units of postage.

FIGURE 1



Schematic representation of 1:4 impedance transformer and implementations using a rod and toroid. The dots refer to the relative positioning of the windings. Each bifilar turn consists of two conductors tightly coupled to one another.

PHOTO A



An implementation of a 1:4 transformer using a ferrite rod. The design shown is unbalanced; ground is applied to one side of the transformer versus the center. Small applications of epoxy hold the rod in place.

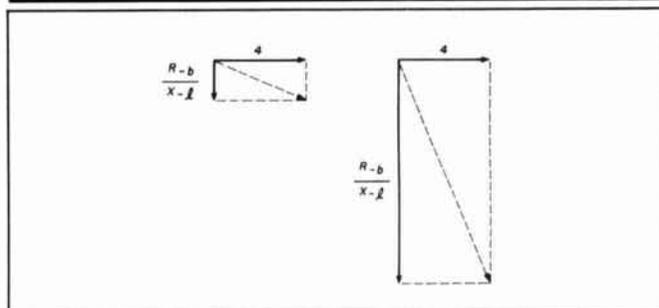
in relation to R_b , the first term in the denominator swamps the second and the expression reduces to:

$$Z_a = R_b/4 \quad (2)$$

The input impedance is $R_b/4$ and is purely resistive. The fact that the inductive reactance is large in relation to R_b is particularly important. It raises questions about how large "large" is, and about what happens if the inductive reactance isn't "large" enough.

Figures 3A and B provide a graphical analysis of Equation 1. Figure 3A shows that the ratio of the resistive portion of Z_a to R_b approaches 0.25 as the frequency increases for all values of R_b divided by L_1 — that is, as the inductive reactance becomes "large" in relation to R_b . You'll note that at a particular frequency the ratio approaches 0.25 as R_b decreases in relation to L_1 , or as L_1 increases in relation to R_b . Figure 3B is interesting because it shows that the

FIGURE 2



The j operator indicates the quantities add at right angles. If R_b/X_l is small in relation to 4.0, the denominator is close to 4.0. However, if R_b is large in relation to 4.0, the denominator is R_b/X_l .

reactive component seen on the "a" side may either increase or decrease with increasing frequency. Note that the reactive component is always positive (inductive).

Analysis of an inadequate design

The following example shows how to use Figures 3A and B to analyze the performance of an impedance transformer.

As one of my earliest projects, I attempted to match my 50-ohm transmitter to an end-fed wire which was about 135 ohms resistive at 3.5 MHz (see Figure 4). I reasoned that $135/4$ wasn't 50, but it was close enough. I wasn't prepared for the 10 ohms plus the substantial reactive component which resulted, and I dropped the project in frustration.

I used the T-200-2 iron powder core (usually red in color).^{*} It's probably the most common toroid-type core used at high HF power levels by the Amateur fraternity.

Consider a nine-bifilar turn arrangement at 3.5 MHz:

$$Ll \text{ (in } \mu H) = (\text{turns}/100)^2 \times Al \quad (3)$$

where Al is in μH per 100 turns.

^{*}The ARRL Handbook discusses toroid inductors in its "Electrical Fundamentals" chapter. It doesn't deal specifically with ferrite rods. The formulas for calculating inductance and turns are the same as the ferrite toroids max. 61, except that Al is 49.

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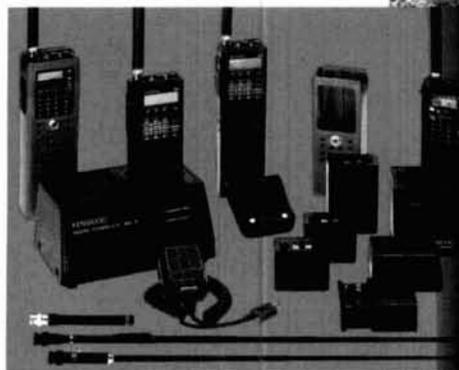
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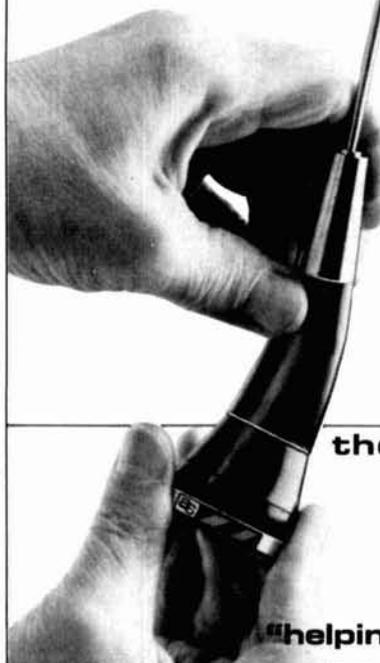


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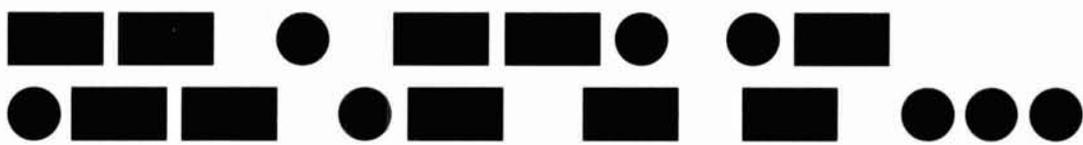
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$$Ll \text{ (in } \mu\text{H)} = (9/100)^2 \times 120 = 0.972 \mu\text{H} \quad (4)$$

Therefore, the inductive reactance at 3.5 MHz is:

$$Xl = 2\pi \times 3.5E \times 10^6 \times 0.972E \times 10^{-6} = 21.4 \text{ ohms} \quad (5)$$

This falls far short of the requirement that Xl be "large" in relation to R_b , which in this case was 135 ohms.

You can quickly calculate the resistive component of Z_a seen by the transmitter by referring to **Figure 3A**. R_b/L_1 is nominally 138 ohms per μH . Using the curve for 150 ohms per μH as a guide, you can interpolate that the resistive part of Z_a/R_b at 3.5 MHz is nominally 0.075. This makes the resistive ratio of Z_a to R_b 1:13 — a far cry from 1:4. Additionally, the input resistance seen by the transmitter is close to 10 ohms.

Find the reactive component using **Figure 3B**. By picturing a curve between $R_b/L_1=100$ and 200 you can estimate the reactive component ratio to be nominally 0.1. That is, $0.1 \times 135 \text{ ohms} = 13.5 \text{ ohms}$. This means that at 3.5 MHz, the 135-ohm resistance is "reflected" to the transmitter as 10 ohms resistive plus 13 ohms reactive. Note that the reactive element is inductive. All in all, mine was not a successful venture.

Developing a workable design for an end-fed wire

Your next step is to design a functional transformer for this antenna. According to **Figure 3A**, you need an R_b/L_1 ratio of less than 15 ohms per μH to get close to a 1:4 impedance ratio at 3.5 MHz. Because R_b is equal to 135 ohms, L_1 must be larger than $135/15=9 \mu\text{H}$. **Figure 3B** shows that the reactive component at 3.5 MHz is $0.04 \times 135 = 5.4 \text{ ohms}$, which is probably acceptable in relation to 33 ohms resistive. Use the relationship for iron powder toroids to find the number of turns required:

$$\begin{aligned} \text{Turns} &= 100 \times \sqrt{(\text{desired } Ll \text{ in } \mu\text{H}/Al)} \\ &= 100 \times \sqrt{(9/120)} \\ &= 27.4 \text{ turns} \end{aligned} \quad (6)$$

Note that each turn is bifilar consisting of two wires. Getting 55 single turns of no. 12 or 14 enameled wire on a T-200 core isn't possible. My own experience is that 15 bifilar turns is a practical maximum.

If you are confronted with the same problem at 7.0 MHz, look at **Figure 3A** and you'll see that an R_b/L_1 of 25 ohms per μH is sufficient. Consequently, L_1 must be a minimum of $135/25=5.4 \mu\text{H}$. I calculated the number of bifilar turns to be 21.

Now that I've gone through an analysis and design example, I'd like to offer some generalizations:

On 3.5 MHz a ratio of R_b/L_1 of 15 ohms per μH is adequate. At 7 and 14 MHz, ratios of 25 and 50 are adequate. Clearly, a broadband transformer designed using these criteria for one band will also perform on the higher ones.

So far I've considered an R_b of 135 ohms. We hams are usually interested in transforming R_b values on the order of several hundred ohms, and I concluded that the T-200-2 is very difficult to use on the lower HF bands. More than 21 bifilar turns simply don't seem to fit and I'm suspicious of parasitic capacitance; however, there's an easier alternative.

FIGURE 3A

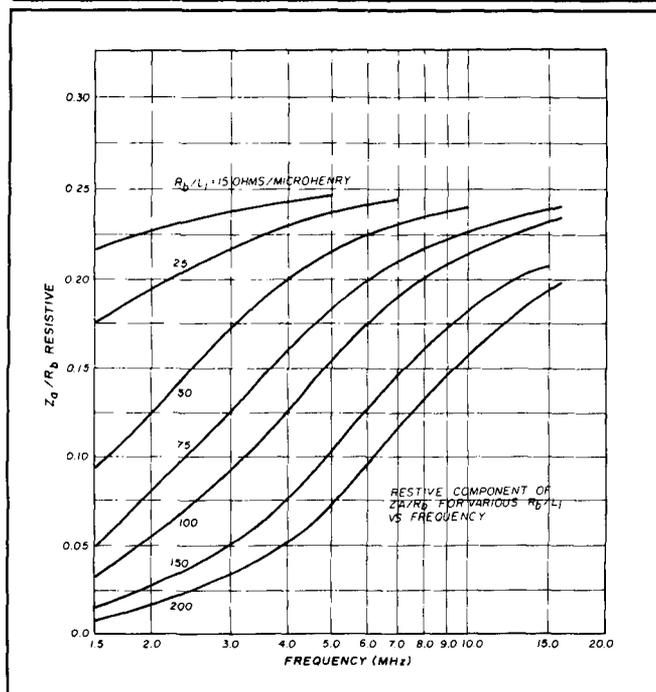
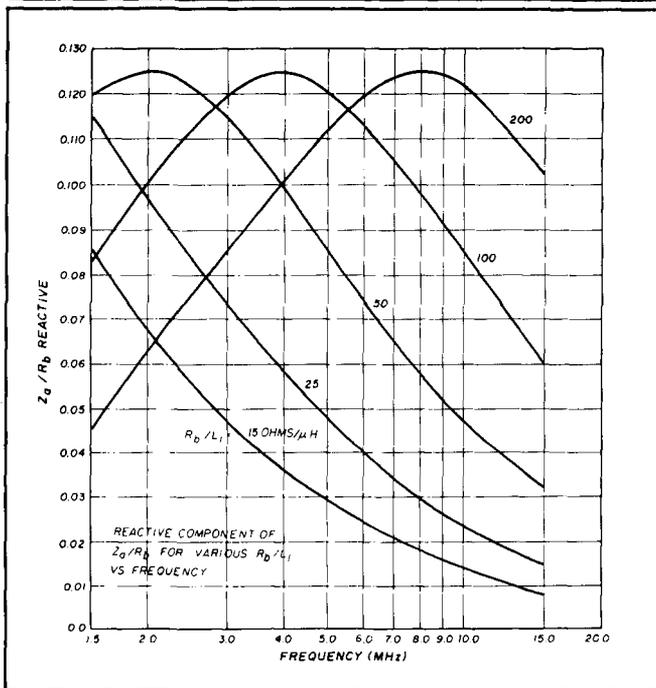


FIGURE 3B



Graphical analysis of Equation 1.

want to experiment with a toroid, the FT-240-61 might be a good prospect.

Consider a 13-bifilar turn arrangement at 3.5 MHz:

$$Ll \text{ (in mH)} = (\text{turns}/1000)^2 \times Al \quad (7)$$

Where Al is in mH per 1000 turns.

$$Ll \text{ (in mH)} = (13/1000)^2 \times 49 = 8.28 \mu\text{H} \quad (8)$$

Assuming R_b is 135, the R_b/L_1 ratio is 16.3 ohms per μH . Using the $R_b/L_1=15$ ohms per μH curve in **Figure 3A** as

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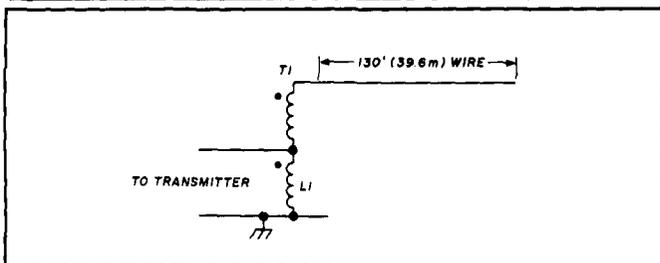
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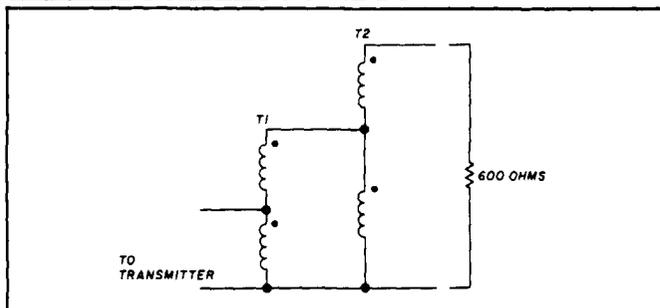
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FIGURE 4



Using a noise bridge I determined the impedance of a 130-foot ended wire was nominally 135 ohms resistive. The transformer reflected this to present a $135/4 = 34$ ohms load to the transmitter.

FIGURE 5



For large impedances two 1:4 transformers may be cascaded to realize a 1:16 impedance transformer. However, the inductive reactance of T2 must be large in relation to the load impedance. (See text.)

Ferrite toroids appear to be the answer. Unfortunately, they're relatively expensive and more susceptible to core saturation. I focused on the R61-050-750 ferrite rod; if you want to experiment with a toroid, the FT-240-61 might be a good prospect.

Consider a 13-bifilar turn arrangement at 3.5 MHz:

$$Ll \text{ (in mH)} = (\text{turns}/1000)^2 \times Al \tag{7}$$

Where Al is in mH per 1000 turns.

$$Ll \text{ (in mH)} = (13/1000)^2 \times 49 = 8.28 \mu\text{H} \tag{8}$$

Assuming R_b is 135, the $R_b/L1$ ratio is 16.3 ohms per μH . Using the $R_b/L1=15$ ohms per μH curve in Figure 3A as a guide, you can calculate the resistive Z_a/R_b ratio at 3.5 MHz to be 0.23. This means the resistive portion of Z_a seen by the transmitter is $0.23 \times 135 = 33$ ohms, very close to $135/4$. The reactive component, taken from Figure 3B, is less than 5 ohms.

As the frequency increases, the $2\pi fL1$ becomes increasingly larger in relation to the Z_b of 135 ohms, and the transformer operates correctly as a 1:4 transformer. The finished product is shown in Photo A.

Using cascaded 1:4 transformers for matching to larger impedances

You can modify this basic approach to match impedances exceeding 400 ohms to a nominal 50 ohms. See Figure 5 for the circuit.

Consider a 600-ohm feedpoint impedance. A 1:16 impedance transformer consisting of two cascaded 1:4 trans-

formers changes the impedance to $600/16=37.5$ ohms, which is "close" to 50 ohms. Note that the R_b seen by the second transformer is 600 ohms. To meet a requirement of $R_b/L1$ greater than 15, $L1=600/15=40 \mu\text{H}$ (or 0.04 mH). The number of bifilar turns on a ferrite rod are calculated by:

$$\begin{aligned} \text{Turns} &= 1000 \times \sqrt{(0.04/49)} \\ &= 28.5 \text{ bifilar turns (57 conductors)} \end{aligned} \tag{9}$$

This is tight, but not impractical. But the 58 bifilar turns (116 conductors) that would be required on a T-200-2 powder iron toroid, are simply not possible.

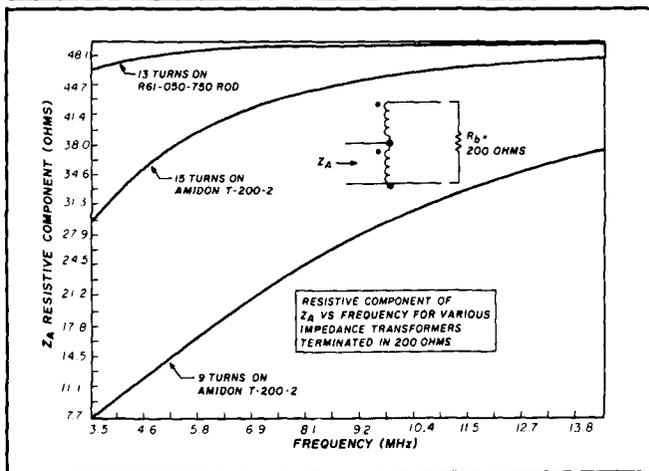
Note that the first transformer sees an R_b of $600/4=150$. This transformer can be implemented simply by using the 13 turns on a ferrite rod discussed above.

Observations

Plots of the resistance and the reactance seen on the "a" side for a number of different 1:4 transformers terminated in 200 ohms are shown in Figures 6 and 7. Using a noise bridge, I found that actual resistance measurements agreed closely with this theoretical performance.

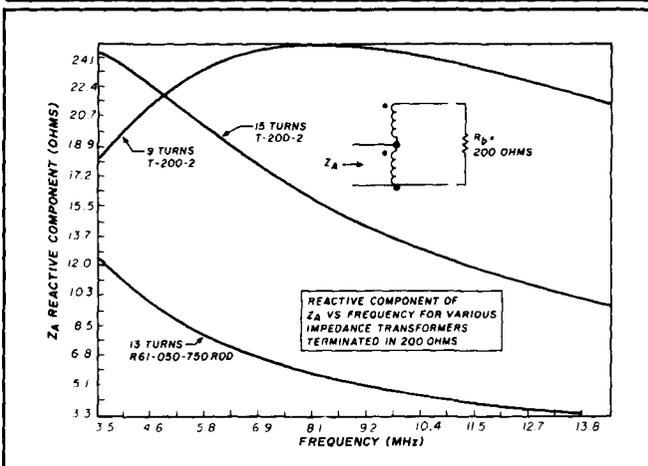
The results speak for themselves. If you want to match several hundred ohms on the 40 and 80-meter bands,

FIGURE 6



Resistive component response for various transformer materials.

FIGURE 7



Reactive component response for various transformer materials.

remember that the T-200-2 is not a Zb/4 transformer. The resistive portion of the impedance seen on the "a" side isn't anywhere close to one quarter and there's a substantial reactive component. On the other hand, the ferrite rod gives pretty good performance as a 1:4 arrangement across the HF spectrum from 80 meters up, and it might prove usable on 160 meters.

This doesn't rule out the T-200-2 as a matching transformer. You could null out the reactive component with a series capacitor, and use the "1:4 transformer" to achieve the ratio you want by controlling the number of turns. For example, if you have a 300-ohm termination that you want to reflect as 50 ohms at 7.0 MHz, $50/300=0.167$. Using **Figure 3A**, locate the point at X coordinate (7 MHz) and Y coordinate (0.167). I estimate that an Rb/L1 of nominally 130 ohms per μH passes through this point. Therefore $L1=Rb/130=300/130=2.3 \mu\text{H}$. Using a T-200-2 iron powder toroid:

$$\begin{aligned} \text{Turns} &= 100 \times \sqrt{(2.3/120)} \\ &= 13.8 \end{aligned} \quad (10)$$

To determine the reactive component use **Figure 3B**. At 7.0 MHz and $Rb/L1=130$, the reactive Za/Rb is nominally 0.11. Therefore, $Zb \text{ reactive}=0.110 \times 300=33$ ohms. This inductive component might be nulled by using a series capacitor on the "a" side: $C=1/(2\pi \times 7 \times 10^6 \times 33)=688$ pF.

Such an arrangement will probably work well across the 40-meter band. However, it won't work on higher frequency bands. You're simply using the deficiency of the 1:4 arrangement to obtain a 1:6 match at a specific frequency.

Use of 1:4 transformers with ladder line

Using my new-found knowledge, I set up the circuit in **Figure 8**. I put two 1:4 transformers (like the ones mentioned above) back to back. A 100' length of 450-ohm ladder line stretching out the back door, around the house, in the front door, and back to the shack provided me with access to both the "transmit" and "antenna" sides for power measurements.

Theoretically, the 50-ohm dummy load is reflected (transformed) as 200 ohms, and on the other side of the line the 200 ohms is reflected as 50 ohms. This results in a relatively high VSWR on the transmission line, but relatively little power dissipation.

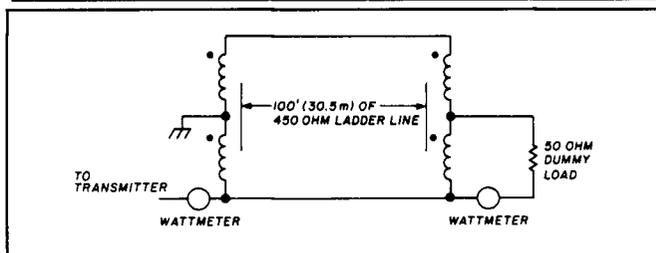
The results seemed too good to be true. The loss in the transmission line from the low end of 80 meters through the high end of 20 meters was less than 10 percent — including the two transformers. Viewed at the transmitter side of the arrangement, the SWR was better than nominally 2:1. On 160 meters the SWR was high, but correctable with a tuner. Even on 160 there was relatively little transmission line loss! Results on the 15 and 10-meter bands were difficult to interpret; this might be due to a less than ideal dummy load.

I have subsequently run tests on various lengths of ladder line and seem to be able to repeat my results. This leads me to conclude that most of the lost power is dissipated in the transformers. However, the efficiency of each transformer appears to be 95 percent.

Putting it all together

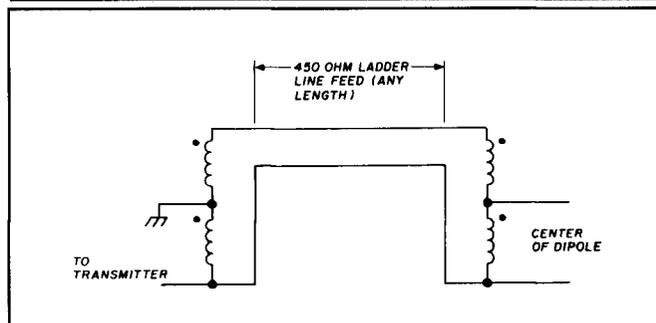
My next step was to apply this "discovery" to feeding dipole antennas normally fed with 50-ohm coaxial cable.

FIGURE 8



Two 1:4 transformers were connected back to back on the two sides of a 100-foot length of 450 ladder line. Ninety percent of the power delivered by the transmitter with a 50-ohm output impedance was delivered to the dummy load for 160 through 20 meters.

FIGURE 9



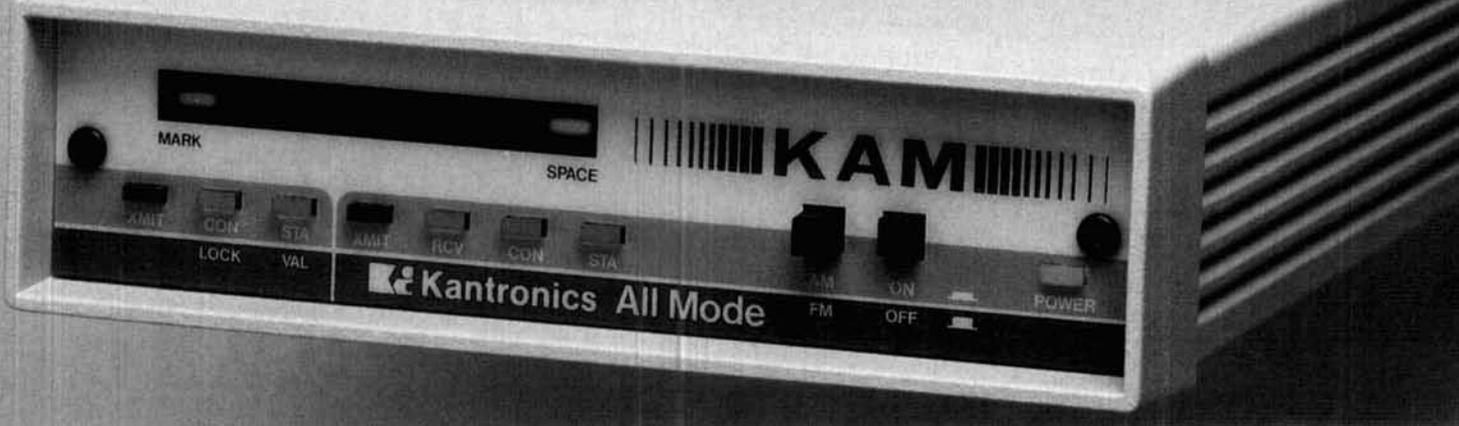
It appears feasible to feed 50-ohm loads using any length of ladder line terminated at each end with a 1:4 ferrite rod transformer over the HF bands from 80 meters and up. Note that the ladder line is balanced, transmitter ground is applied at the center of the transformer.

My previous attempts to feed dipoles with ladder line had drawbacks. One of my techniques involved terminating the ladder line directly at the center of the antenna, and on the transmitter side using a tuner capable of assuming different configurations. This let me load the antenna with "brute force."* The technique worked well, but I'm a firm believer in having a lot of antennas and I can't afford the space or money necessary to devote a tuner of this type to each antenna. My other technique was to use a delta match consisting of wires fanning down to meet the transmission line. I never had much luck with this method, and the family was forever getting entangled in wires that fell from the sky.

The arrangement in **Figure 9** has proven successful in feeding simple antennas without a tuning arrangement. The 1:4 transformer between the rig and the transmission line is an unbalanced-to-balanced arrangement, and ground is fed to the center of the transformer. The second transformer is mounted at the center of the dipole.

The results have been good. I've used this approach with standard dipoles on 80, 40, and 20 meters. The 40-meter dipole also performed well on 15 meters. **Figure 10** shows summarized results. In all cases except on 80 meters the SWR was better than 2:1 across the entire band. Using a very simple fixed series L/fixed shunt C tuner arrangement, I was able to obtain a match of close to 2:1 across the entire 80-meter band.

*Both matching techniques are discussed in detail in Doug DeMaw's book *W1FB's Antenna Notebook*, ARRL, 1987.



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14

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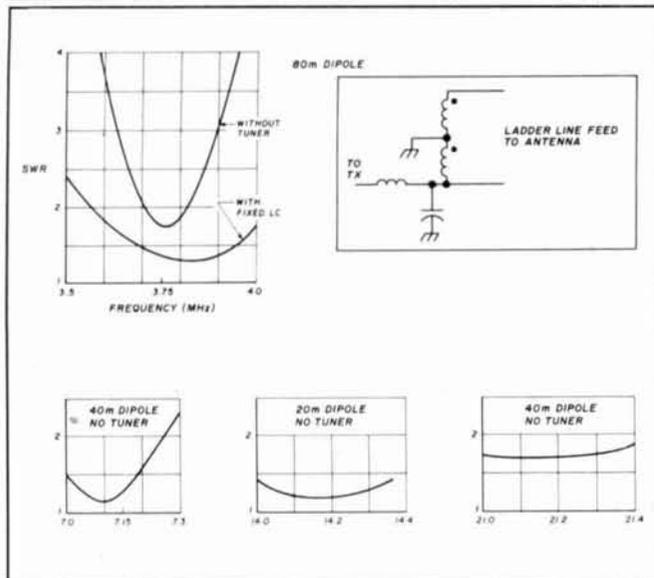


16

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FIGURE 10



Three separate dipoles (80, 40, and 20 meters) were erected, each with separate feeds consisting of ladder line terminated with ferrite rod transformers. The 40-meter dipole was also used on 15 meters. Performance on 40, 20, and 15 was excellent with an SWR consistently less than 2:1 across the entire band. No tuner was required. On 80 meters, the bandwidth was relatively narrow and a very simple tuner consisting of a series L, shunt C was inserted on the transmitter side of the transformer.

One of the beauties of ladder line is that it's relatively lossless. Most of us don't have enough property to run 500 feet from the shack to that ideal antenna location — but some may, and this is the way to go.

Many hams want to ragchew on 3.9 MHz and also work CW or the DX window on the low end of the band using the same antenna. This poses a problem when feeding an 80-meter dipole with coax. You must choose a compromise resonant frequency and use a tuner for frequencies far removed from it. You can measure full power at the tuner, but the loss in the coaxial feed will be high. By using the ladder line in conjunction with the 1:4 transformers at each end, you can use a tuner on the transmitter side of the transformer and have confidence that most of the power is reaching the antenna.

I've run all my dipoles at 700-watts output CW for several months with no evidence of flashover, saturation, or core heating. Even so, use some care. Attempting to load an 80-meter dipole on 40 meters presents the transformer with an impedance of several thousand ohms. This results in very high potentials, which may saturate the core and cause the windings to flash over.

Construction notes

The design which uses 13 turns on a ferrite rod has become my standard 1:4 transformer for impedances less than 400 ohms. The designs intended for indoor use may be either self-supporting or mounted on any miscellaneous mini box. The outdoor design is packaged in 3/4" PVC pipe. I used epoxy to secure the end caps and sealed the seams externally with RTV silicone. I fastened the entire assembly to a conventional ceramic insulator using cable ties.

Electrical parts are relatively easy to find. But I would stay away from miscellaneous toroids sold at hamfests. The computer industry manufactures many types of toroids to suppress the electromagnetic interference caused by high-speed switching noise, and these seem to have flooded the market. They are not suitable for this transformer application. There appears to be no standard color coding of ferrite toroids; unless you have a lot of faith in the seller, I wouldn't waste my time.

Three *Ham Radio* advertisers carry toroids and rods: Amidon, Palomar, and Radiokit. Amidon and Radiokit also carry the Scotch no. 27 glass tape I used to bind the two conductors tightly together. They also have a complete line of heavy Thermaleze™ insulated wire; I used 14 gauge.

Summary

I've provided a way to design and analyze 1:4 impedance transformers and given examples of some successful designs. I've had great success feeding dipoles with ladder line in the same way I'd feed them using coaxial cable. The materials are inexpensive and readily available. You don't need special tools, exotic metalwork, complex instrumentation, or large amounts of money. Give these designs a try, extend them to other types of antennas, and let me know your results. 

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CT-125	10 Hz-1.25 GHz	< 25mV @ 50 MHz < 15mV @ 500 MHz < 100mV @ 800 MHz	1 PPM	9	0.1 Hz, 1Hz, 10Hz	189.95
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A NOVEL METHOD FOR MEASURING CABLE ATTENUATION

By A. E. Popodi, OE2APM/AA3K, Moosstrasse 7, A-5020 Salzburg, Austria

The simplest way to determine cable loss is to measure the difference between input and output power with an RF wattmeter. A more accurate method is to measure the cable input and output voltages with an RF voltmeter and calculate the cable loss in nepers from the formula $\alpha = \log_E \frac{V_{in}}{V_L}$ where V_{in} is the input voltage and V_L

the output (load) voltage. Both methods require a power source with an output impedance matching that of the cable and a known characteristic impedance Z_0 . Accuracy suffers if there is a matching error. Ground currents in the test setup can cause false readings. This method becomes very unreliable if α is small (i.e., the cable is short). If you want to measure the loss of a 2-meter length of RG-213 cable at a frequency of 28 MHz using this method, you'll find that the input and output voltage differ by only 0.89 percent. It's very difficult to make credible and meaningful measurements on short cables because of this small input/output voltage differential.

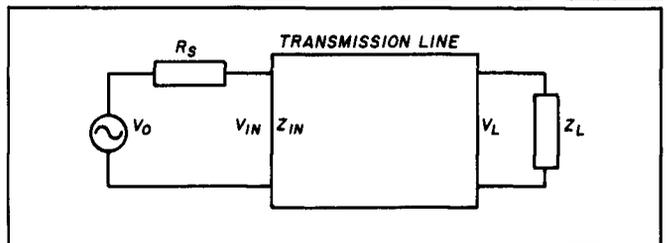
The method I've developed doesn't suffer from this shortcoming. It's based on the unique properties of an unterminated, quarter-wavelength transmission line driven from a low-impedance source. (See Figure 1.)

At the quarter-wavelength frequency, the cable input impedance (Z_{in}) is very low — typically less than 1 ohm. The input voltage (V_{in}) is at minimum and the cable output voltage (V_L) is high. Resistor R_s is the source impedance of the voltage source V_0 .

The general expression for the ratio of input to output voltage is given in Equation 1:

$$\frac{V_{in}}{V_L} = \cosh(\alpha + j\beta) + \frac{Z_0}{Z_L} \sinh(\alpha + j\beta) \quad (1)$$

FIGURE 1



Block diagram of the setup for measuring cable attenuation.

where Z_0 is the characteristic impedance, Z_L is the load impedance, α is the cable attenuation in nepers, and β is the cable length in radians — in our case $\frac{\pi}{2}$. For the ideal case of the unterminated cable, this expression reduces to:

$$\frac{V_{in}}{V_L} = \cosh\left(\alpha + j\frac{\pi}{2}\right)$$

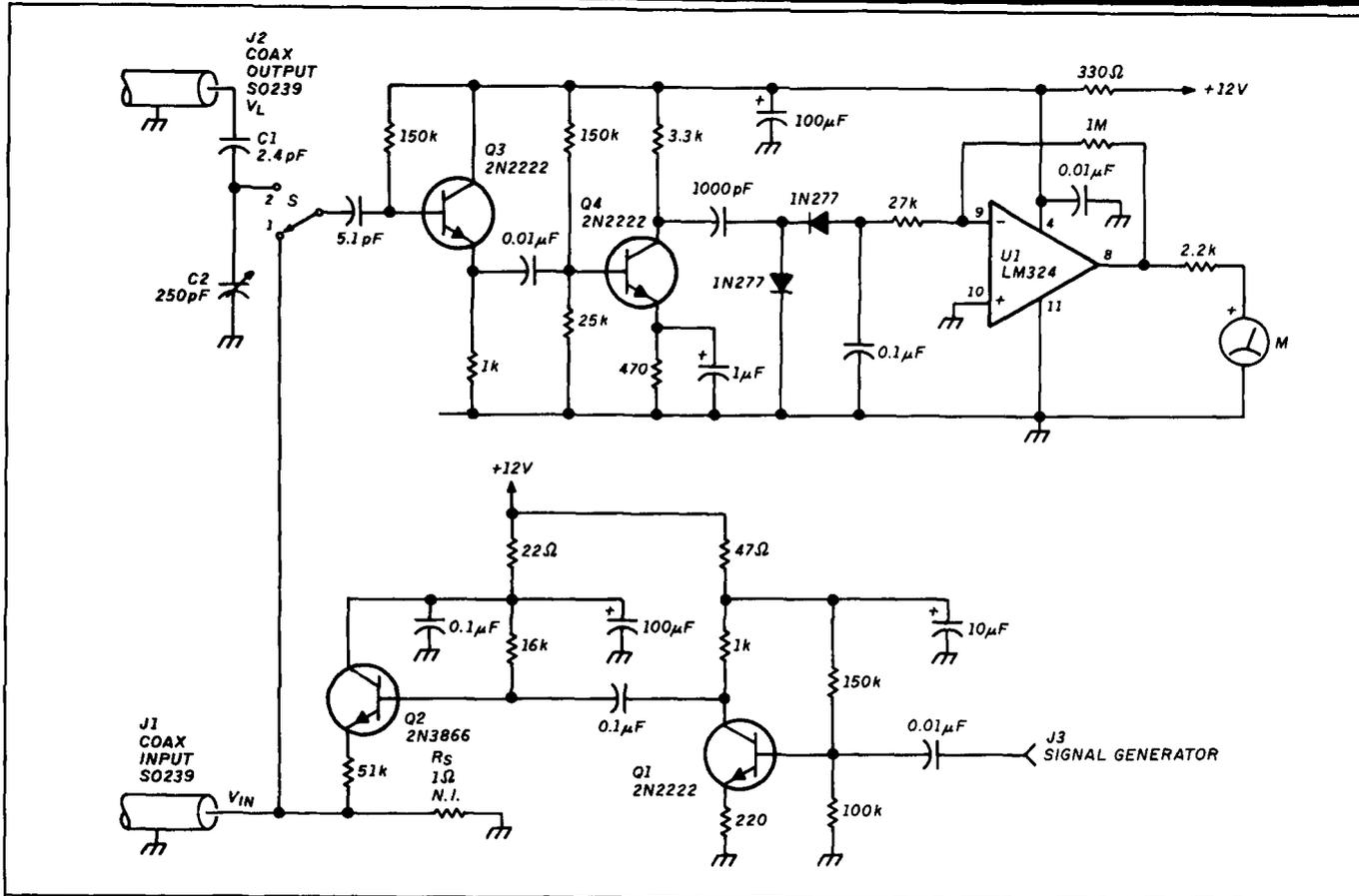
Using the formula $\cosh(\alpha + j\beta) = \cosh \alpha \cos \beta + j \sinh \alpha \sin \beta$ we have:

$$\frac{V_{in}}{V_L} = j \sinh \alpha \quad (2)$$

Since α is a small number, $\sinh \alpha$ is nearly equal to α . (For example, if $\alpha = 0.147$ neper, the error caused by this simplification is only 0.5 percent.) The operator (j) indicates a 90-degree phase shift between the two signals. Using absolute values, you have:

$$\frac{V_{in}}{V_L} = \alpha \text{ in nepers} \quad 1 \text{ neper} = 8.6859 \text{ dB.} \quad (3)$$

FIGURE 2



Schematic of the complete test circuit for determining the voltage ratio for accurately determining the cable attenuation.

The cable attenuation in nepers is very nearly equal to the ratio of the two voltages. As a result, V_L may be 50 times larger than V_{in} , depending on cable loss.

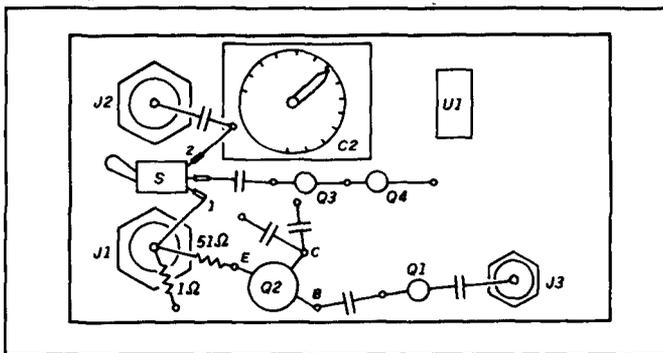
Equation 2 is only valid if load impedance Z_L is infinitely high. In practice, you have a capacitive load impedance (Z_L) and a slightly inductive source impedance (R_s). That means you must use Equations 1 and 6 to evaluate the effect on the voltage ratio $\frac{V_{in}}{V_L}$.

You don't need a precision RF voltmeter to measure this voltage ratio if you use a capacitive voltage divider for V_L and compare the divided voltage with V_{in} in a simple voltage comparator. This gives you the advantage of being able to make Z_{in} very high. Figure 2 shows the schematic of the complete test circuit.

A signal generator is connected to the input of isolation amplifier Q1 that feeds the emitter follower and cable driver Q2. Resistor R_s is a noninductive 1-ohm resistor that presents a low source impedance to the cable input at terminal J1. At the cable output (J2), there is a capacitive voltage divider consisting of a small 2.4-pF fixed capacitor (C1) and a 250-pF variable capacitor (C2). Switch S alternately connects V_{in} (position 1) or the divided voltage V_L (position 2) to a post-amplifier (consisting of emitter follower Q3 and amplifier Q4) with high input impedance. The rectified RF signal is amplified in operational amplifier U1 that drives indicator instrument M.

You must carefully choose the biasing of transistor Q1

FIGURE 3



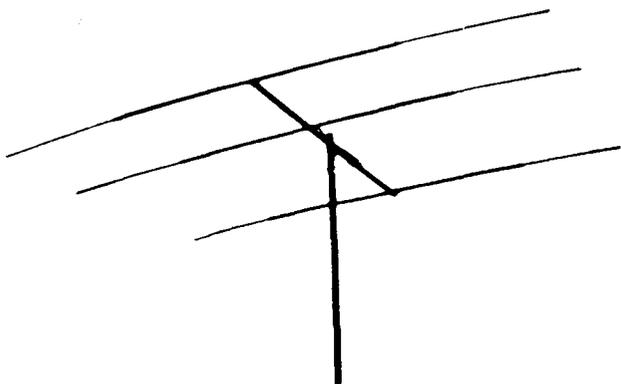
Component layout of the cable-attenuation test circuit.

and emitter follower Q2 to avoid waveform distortion in Q2. Because the cable output waveform is always better than a possibly distorted input signal, voltage comparison may be degraded due to waveform differences between the two signals. Figure 3 shows a recommended layout for the test fixture.

Switch S is mounted between connectors J1 and J2. (I recommend using a silver-plated brass plate as the mounting surface in order to reduce the effect of ground loops.) It's important that the lead lengths to and from capacitor C1 be as short as possible to obtain a frequency-independent capacitive voltage divider.

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- F/B.....20 dB
- Feed Imp.....50 Ohms
- Balun.....4:1, 5kW

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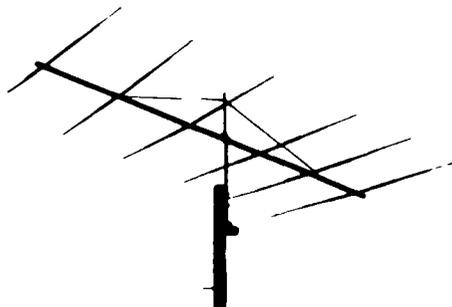
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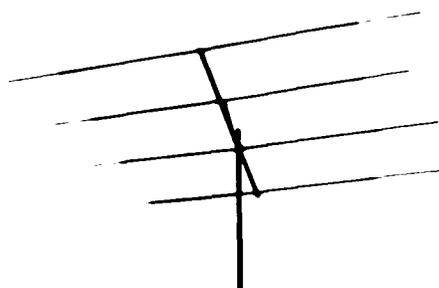
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- Bandwidth... 28-30 MHz
x 1 MHz
- Gain..... 11 dBd
- VSWR..... 1.5:1
- F/B..... 30 dB
- Feed Imp..... 50 Ohms
- Balun..... 4:1, 5 kW

MECHANICAL

- Element Length... 18 ft.
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- Turn Radius. 16.5 ft., 8 in.
- Windload..... 4 sq. ft.
- Weight..... 29 lbs.
- Mast..... 2 in.

10M-4



ELECTRICAL

- Bandwidth... 28-30 MHz
x 1 MHz
- Gain..... 7.7 dBd
- VSWR..... 1.5:1
- F/B..... 25 dB
- Feed Imp..... 50 Ohms
- Balun..... 4:1, 5 kW

MECHANICAL

- Element Length... 18 ft.
- Boom Length..... 10 ft.
- Turn Radius..... 10.5 ft.
- Windload..... 2.25 sq. ft.
- Weight..... 12 lbs.
- Mast..... 2 in.

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x 250 kHz
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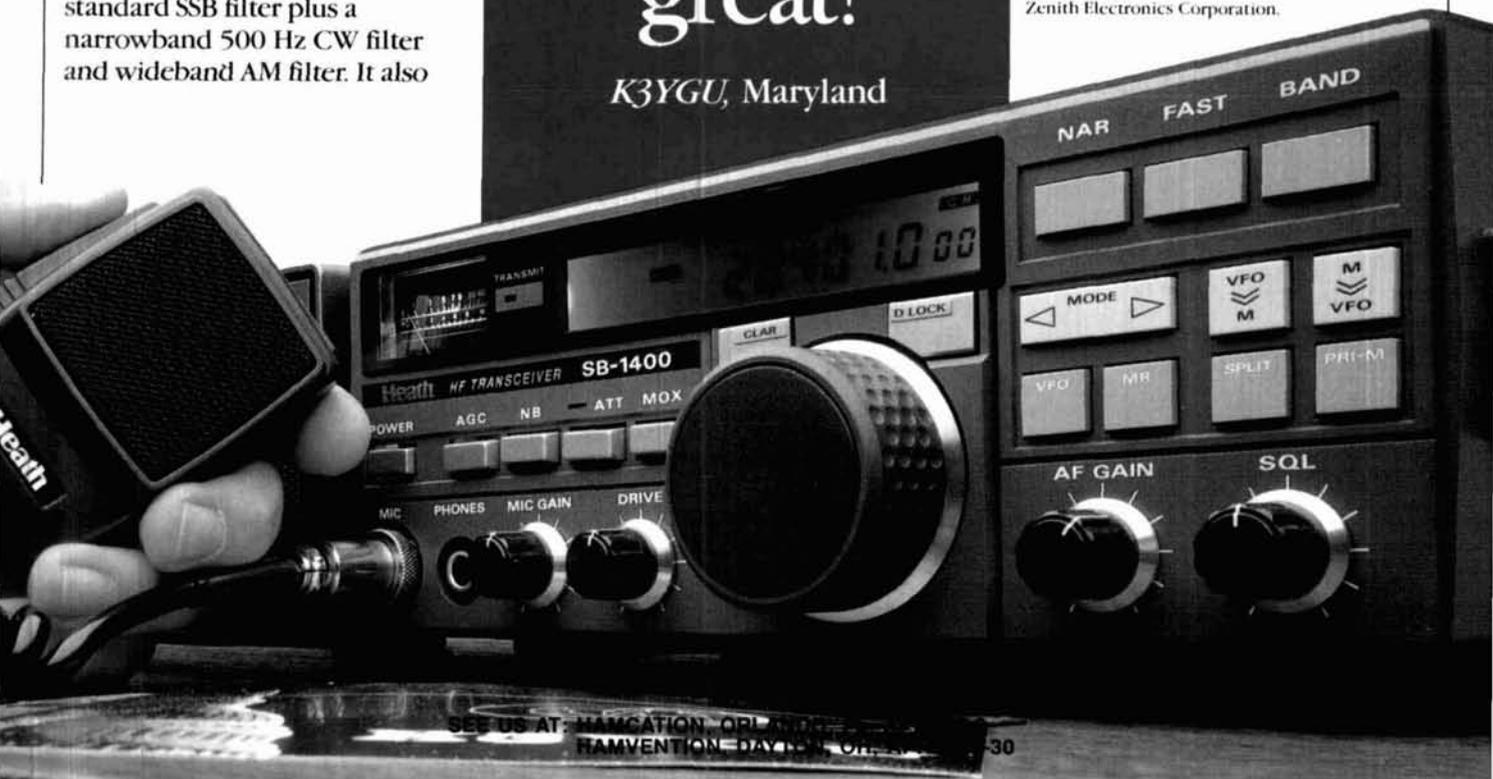
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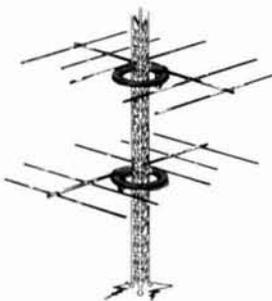
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Test procedure

Connect the cable to J1 and J2. Put switch S in position 1 and adjust the signal generator near the estimated quarter-wave frequency for an input voltage minimum, with a meter deflection of perhaps 80 percent. With the switch in position 2, adjust variable capacitor C2 for equal meter deflection. Capacitor C2 must be measured in the circuit with a capacitance bridge, with capacitor C1 disconnected, and with switch S in position 2. You can calibrate the dial of C2 directly in nepers or picofarads. Calculate the divider ratio from Figure 4. If, for instance, C2 = 153 pF and C1 = 2.4 pF, the voltage ratio is:

$$\frac{V_L}{V_{in}} = \frac{153}{2.4} + 1 = 64.75$$

and the cable loss is:

$$\alpha = \frac{1}{64.75} = 0.0154 \text{ neper or } 0.133 \text{ dB}$$

How to find cable loss at other frequencies

Unfortunately, my method supplies cable loss for only one frequency. But if you make a second measurement at three times the frequency (pertaining to a three-quarter wavelength cable), you now have two sets of attenuations α_1 and α_2 and two frequencies F1 and F2. Using the interpolation method described by K2BT,¹ you can calculate the loss at any other frequency with the following procedure:

First calculate two constants, m and n:

$$m = \frac{\alpha_1 F_2 - \alpha_2 F_1}{F_2 \sqrt{F_1} - F_1 \sqrt{F_2}} \quad n = \frac{\alpha_2 \sqrt{F_1} - \alpha_1 \sqrt{F_2}}{F_2 \sqrt{F_1} - F_1 \sqrt{F_2}}$$

The cable loss α_3 at the frequency F3 is then:

$$\alpha_3 = m\sqrt{F_3} + nF_3 \quad (4)$$

Here's a practical example:

Coaxial cable RG-213, l = 9.58 meters. The full-wave frequency of the 9.58-meter cable is:

$$\frac{300}{9.58} \cdot 0.66 = 20.668 \text{ MHz}$$

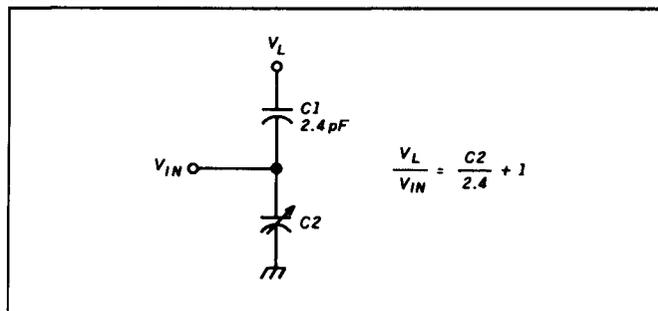
assuming a velocity constant of 0.66. This means that for a frequency of 5.167 MHz the cable is a quarter wavelength long. This brief calculation only helps you find the approximate value of the measurement frequency. Its exact value depends on the minimum value of V_{in} . The next minimum occurs at the three-quarter wavelength frequency of 15.417 MHz. In general, the first measurement frequency is the lowest frequency on the frequency dial where V_{in} has its first minimum.

Assuming that the minimum of V_{in} occurs at 5.116 MHz, the capacitor settings are: C2 = 153 pF and C1 = 2.4 pF.

$$\frac{V_L}{V_{in}} = \frac{153}{2.4} + 1 = 64.75 \quad \alpha l = \frac{1}{64.75} = 0.0154 \text{ N}$$

The measurement at frequency F2 = 15.348 MHz is: C2 = 79 pF and C1 = 2.4 pF.

FIGURE 4



Capacitive voltage divider circuit used in the test circuit at the output end of the coax under test.

$$\frac{V_L}{V_{in}} = 33.9l \quad \alpha_2 = 0.0295 \text{ N}$$

Calculating factors m and n from $\alpha_1 = 0.0145 \text{ N}$, F1 = 5.116 MHz, and $\alpha_2 = 0.0295 \text{ N}$, F2 = 15.348 MHz gives m = 0.00582 and n = 0.000436.

If you want the loss at frequency F3 = 28 MHz:

$$\alpha_3 = m\sqrt{F_3} + nF_3 = 0.043 \text{ N}$$

Measurement accuracy

Now you need to find out how large an error results if you assume the simple relationship $\frac{V_{in}}{V_L} = \alpha$ which the test set is delivering.

In reality, the source impedance R_s , even when purely resistive, affects the tuning frequency for minimum input voltage V_{in} . The load impedance Z_L is not infinitely large, and also affects the tuning. In our example, capacitor C1 = 2.4 pF has a capacitive reactance of $Z_L = -j2962 \text{ ohms}$ at 5116 MHz using Equation 1 you get:

$$\frac{V_{in}}{V_L} = \cosh(\alpha + j\beta) + j 0.003857 \sinh(\alpha + j\beta) \quad (5)$$

According to the test procedure, you must adjust the frequency for minimum input voltage. You can calculate V_{in} from Figure 1 to:

$$V_{in} = V_0 \frac{Z_{in}}{Z_{in} + R_s}$$

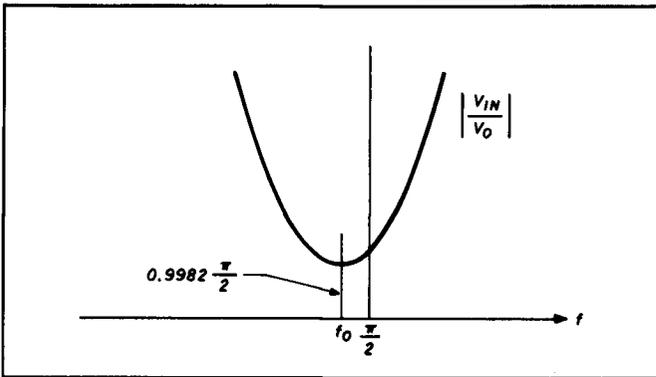
Since

$$Z_{in} = Z_0 \frac{Z_L + Z_0 \tanh(\alpha + j\beta)}{Z_0 + Z_L \tanh(\alpha + j\beta)}$$

we can calculate $\frac{V_{in}}{V_0}$ to:

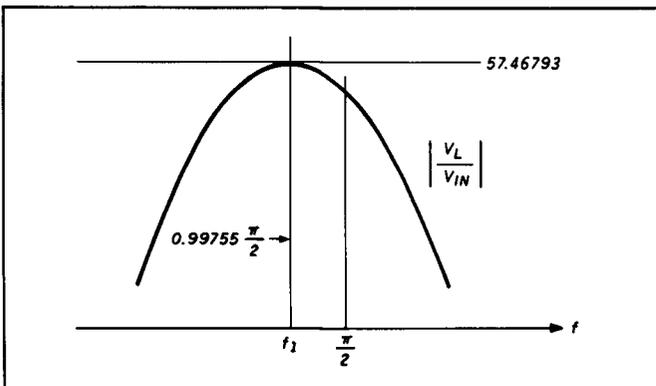
$$\frac{V_{in}}{V_0} = \frac{Z_{in}}{Z_{in} + R_s} = \frac{1 + \frac{Z_0}{Z_L} \tanh(\alpha + j\beta)}{Z_0 \left(1 + \frac{Z_0}{Z_L} \tanh(\alpha + j\beta) \right) + R_s \left(\frac{Z_0}{Z_L} + \tanh(\alpha + j\beta) \right)} \quad (6)$$

FIGURE 5



Typical plot for the results of Equation 6.

FIGURE 6



Plot of V_L/V_{IN} versus frequency for $Z_1 = -j12962$ ohms and $\alpha = 0.0174$ nepers.

A typical plot of Equation 6 is shown in Figure 5 which is drawn for $R_s = 1 + j0.2$ ohms, $Z_L = -j12962$ ohms, and $\alpha = 0.0174$ N. It can be shown that the input voltage minimum occurs to the left of the $\frac{\pi}{2}$ point (frequency F_0) if there's a capacitive load Z_L and resistive R_s . If you have a slightly inductive source impedance R_s and an infinitely high Z_L , the voltage minimum will occur to the right side of the $\frac{\pi}{2}$ point. Which effect dominates determines the location of the minimum. However, the deviations are very small. Plots 5 and 6 are frequency independent; the point π denotes the quarter-wavelength frequency. All voltage ratios are in absolute values.

Having found the frequency F_0 from Figure 5 by using Equation 6, you must insert its value into Equation 5 to find $\frac{V_L}{V_{IN}}$. Figure 6 is the plot of $\frac{V_L}{V_{IN}}$ versus frequency for $Z_L = -j12962$ ohms and $\alpha = 0.0174$ N (resistor R_s doesn't affect this ratio).

The maximum value of $\frac{V_L}{V_{IN}}$ occurs at the frequency F_1 of $0.99755 \frac{\pi}{2}$, giving a $\frac{V_L}{V_{IN}}$ value of 57.46793, whereas the true voltage ratio for the cable attenuation of $\alpha = 0.0174$ N is:

$$\frac{V_L}{V_{IN}} = \frac{1}{0.0174} = 57.47126$$

This represents an error of only 0.0058 percent for the ideal case when the minimum of V_{IN} occurs at the same frequency as the maximum of $\frac{V_L}{V_{IN}}$. This isn't always the case, but the deviation is small. You must measure V_L at the frequency where V_{IN} has its minimum, not at the frequency where V_L has its own maximum, because the maximum of V_L doesn't coincide with the maximum of $\frac{V_L}{V_{IN}}$.

In reality, the frequency F_0 may deviate slightly from this ideal F_1 value, depending on the canceling effect of R_s and Z_L . This error-reducing effect is a welcome and unexpected benefit, since R_s is always slightly inductive because of the difficulty in realizing a noninductive 1-ohm resistor.

In the example, the frequency F_0 for the lowest input voltage is $0.9982 \frac{\pi}{2}$. The corresponding $\frac{V_L}{V_{IN}}$ ratio, as calculated from Equation 5, is 57.3697. This amounts to an error of only 0.18 percent. The error would be 0.4 percent without the capacitive load. In other words, the "minimum V_{IN} " method is quite accurate, and its accuracy depends mainly on the capacitive voltage divider. It shows that the measurement error in the practical case of capacitive load and inductive source impedance is even smaller than that of the ideal case with infinitely high load impedance.

If you make a direct RF voltage measurement of V_{IN} and V_L without the test set, you must connect a capacitor whose value is equal to the meter capacitance at the cable output while you adjust the frequency for minimum input voltage. Remove this capacitor when you measure V_L with the meter, but don't alter the frequency setting. Even if you don't make this substitution, the loss measurement is still accurate enough for most applications.

Summary

Accurate and credible measurement of cable loss (calculated as the difference between input and output power) is difficult if the loss is small (for a short cable or at low frequencies).

The ratio of output to input voltage of a transmission line that is unterminated and driven from a low source impedance has a maximum value when measured at odd multiples of the quarter-wave frequency. The reciprocal of this ratio is very nearly equal to the cable loss in nepers within an error of less than 1 percent.

There are two ways of finding this maximum. The first involves the direct measurement of V_{IN} and V_L with a calibrated RF voltmeter. The second is based on the voltage comparison between the capacitively divided output voltage and the input signal; it doesn't require a calibrated RF voltmeter. The capacitor dial can be calibrated in picofarads or directly in nepers.

By repeating the measurement at the three-quarter wavelength frequency, you'll get two sets of attenuation and frequency values. You can find the cable loss at any frequency between the measuring points by using a simple interpolation method. However, actual tests show that the equations also render excellent results for frequencies much higher than the measuring frequency. The two error sources of this method, the inductive source impedance and the capacitive load, tend to cancel each other and the measurement accuracy is mainly dependent on the accuracy of the capacitive voltage divider.

One great advantage of this method is that you don't have to know the characteristic impedance of the cable or its velocity factor; therefore, no mismatching error exists. The method is especially useful when the cable loss is small, making conventional loss measurement unreliable. Even a piece of coax as short as 1 meter can be measured accurately.

Practical cable tests, together with the interpolation method, produced excellent agreement with published data sheets. 

REFERENCE

1. Forrest Gehrke, K2BT, "Real Coax: Impedance and Phase Relationships," *Ham Radio*, April 1987, pages 8-14.

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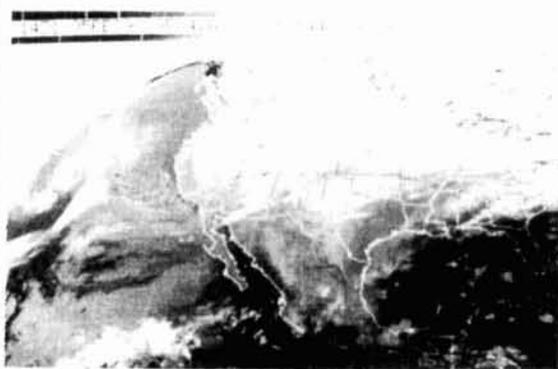
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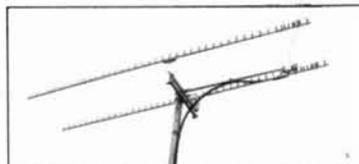
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ANTENNA TIME

Spring is just ahead and it's time to start thinking about worthy antenna projects. I've received a number of letters about unusual antennas that I think are interesting. Here are some of them for your consideration.

The 10-meter "Hentenna" loop

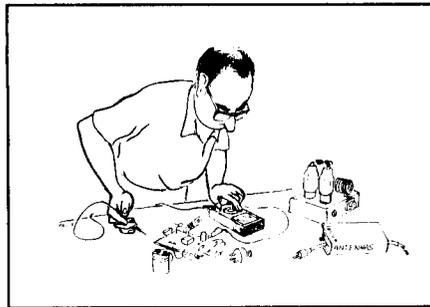
Harold Muensterman, N9DEO, sent me data on this clever little antenna developed by JE1DEU of Sagamihara City, Japan. Local hams were amused by the loop; hence the name — "hen" means curious in Japanese. "Hen-antenna" was quickly shortened to "Hentenna." It's shown in **Figure 1**. This antenna's virtue is that it has very little "wingspread" and is quite unobtrusive.

The array has two one-sixth wave radiators separated vertically by a half wavelength. To feed them, connect the tips and tap the vertical wires with a coax feedline. Polarization is horizontal.

Hentenna construction is simple. You use a single mast; try a TV-style push-up one. Make your horizontal sections out of 5/8-inch diameter aluminum tubing bolted to a mounting plate, and attach the plate to the mast with U-bolts. Use enamel-coated copper wire for the antenna's vertical sections.

Feed the Hentenna with a balun and coax line. Run your feed wires from the balun to the vertical wires. Adjust for lowest SWR by moving the feed wires up or down the vertical wires. Copper alligator clips are ideal for this; you can remove them and make joint solders when you find the correct points. The points should be about 36 inches above the bottom tube for 10 meters.

The Hentenna provides a figure-eight pattern at right angles to the



antenna plane. Gain is estimated at about 2.5 dB over a dipole. Bandwidth is very broad. By changing the length of the vertical wires, you can move the design frequency to any point in the 10-meter band.

A hanging unipole antenna for 160 meters

Phil Morgan, WD0P, uses a simple folded monopole for 160 operation (**Figure 2**). Phil says, "Being a cheap-skate, I put up this antenna made out of Radio Shack loudspeaker cable. I used a tree limb about 46 feet above the ground. I zipped 45 feet of this wire down the middle, soldered the leads together at one end and spaced it

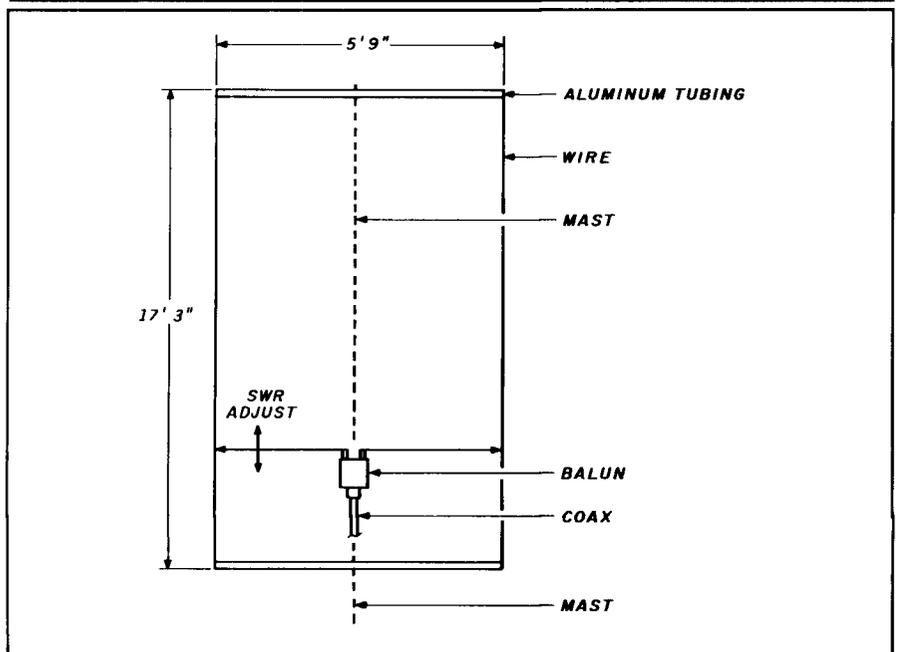
down its length with a half-dozen spreaders made of 6-inch pieces of 1/2-inch PVC tubing. I grounded one wire at the bottom and fed the other. Since the antenna was too short for 160 operation, I soldered a single wire to the top of the vertical portion and trimmed its length to resonate the whole works in the middle of the 160-meter band. I use my Heath SA-2060 Transmatch to give me access to the whole band.

"I have a pretty good ground system — an old, abandoned underground water tank, plus four 50-foot radials made of 2-foot wide chicken wire fencing laid on the ground. I'm not a big DXer, but have worked Hawaii and Cape Verde Islands on 160 using this lash-up. If you change the length of the single wire, the antenna will also work on 80 meters."

Any information on GFI?

In closing, Phil asks if anyone has

FIGURE 1



"Hentenna" for 10 meters. Adjust tap points for lowest SWR.

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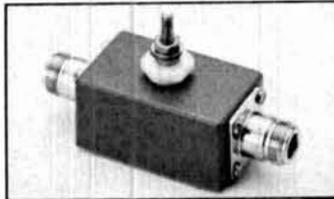
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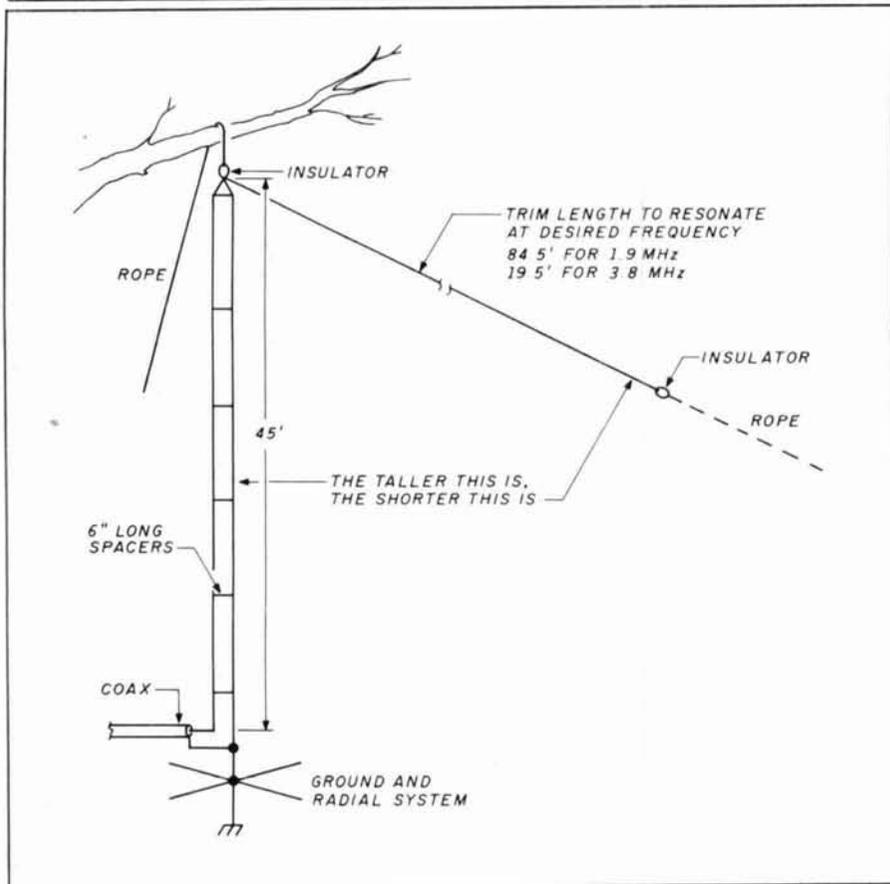


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FIGURE 2



The 160/80 meter antenna at WD0P.

information on Ground-Fault-Interrupter interference (GFI). He's run into these pesky devils in trailer parks and finds they're very sensitive to RF. He notes that the slightest tap of the key or word spoken into the mic will cause these GFIs to trip. Phil has also found GFIs on boat docks, so hams operating around a marina may have the same problem. Any answers to this? Write me if you have a solution.

A six-element quad on an 80-foot boom!

I received a note and photo from Victor Trachenco, UA6LA, Rostov-on-Don, USSR. He's the proud owner of a six-element, 20-meter quad (Photo A). This monster is built on a well-trussed 80-foot boom. So when you hear the blockbuster signal from UA6LA, you'll know it comes from this giant antenna!

The K6WZ tilt-over tower

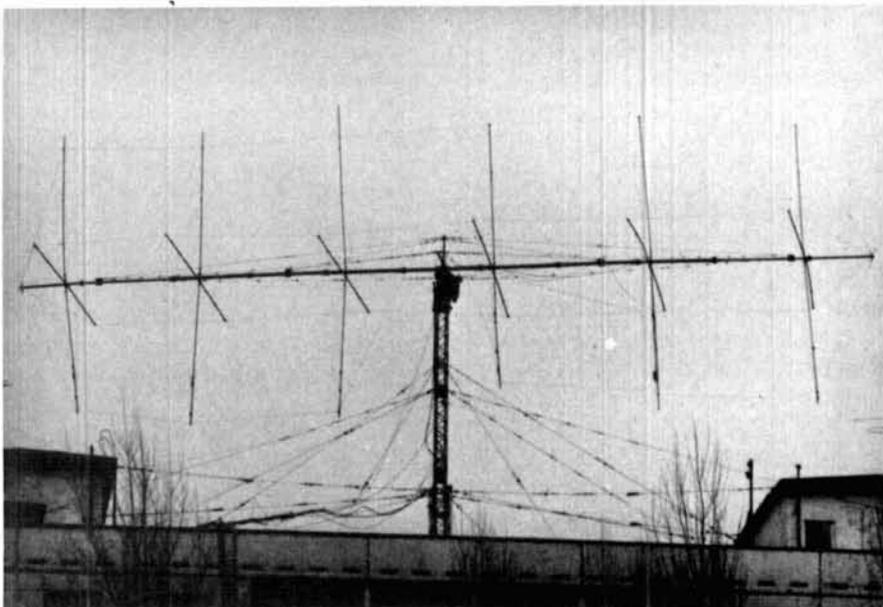
I mentioned in a previous column that I'd seen a lot of tilt-over towers in New Zealand, but not many in the United States. Some of you have been kind enough to send me information on homemade tilt-over towers. Here are two interesting designs.

Carl Stevenson, K6WZ, of Herington, Kansas, built the wood tower shown in Photo B and Figure 3. It's about 30 feet high. The fixed, bottom portion is made of two 2 x 6 pieces of pressure-treated lumber 16 feet long. They are treated, before assembly, with Thomson's Waterseal. The 2 x 6's are sunk about 4 feet in the ground.

The top tilt-over portion of the tower is made of a single section of 2 x 6 material 20 feet long. It's pivoted near the top of the two lower sections with a heavy bolt. The pivot point on the top section is placed so that about 17 feet extend above the lower supports, and there's a 3-foot lever arm below the pivot point. The pivot is made of a short piece of half-inch pipe with a heavy bolt running through it.

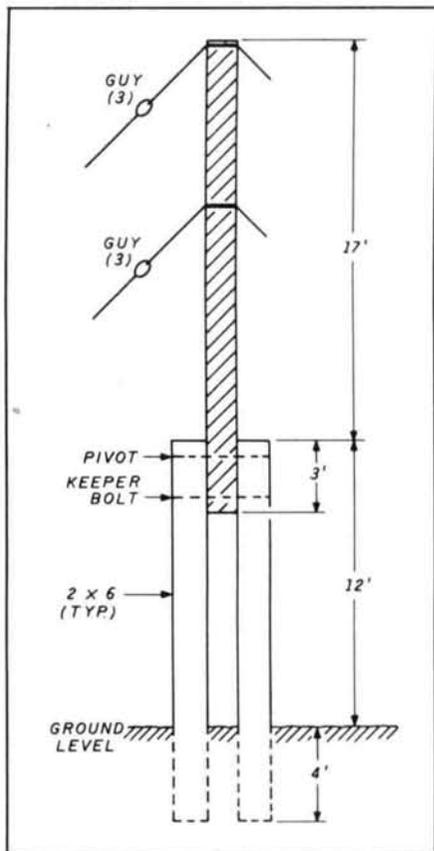
A rotating mast of 10-foot TV mast sections runs up the wood mast. The rotor is about 8 feet above ground and turns the mast sections which support the antenna. The mast sections are attached to the wood tower by large eye bolts whose "eyes" are opened just enough to pass the rotating metal mast. The rotor sits on a small metal shelf bolted to the tilt-over section. To counterweight the tower, a few pieces

PHOTO A



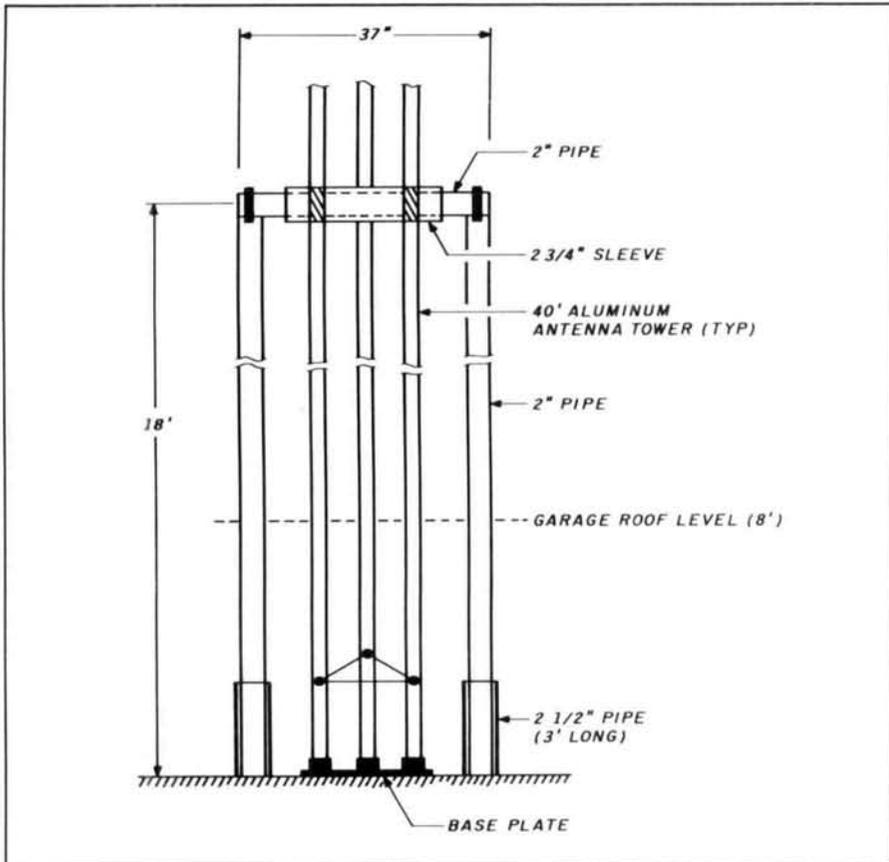
The 6-element, 14-MHz Quad at UA6LA is built on an 80-foot, trussed boom.

FIGURE 3



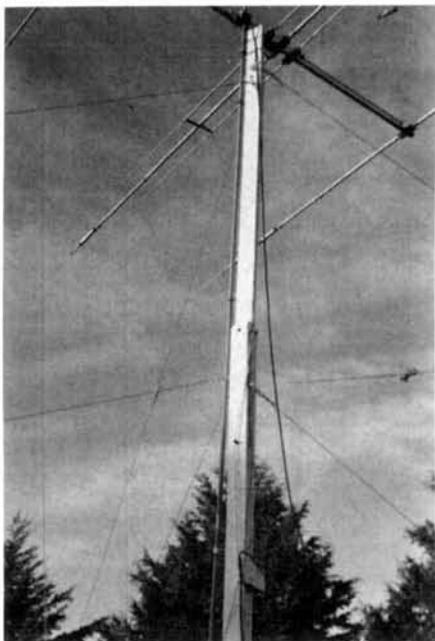
Layout of K6WZ tilt-over tower.

FIGURE 4



Tilt-over tower support at K9BX.

PHOTO B



The tilt-over tower at K6WZ. Made of 2 x 6 lumber. The tower is about 30 feet high. Rotor is placed near ground level and antenna is turned by supporting pipe.

of scrap metal are bolted to the shelf below the rotor.

To tilt the tower, Carl loosens two top back guys and removes the keeper bolt. He uses a little care and safety ropes, and over it comes!

Carl built the original tower in California 15 years ago. It was so successful that he built a duplicate when he moved to Kansas. He's worked 214 countries on RTTY with this low-cost, simple tilt-over antenna system, using a small triband beam!

The K9BX tilt-over tower

"Doc" Roberts, K9BX, of Rothschild, Wisconsin, has adapted a commercial tower to a tilt-over design (Figure 4). He's had this arrangement for 20 years and is very pleased with it.

The tilt-tower support frame is 18 feet tall and made of salvaged pipe. Each arm is built from 2-1/2 inch schedule 40 pipe on the bottom section, and 2-inch pipe on the top section. Running across the top of the tower is a 37-inch crossarm of 2-inch pipe welded into ears at the tops of the vertical legs. A 2-3/4 inch pipe, about 24 inches long,

fits over the crossarm of the tower and functions as a pivot point. It is fastened to the tower.

The base of the tower (a 40-foot aluminum model) is bolted into a base mount anchored in concrete. The tower sits alongside a flat-roofed garage; two of the vertical legs are anchored to the garage about 8 feet above ground.

To lower the tower, a pulley and rope system is attached to the bottom, the bolts anchoring the tower to the base plate are removed, and the tower is lowered manually. When the tower reaches a horizontal position, the rope is secured. The tower is then about 7 feet above the roof level, and the antenna is in an ideal position to work on. (Doc has a two-element quad on the tower.)

Doc says that after 20 years of wind action on the tower, the bolt holes near the tilt-over joint have become slightly elongated. He's reinforced these points and has added a set of top guy wires. Up to this time, it had been a freestanding tower, anchored only at the top of the garage.

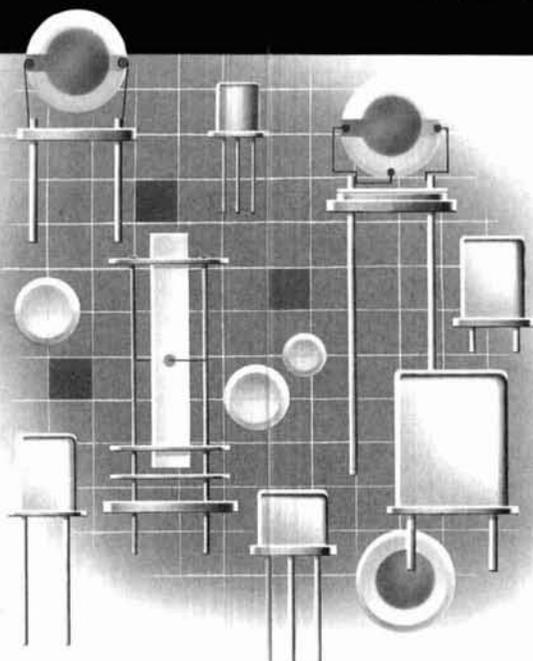
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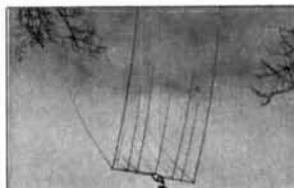
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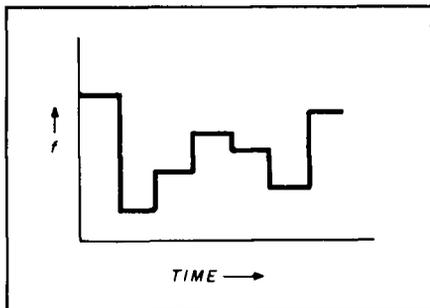


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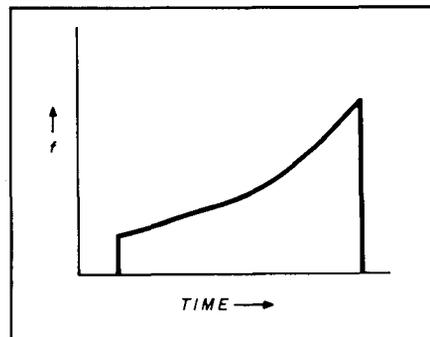
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FIGURE 5



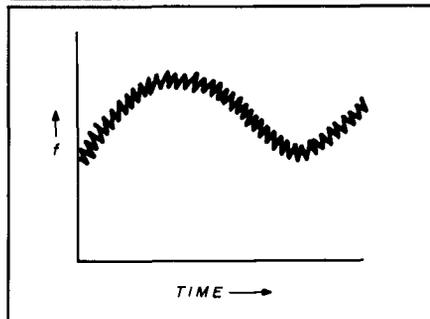
Frequency-agile ("hopping") signal.

FIGURE 6



Radar "chirp" signal.

FIGURE 7



"Jitter" signal.

Things that go "bump" in the night!

Back in the "good old days," emission was either phone or CW. As time went on, things became more complicated. A0 Emission was unmodulated carrier, A1 was CW, A2 was tone-modulated CW, A3 was phone, A4 was facsimile, and A5 was television. There were also "F-type" emissions covering FSK, FM, and so on. Things were getting confusing!

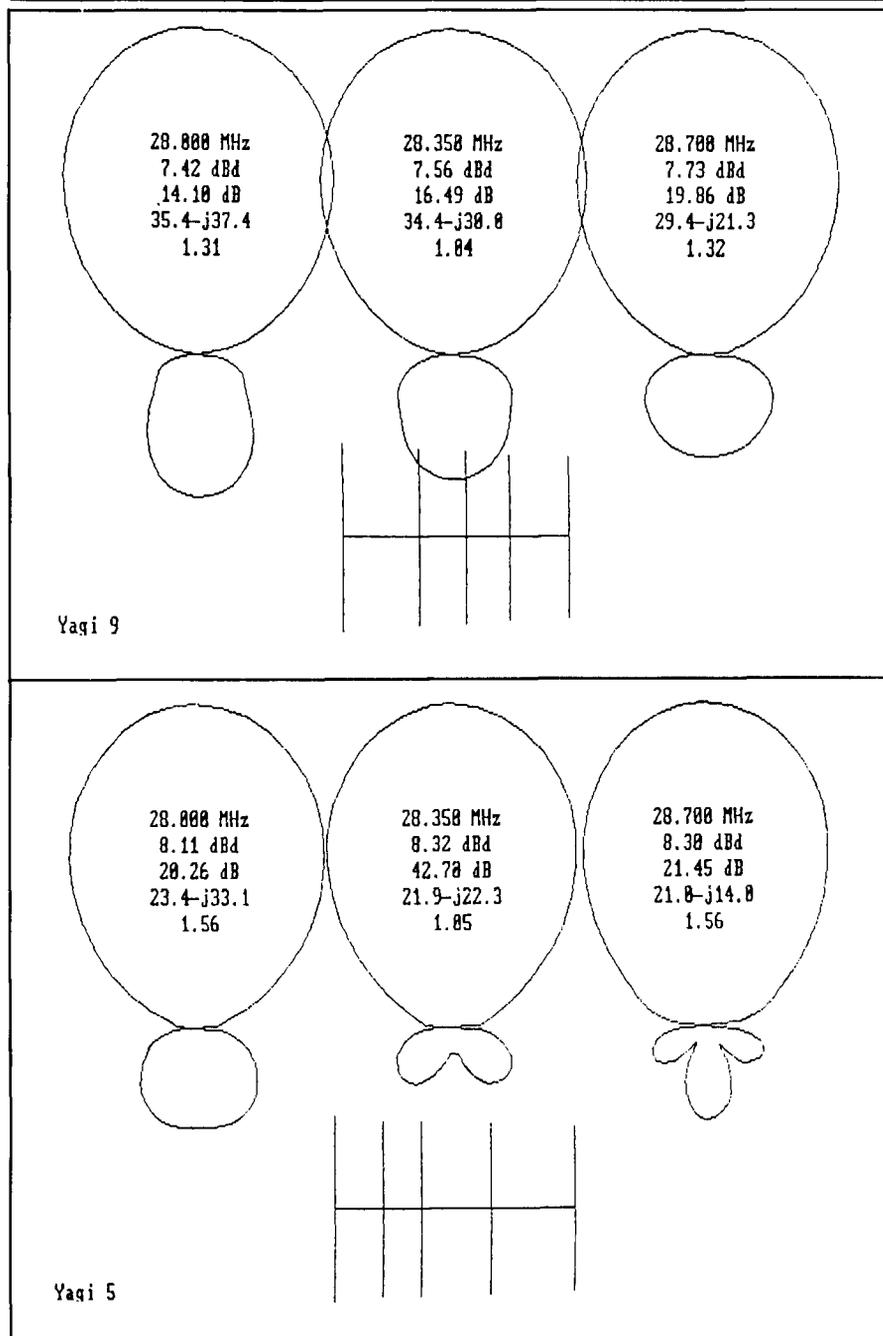
Today there's a half-page of "emission classifications" in the *ARRL Handbook*! Single sideband, suppressed

carrier (SSB) is listed as J3E (formerly A3J). That's simple enough. But if you're running "angle modulation" (FM) with 5-kHz deviation and a maximum modulation frequency of 3 kHz, your emission classification is 16K0F3EJN. If you use phase modulation, your classification is 16K0G3EJN. And poor old Morse code is now 150HA1AAN (if you're sending 60 wpm).

What this means is that hams now hear a lot of funny signals in and out of the ham bands. They may or may not have emission classifications but, for the most part, they certainly aren't ham signals. Here are some of the more interesting ones.

Figure 5 is a sketch of a frequency-agile ("hopping") signal which varies frequency in a predetermined manner, with the receiver locked to the rapidly

FIGURE 8



Top: Commercial 10-meter beam. Bottom: Commercial beam optimized. Constants listed are frequency, gain, F/B ration, input impedance, and SWR.

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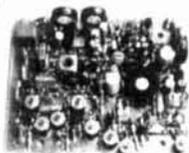
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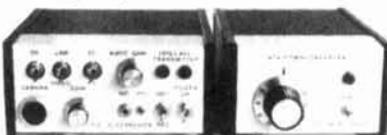
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shifting signal. The hopping sequence often lands the signal in a ham band; you hear a quick burst of voice or other modulation, and the signal is gone. Does anyone know who the "hopper" is? I don't.

Figure 6 shows a representation of a "chirp" emission. This is often used with a radar, where the frequency varies with time. Figure 7 illustrates a "jitter" signal. The jitter varies with frequency, within a predetermined time frame.

There are other weird signals including: the well-known Soviet "Woodpecker", OTHR (Over-the-Horizon Radar), and Soviet missile tracking radars that occasionally pop up in the 10 and 12-meter bands. American OTHR is also on the air, but hops about randomly, and so far has caused no lasting interference in the hams bands. Other OTHRs on the air in various countries sometimes appear in an Amateur band.

Countless other curious things (like single-letter beacons) abound, and they often cause QRM in and out of the ham bands. There are also mysterious "numbers" stations, which repeat coded number groups throughout the HF range.

All of these signals cause QRM in an already jammed radio spectrum. Unfortunately, it's often easy to turn them on but not so easy to turn them off! Happy listening!

The Yagi optimizer disk

Brian Beezley, K6STI, has come up with another interesting disk for IBM-PC and compatible users. It's called the "Yagi Optimizer, YO version 1.00." This program automatically optimizes

a Yagi antenna for maximum forward gain, best pattern, and minimum SWR. The package includes models for matching networks, element tapering, element-to-boom mounting plates, frequency scaling, and element taper scaling.

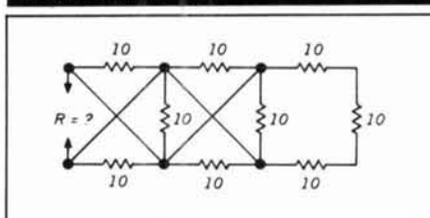
The YO program also plots antenna radiation patterns at the central design frequency and at the band edges. Best of all, YO has been designed to work alone or in conjunction with MININEC3, the high-accuracy general purpose antenna analysis program. This makes it convenient to analyze and optimize a Yagi previously analyzed with MININEC3.

Figure 8 shows an example of "before" and "after" optimization of a commercial 5-element 10-meter Yagi. Interested? Write to Brian (507-1/2 Taylor Street, Vista, California 92084) for full details.

The Dead Band Contest

I appreciate all the letters and cards I've received in response to this little

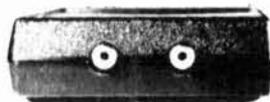
FIGURE 9



What is input resistance?

exercise. Thank you all very much! This month's quiz is an easy one. I'll give the answer next month. Find the input resistance to the network in Figure 9. Each resistor is 10 ohms. 

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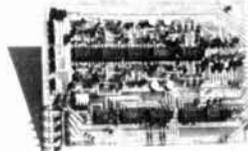
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Tell 'em you saw it in HAM RADIO!

THE MICROFARAD COUNTER

By Hans Evers, PA0CX, Wintererstrasse 3, D-7800 Freiburg, W. Germany

You can often find second-hand electrolytic capacitors at amazingly low prices at flea markets and surplus stores. The reason? Nobody trusts them (the capacitors, that is).

It isn't always easy to check the quality of a polarized capacitor. The number of microfarads may exceed the range of your measuring bridge, and it's rather unusual to find provisions for applying the necessary DC polarization during the measurement. Also, a quick assessment of possible leakage in a large capacitor could lead to problems.

With these considerations in mind, I decided to build a basic test box. My efforts resulted in an almost suspiciously simple schematic diagram. Yet the test box (though small enough to fit in your pocket) measures not only how many microfarads there are with reasonable accuracy, but whether the capacitor under test leaks. Although the circuit consists of only a few discrete components, it boasts an elementary "digital display."

The amount I spent for materials (practically all were supplied by my modest junkbox) was rapidly paid off when I used my new gadget to sort a shoe box full of old, partly used electrolytic capacitors. The contents of the box had become the subject of more and more distrust over the years, and I was actually at the point of throwing the whole lot overboard.

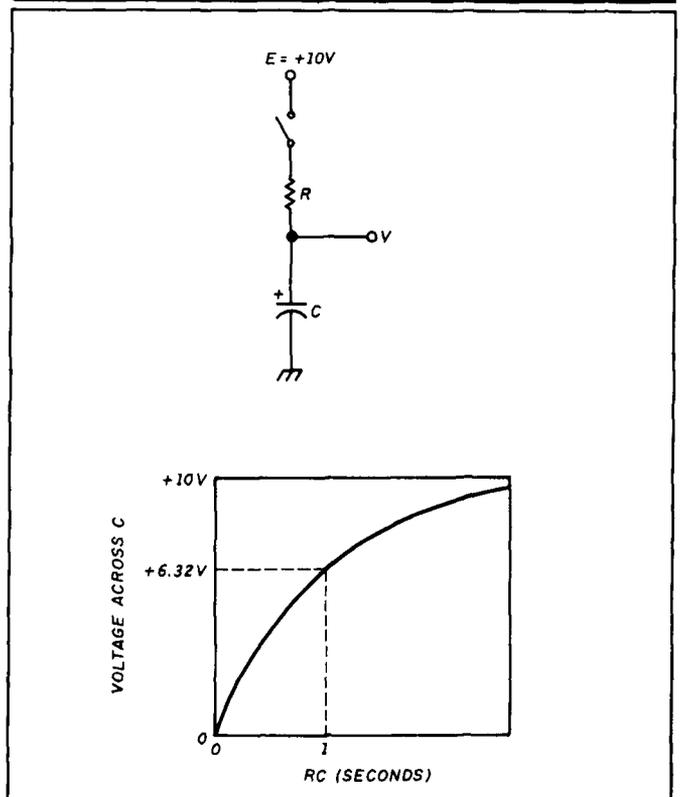
The principle

The principle of the microfarad counter is that of a capacitor charged through a resistor. When voltage is applied, the capacitor voltage builds up exponentially starting from zero, just as the textbooks specify. Something interesting occurs when the elapsed time (in seconds) becomes equal to R (in ohms) times C (in farads). You'll find the voltage across the capacitor has grown from zero to 63.2 percent of the supply voltage*.

If, for example, the supply voltage is 10 volts and $R=1$ meg (as shown in Figure 1), you reach a voltage of 6.32 across a $1\text{-}\mu\text{F}$ capacitor after 1 second. For an $n\text{-}\mu\text{F}$ capacitor this would take n seconds. Thus, microfarads can be measured by counting seconds. It isn't too difficult to give the circuit a differ-

* $1 - \frac{1}{e}$ times 100 percent, to be exact

FIGURE 1

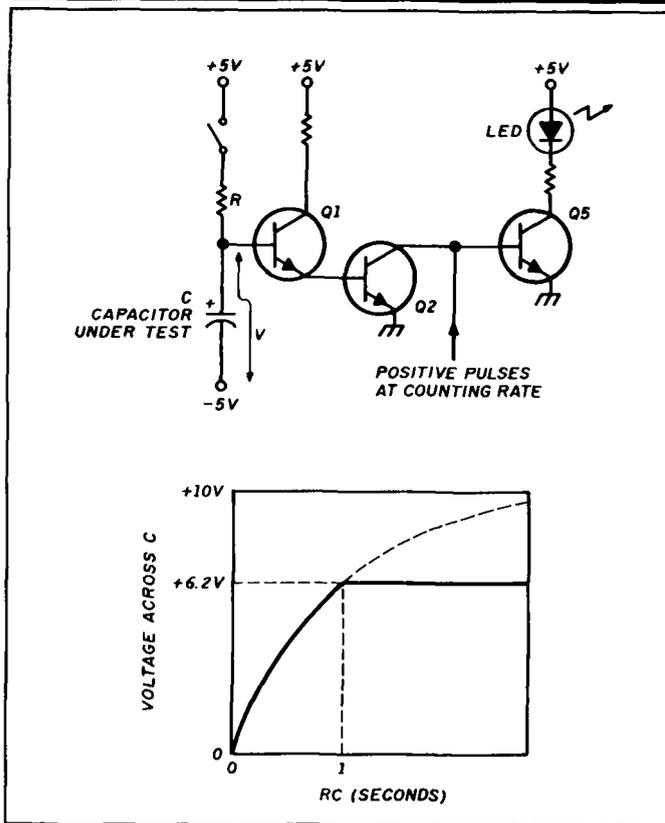


Charging a capacitor through a resistor.

ent "measuring range." If you make $R = 100$ k (ten times smaller), each second will represent 10 microfarads.

Putting this simple principle into practice is altogether something else. The voltmeter necessary to determine the 6.32 volts would unavoidably establish a bypass around the capacitor. This, in turn, would behave like a "leaky" capacitor. Because of the leakage, it would take more time to reach the threshold

FIGURE 2



Principle of the Microfarad Counter.

voltage, making the capacitor look larger than it actually is.

You're left with two choices. You can measure the voltage over R and then subtract it from the supply voltage. Or use a voltage indicator that, at least until the threshold is reached, looks like an electrical insulator. I found that the second option was the simplest. A wristwatch (or even a stopwatch) isn't the most practical time indicator in this case.

An improved counting system

There's a better system for counting the seconds. Figure 2 shows how it's done. The stopwatch has been replaced by a light-emitting diode (LED) that produces light flashes for counting the charging time — or rather the microfarads. This counting mechanism is switched on as soon as the capacitor starts charging, and is switched off when the capacitor voltage reaches 6.2 volts (near enough to the ideal 6.32 volts, for the moment). Q2, the actual switching element, acts as a temporary short across the base of LED driver Q5.

The initial capacitor voltage is zero with the power supply connected. The base of Q1 begins by looking at -5 volts; Q1 blocks the base current of Q2 which, therefore, can't conduct. So while the capacitor starts charging, nothing prevents the LED from blinking.

The capacitor voltage grows, and when the Q1 base voltage arrives at +1.2 with respect to ground, both Q1 and Q2 start conducting because their base-emitter junctions now have the required 0.6 volt across them. The emitter-collector path of Q2 forms a short and the LED stops blinking. When this happens, the capacitor (situated between -5 volts at the bottom and +1.2 volts at the top) has been charged up to 6.2 volts, indicating that $t=RC$ or: $C = \frac{t}{R}$.

Because the base voltage of Q1 remains virtually constant, the voltage across the capacitor doesn't rise, and remains stabilized at 6.2 volts. With the measurement completed, the current through R no longer flows into the capacitor; it finds its way through the transistors to ground instead. The microfarad meter isn't complete yet. As you saw before, the capacitor would cause a false reading if there's leakage. So for the measurement to make sense, it's essential to know whether the capacitor under test is leakproof.

Leakage detection

A conventional leakage-current meter would be quite something to design. Even leakage currents as low as several microamperes can be significant (remember they are DC), especially in smaller sized capacitors. The measured current should be totally independent of charge and discharge currents; this requires that a capacitor voltage remain untouched when the meter circuit is introduced. Fortunately, the concern when using the test box is whether the leakage is serious enough to spoil your measurements. You need not worry about the actual amount of current leaked.

I based my method on the following statement: "Only if you extract the same amount of electricity from the capacitor as you've put in can you be sure that the capacitor isn't leaking." In other words, if the capacitor after testing is discharged under the same conditions as it was charged, the charging and discharging processes should take the same amount of time.

You have to use an unusual technique to discharge the capacitor under the same conditions. Instead of being discharged passively across a parallel resistance (a method that would be unsuitable here as the process would follow a different portion of the characteristic), the capacitor is discharged actively when it's supplied with the same current in the opposite polarity.

The capacitor connections (charged to 6.2 volts as the result of the previous microfarad measurement) are reversed with a toggle switch. The positive side of the capacitor is now connected to a point that carries +1.2 volts and the negative side is connected to R (see Figure 3).

Initially, the base of Q1 sees a voltage of +1.2 volts in series with -6.2 volts, equaling -5 volts. The current through R discharges the capacitor. When the capacitor is fully discharged, the voltage across it is zero, and the base voltage of Q1 has arrived at +1.2 volts. This stops the process.

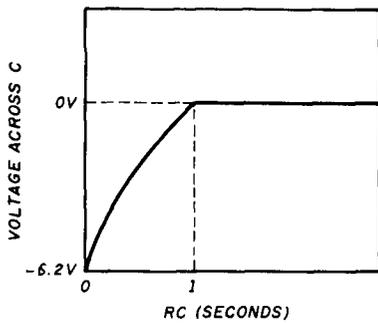
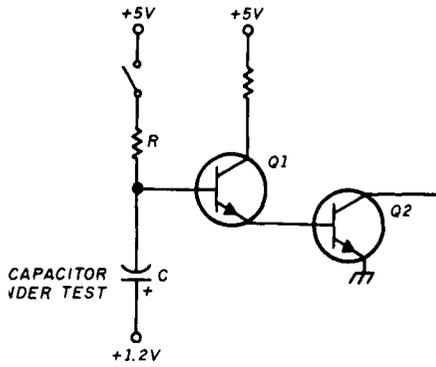
Q1 doesn't differentiate between charging or discharging, and the LED blinks in both cases. When measuring a healthy leak-proof capacitor, the LED blinks the same number of times for discharge as it does for charge. If there are fewer flashes, some of the electric charge was lost and you can conclude that the capacitor leaks.

During the entire charge-discharge cycle, the capacitor is never subjected to any voltage higher than 6.2 volts. The capacitor is left completely discharged after the capacity and leakage check.

The counter

It seemed appropriate to do some basic research before designing the blinking LED mechanism. I found that the once-per-second flashing rhythm would be exasperatingly slow. After doing some experimenting, I decided that a more suitable counting rate would be two flashes per second. The blinking is relatively fast, but not so fast that you risk losing count.

FIGURE 3



How the capacitor under test is discharged.

PARTS LIST

CAPACITORS

- 1 0.68 μ F 25 volt tantalum
- 1 4.7 μ F 25 volt electrolytic or tantalum

POTENTIOMETER

- 100 k Radio Shack RS271-220

RESISTORS (all resistors 1/4 watt)

- R1—R5 see text, 5 percent
- 1 22 ohm
- 1 560 ohm
- 1 4.7 k
- 1 100 k
- 1 560 k
- 1 5.1 k
- 1 56 k

SEMICONDUCTORS

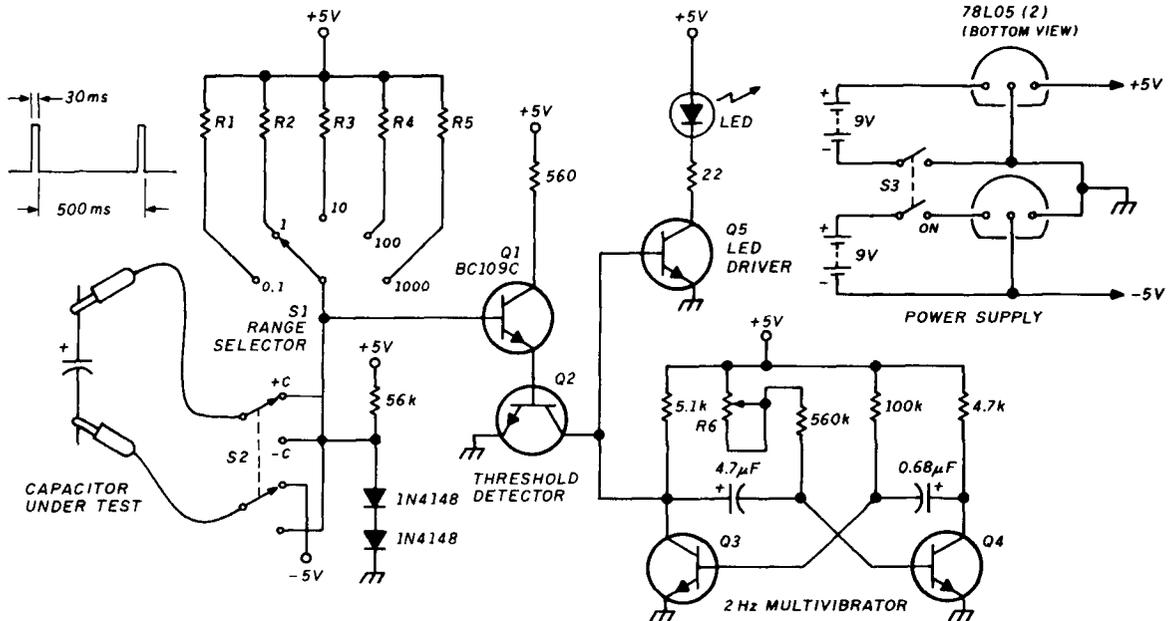
- Q1—Q5 ECG 123AP or equivalent
- 2 78L05
- 2 1N4148 or equivalent
- 1 LED

MISCELLANEOUS

- 1 PCB FAR Circuits*
- 1 DPST miniature toggle switch
- 1 DPDT miniature toggle switch
- 1 SP5P rotary switch
- 2 9-volt battery connectors

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FIGURE 4



Schematic diagram.

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It's obvious that, with the LED flashing twice as fast, you have to correct the RC time to maintain the principle of one count per microfarad. I compensated for this by changing R2 in **Figure 4** from 1 meg to 510 k. Although this is slightly more than the ideal of 0.5 meg, the difference takes care of the approximately 1.5-percent error in the 6.2-volt threshold (which ideally should have been 6.32 volts). I also had to consider the actual light/flash duration. I soon found out that a 50/50, on/off LED ratio isn't convenient for visual counting and that shrinking the length of the light pulse made counting easier. There are limits, of course — you can't make the length so small that visibility begins to suffer. Thirty ms seems to be a good compromise.

The drastic reduction in light/pulse length has more advantages. You can subject the LED to considerably more current than the 20 mA usually recommended. The 68 mA that flows through the LED (mainly determined by 22 ohms in the Q5 collector lead, and to a lesser extent by the collector resistor of Q3) visibly increases the brightness. Nevertheless, the average LED current still isn't more than a very reasonable 4 mA.

Calibration and accuracy

Calibration doesn't require a capacitor of any standard value. The accuracy of calibration depends entirely on the voltages (the output voltages of the 78L05s, as well as the constant voltage drops over the diodes and transistor junctions), the resistors (R1 and R5 may be as accurate as desired), and the counting time. You can adjust the factor time by shorting the measuring leads (to make the LED flash) and by tuning R6 for exactly two counts per second.

The microfarad counter's total accuracy depends largely on the time you spend on measurement. If you measure a certain capacitor in 5 seconds, the unavoidable uncertainty of the last digit may cause an error of 10 percent. But if you're willing to spend almost a minute on the same measurement by switching to a lower range, and if the capacitor under test is leakproof, the digital error will be limited to 1 percent. In this respect, the elementary counting system of the microfarad counter is in good company with other forms of digital displays, which also leave an uncertainty of at least plus or minus one digit.

Other design considerations

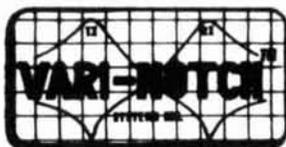
I designed the circuit with low power supply requirements in mind. See **Figures 5** and **6** for the pc board layout and component placement guide. That's why I chose the LED as an indicator, rather than something like an acoustic bleeper. To ensure long-term battery consumption, it seemed necessary to use two batteries with a double-throw "power" switch. The 78L05 regulators appeared to provide the cheapest solution for maintaining reliable voltages.

The battery for positive supply must deliver an average current of 11.5 mA (it can be as high as 20 mA in the $\times 1000$ position). The battery for the negative supply has to produce 3.5 mA; this means it should last longer.

Use any common, low-power, silicone, NPN-type transistors. I used BC 237As from my junkbox. You must make an exception for Q1. A BC 109C works well because of its high beta at very low currents (in the "times 0.1" position the base current is only 0.8 μ A!). If you can live without the times 0.1 range, you can use a more conventional transistor for Q1.

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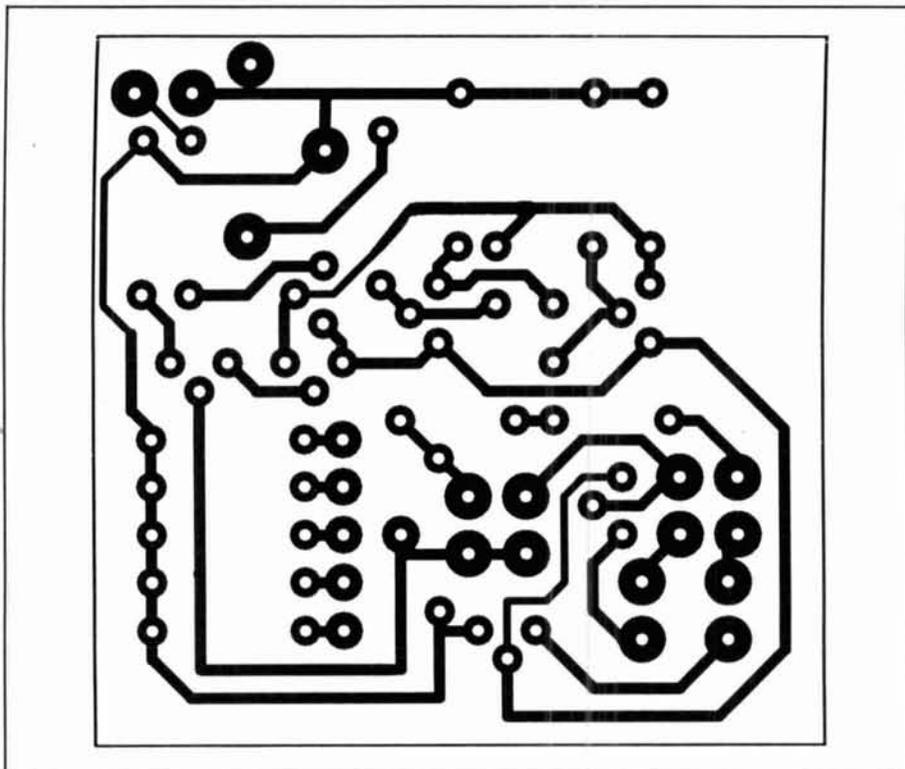
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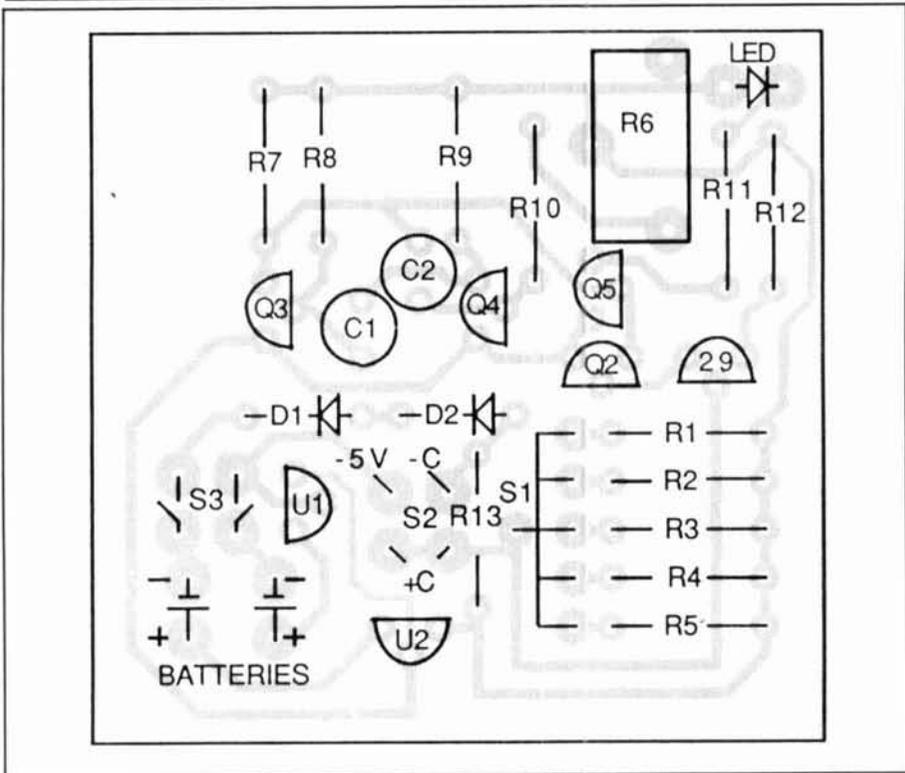
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FIGURE 5



Foil side of pc board.

FIGURE 6



Component placement guide (component side).

Precautions

Some electrolytic capacitors have a maximum voltage rating lower than 6.5 volts. Don't measure these with the microfarad counter; they may be damaged. Capacitors with maximum voltage ratings at the other end of the scale also require a word of warning. When dealing with high-voltage capacitors, whether they are electrolytic, tantalum, or just ordinary paper-insulated type, you must always be aware of residual electric charges. They can develop *even if the capacitor has been entirely discharged by a complete, long-lasting, full short*. It's a good habit to short any capacitor (even if it's been lying around for some time) before doing something with it.

In any case, it seems advisable to put Q1 and Q2 where they can be easily replaced if disaster strikes.

Operating instructions

WARNING: Residual electric charge on high-voltage capacitors may damage the test instrument. Discharge the capacitor before connecting.

- Connect the capacitor, with the switch in position "–C." Observe correct polarity.
- Set the switch in position "+C." The LED will start flashing. Count the flashes until they stop. Each flash represents 1 μF (or 10 μF , 100 μF , and so on, depending on the position of the range switch).
- Switch to position "–C." The LED will start flashing again. Count the number of flashes until they stop. If the number is less than before, the capacitor is leaking current.

Observations

The use of electrolytic capacitors is usually regarded as an unavoidable evil. Some common problems you may experience with these caps are: limited life expectancy, exaggerated tolerance in value, and excessive leak current. After building my microfarad counter and using it to test a large variety of capacitors, I've found it necessary to reevaluate my opinion. Not only did there appear to be more leakproof electrolytic capacitors in my collection than I dared hope for, but their accuracy was generally much better than I expected. Most stayed within a tolerance of about ± 15 percent. 



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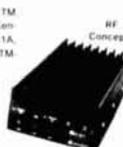


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Practically Speaking

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PART 1 HIGH-FREQUENCY DIPOLE ANTENNAS

An unfortunate myth arose in Amateur Radio circles some time ago. People came to believe that large antenna arrays were absolutely necessary for effective communications — especially for DX work. They tend to overlook basic, but effective, antennas that anyone can erect and make work. The simple *dipole* or *doublet* antenna is a case in point. This antenna is sometimes called the *Hertz* or *Hertzian* antenna, because radio pioneer Heinrich Hertz is said to have used it in his experiments.

The dipole is a balanced antenna with two quarter-wavelength radiators (Figure 1), making a total of a half wavelength. The antenna is usually installed horizontally, producing a corresponding horizontally polarized signal.

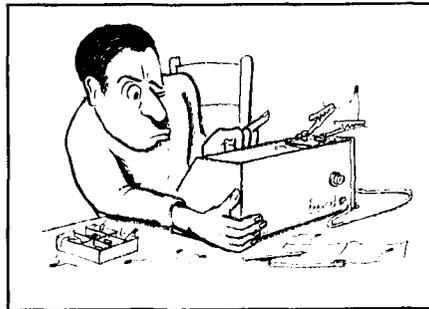
In its most common configuration (Figure 1), the dipole is supported at each end by rope and insulators. The rope supports are tied to trees, buildings, masts, or some combination of structures.

As I said before, the antenna length is a half wavelength. Remember that the physical length of the antenna and the theoretical electrical length often differ by about 5 percent. In free space, a half wavelength is found from:

$$L = \frac{492}{F_{\text{MHz}}} \text{ feet} \quad (1)$$

Equation 1 gives you the physical length of a perfect, self-supporting antenna that's many wavelengths away from any object. But for real antennas, the length calculated using this equation is too long. The physical length is about 5 percent shorter because of the capacitive effects of the end insulators. A more nearly correct *approximation* (remember that word; it's important) of a half-wavelength antenna is:

$$L = \frac{468}{F_{\text{MHz}}} \text{ feet} \quad (2)$$



Where:

L is the length of a half-wavelength radiator, in feet.

F_{MHz} is the operating frequency in megahertz.

Example 1

Calculate the approximate physical length for a half-wavelength dipole operating on a frequency of 7.25 MHz. Solution:

$$L = \frac{468}{F_{\text{MHz}}} \text{ feet}$$

$$L = \frac{468}{7.25} \text{ feet} = 64.55 \text{ feet}$$

or, restated another way:

$$L = 64 \text{ feet } 6.6 \text{ inches}$$

Unfortunately, a lot of people accept Equation 2 as a universal truth — perhaps because of books and articles on antennas that fail to tell it all. For example, you must consider *resonance*. An antenna acts like a complex RLC network. At some frequencies it will appear as an inductive reactance (X_L

= $+jX$), and at others as a capacitive reactance ($X_C = -jX$). At a specific frequency the reactances are equal in magnitude but opposite in sign, so they cancel each other out: $X_L - X_C = 0$. At this frequency the impedance is purely resistive, and the antenna is resonant.

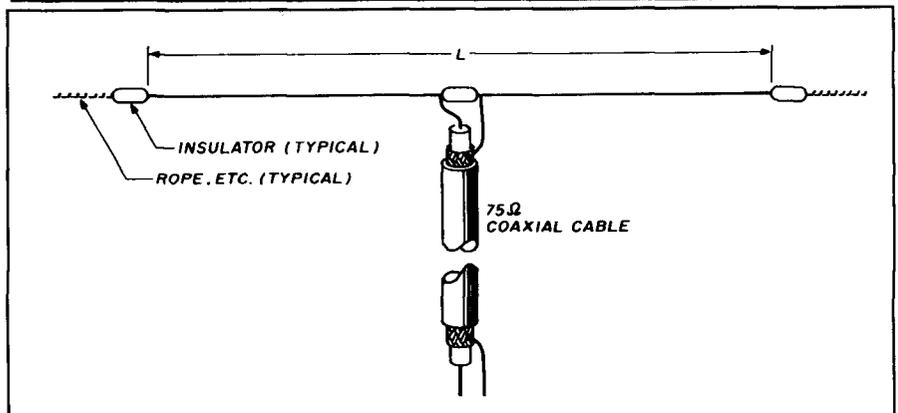
The goal in erecting a dipole is to make the antenna resonant at a frequency that's inside the band of interest — preferably the portion of the band most often used by your station. I'll discuss some of the implications of this later, but for now assume that you have to custom tailor the antenna length. Depending on several local factors (among them nearby objects, the antenna conductor's shape, and the conductor's length/diameter ratio), you may have to add or trim the length a bit to reach resonance.

The dipole feedpoint

The dipole is a half-wavelength, center-fed antenna. Figure 2 shows the voltage (V), current (I), and impedance (Z) distributions along the length of the half-wavelength radiator element. The feedpoint voltage is at a minimum and the current is at a maximum, so you can assume that the feedpoint is a current "loop" or "antinode."

The impedance of the feedpoint at resonance is $R_0 = V/I$. R_0 is made up of two resistances. First there are ohmic losses that generate nothing but heat when the transmitter is turned on. These losses result because conduc-

FIGURE 1



Standard coaxial cable fed half-wavelength dipole antenna.

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tors have electrical resistance, and because electrical connections aren't perfect (even when properly soldered). Fortunately, in a well-made dipole these losses are almost negligible. The second contributor is the antenna's radiation resistance (R_r). This resistance is a hypothetical concept that accounts for the fact that the antenna radiates RF power. The radiation resistance is the fictional resistance that would dissipate the amount of power radiated away from the antenna.

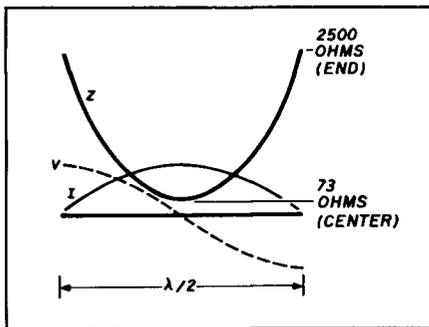
For example, suppose you're using a large diameter conductor as an antenna, and it has negligible ohmic losses. If you apply 1,000 watts of RF power to the feedpoint, and measure a current of 3.7 A, what is the radiation resistance?

$$R_r = P/I^2$$

$$R_r = (1,000 \text{ watts})/(3.7)^2$$

$$R_r = 73 \text{ ohms}$$

FIGURE 2



Plot of current, voltage, and impedance distribution along half-wavelength dipole.

It's important to match the feedpoint impedance of an antenna to the transmission line impedance. Maximum power transfer always occurs when the source and load impedances (in any system) are matched. If some applied power isn't absorbed by the antenna (as happens in a mismatched system), then the unabsorbed portion is reflected back down the transmission line towards the transmitter. This results in standing waves, and the so-called standing wave ratio (SWR or VSWR).

Matching antenna feedpoint impedance may seem easy because the free space feedpoint impedance of a simple dipole is about 72 ohms. You'd think this would be a good match to 75-ohm coaxial cable. Unfortunately, the 72-ohm feedpoint impedance is almost a myth in practical situations. **Figure 3** shows a plot of approximate radiation resistance (R_r) versus height

above ground (as measured in wavelengths). As before, you must deal in the approximations found in **Figure 3**; here the ambiguity is introduced by ground losses.

Despite the fact that **Figure 3** is based on approximations, you can see that radiation resistance varies from less than 10 to almost 100 ohms as a function of height. At heights of many wavelengths, this oscillation of the curve settles down to the free space impedance (72 ohms). At the higher frequencies it may be possible to install a dipole many wavelengths high. On the 2-meter band (144 to 148 MHz) one wavelength is around 6.5 feet (2 meters \times 3.28 feet/meter), so it's relatively easy to achieve "many" wavelengths at reasonably attainable heights. In the 80-meter band (3.5 to 4.0 MHz), however, one wavelength is about 262 feet, so many wavelengths is a practical impossibility.

There are three tactics you can follow. The first is to ignore the problem altogether. In many installations, the height above ground will be such that the radiation resistance is close enough to present only a slight impedance mismatch to a standard coaxial

cable. You'd calculate the VSWR as the ratio (among other ways):

$$Z_o > R_r: \quad VSWR = Z_o/R_r \quad (3)$$

$$Z_o < R_r: \quad VSWR = R_r/Z_o \quad (4)$$

Where:

Z_o is the coaxial cable characteristic impedance.

R_r is the radiation resistance of the antenna.

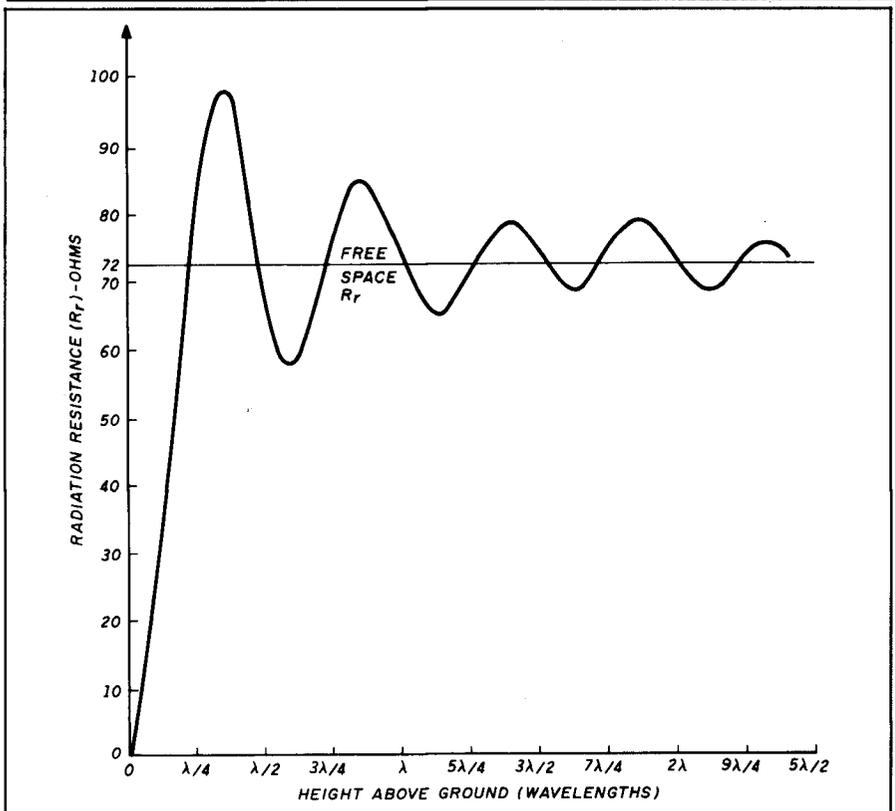
Consider an antenna mounted at a height somewhat less than a quarter wavelength, with a radiation resistance of 60 ohms. While not recommended as good engineering practice, there are many practical reasons why it's necessary to install a dipole at less than optimum height. If so, what are the implications of feeding a 60-ohm antenna with either 52 or 75-ohm standard coaxial cable? Some calculations are revealing:

For 75-ohm coaxial cable:

$$VSWR = Z_o/R_r$$

$$VSWR = 75 \text{ ohms}/60 \text{ ohms} = 1.25:1$$

FIGURE 3



Feedpoint impedance versus height above ground.

ATTENTION: WOMEN WHO SOUGHT EMPLOYMENT WITH THE VOICE OF AMERICA (VOA), THE UNITED STATES INFORMATION AGENCY (USIA), OR THE UNITED STATES INTERNATIONAL COMMUNICATION AGENCY (USICA) BETWEEN OCTOBER 8, 1974 AND NOVEMBER 16, 1984.

YOU MAY BE A VICTIM OF SEX DISCRIMINATION ENTITLED TO A MONETARY AWARD AND A POSITION WITH THE AGENCY.

UNITED STATES DISTRICT COURT FOR THE DISTRICT OF COLUMBIA

CAROLEE BRADY HARTMAN, et al.,
Plaintiffs,

v.

CHARLES Z. WICK,
Defendant

Civil Action No. 77-2019
Judge Charles R. Richey

PUBLIC NOTICE

On November 16, 1984, the United States District Court for the District of Columbia found in this class action lawsuit that the United States Information Agency (USIA or the Agency), including the Voice of America (VOA), is liable for sex discrimination against female applicants for the following positions at the Agency. The USIA was also formerly known as the United States International Communication Agency (USICA). On January 19, 1988, the Court issued its opinion ordering relief in a variety of forms to potential class members. Accordingly, this case is now in the remedial phase.

JOBS COVERED

Specifically, the Court has found that the Agency has discriminated against women in hiring in the following jobs:

- Electronic Technician (Occupational Series 856)
- Foreign Language Broadcaster (Occupational Series 1048)
- International Radio Broadcaster (Other) (Occupational Series 1001)
- International Radio Broadcaster (English) (Occupational Series 1001)
- Production Specialist (Occupational Series 1071)
- Writer/Editor (Occupational Series 1082)
- Foreign Information Specialist/Foreign Affairs Specialist/Foreign Service Information Officer/Foreign Service Officer (Occupational Series 1085 and 130)
- Radio Broadcast Technician (Occupational Series 3940)

WHO IS INCLUDED

All women who sought employment with the Agency in any of the jobs listed above between October 8, 1974 and November 16, 1984 and were not hired may be eligible for relief. Also included are those women who were discouraged from applying for these positions during that time period. Even those women subsequently hired by the Agency in some capacity may be entitled to participate in the remedial phase of this case.

Women who sought employment with the Agency as Foreign Service Officers or Foreign Service Information Officers may be eligible for different kinds of relief depending upon the date of application and whether they sought employment at the entry level or mid-level. Women who sought employment with the Agency as entry level Foreign Service Officers or Foreign Service Information Officers in the years 1974-1977 must use the procedure outlined below. Women who sought employment with the Agency as mid-level Foreign Service Officers or Foreign Service Information Officers in the years 1974-1984 must also use the procedure outlined below. However, women who sought employment with the Agency as entry level Foreign Service Officers or Foreign Service Information Officers in the years 1978-1984 cannot use the procedure outlined below, since the Court has ordered an alternative form of relief for them and selected women in this group will be notified individually as to their rights.

RELIEF AVAILABLE AND HOW TO OBTAIN IT

Relief available to class members may include a monetary award and/or priority consideration for a current position with the Agency. If you think you may be entitled to relief, you must obtain a claim form, complete it fully, and return it to counsel for the plaintiff class, Bruce A. Fredrickson, Esq., Webster & Fredrickson, 1819 H Street, N.W., Suite 300, Washington, D.C. 20006 (202/659-8515), postmarked no later than July 15, 1989.

You may obtain a claim form in person and/or in writing from several sources: counsel for the plaintiff class, whose address is listed above; in person from USIA, Front Lobby, 301-4th Street, S.W., Washington, D.C. (8:15am-5:00pm), Office of Personnel Management (OPM), Federal Job Information Center (First Floor, Room 1425), 1900 E Street, N.W., Washington, D.C. (8:30am-2:30pm), or from area OPM offices throughout the country; in writing, VOA-Hartman, P.O. Box 400, Washington, D.C. 20044. You should carefully consider all questions on the claim form, sign it, and return it to counsel for the plaintiffs. Do not under any circumstances, return the claim form to the Judge, the Court or the Clerk of the Court. The Judge, the Court and the Clerk of the Court will not accept the claim forms and will not forward claim forms to plaintiffs' counsel.

PROCESSING OF CLAIMS

The process for handling claims has not been finally decided. Thus far, the Court has ordered that responding class members demonstrate their potential entitlement to relief at an individual hearing to be scheduled at a later date. However, the Court has reserved the right to reconsider this procedure in the event the number of claims filed makes this approach unmanageable.

Should individual hearings be used, you will be fully informed as to the date and time of your hearing. Moreover, you will be entitled to legal representation by counsel for the plaintiff class or his designee at no cost to you. Legal counsel will discuss your claim with you prior to your hearing, help you prepare your case and represent you at your hearing. You may, of course, retain your own attorney to represent you, if you so desire.

At the individual hearing, you will be asked to demonstrate your potential entitlement to relief by showing that you applied for one or more of the covered positions during the period October 8, 1974 and November 16, 1984 and that you were rejected, or that you were discouraged from applying. Evidence may be required in the form of testimony, documents, or both. Once you have demonstrated these facts, USIA is required to prove, by clear and convincing evidence, that you were not hired (for each position for which you applied) for a legitimate, non-discriminatory reason, such as failure to possess requisite qualifications. Should USIA make such a showing, you would then be entitled to demonstrate that the Agency's reason is merely a cover for sex discrimination or unworthy of belief.

Following the hearing, the Presiding Official will decide whether you are entitled to relief and, if so, what relief is appropriate. You may be entitled to wages and benefits you would have earned if you had been hired (back pay) from the date of your rejection until the date relief is approved. Under the law, back pay is offset by earnings you may have had during the period. In addition, you may be found to be entitled to front pay (that is, compensation into the future until an appropriate position is afforded you). Similarly, you may be found to be entitled to priority consideration for employment with the Agency. If hired, you may further be entitled to retroactive seniority with the associated benefits and the value of any promotions you would likely have had if you had not suffered discrimination.

REQUIRED STEPS TO FILE YOUR CLAIM

To participate in the remedial phase, you must fully complete the claim form and return it, POSTMARKED NO LATER THAN JULY 15, 1989, to counsel for the plaintiff class. Your failure to do so will result in your losing all rights you may have in this lawsuit. If you have questions about your rights or procedures available to you, you may contact counsel for the plaintiff class:

Bruce A. Fredrickson
Webster & Fredrickson
1819 H Street, N.W., Suite 300
Washington, D.C. 20006
(202/659-8515)

October 4, 1988

Date

/s/ Judge Charles R. Richey

United States District Court
Judge Charles R. Richey

For 52-ohm coaxial cable:

$$VSWR = R_p/Z_0$$

$$VSWR = 60 \text{ ohms}/52 \text{ ohms} = 1.15:1$$

In neither case is the VSWR created by the mismatch very significant.

The second approach is to mount the antenna at a convenient height and use an impedance-matching scheme to reduce the VSWR. You'll find information on suitable impedance-matching methods (including Q-sections, coaxial impedance transformers, and broadband RF transformers) in any good antenna textbook. Homebrew and commercial transformers can cover most impedance transformation tasks.

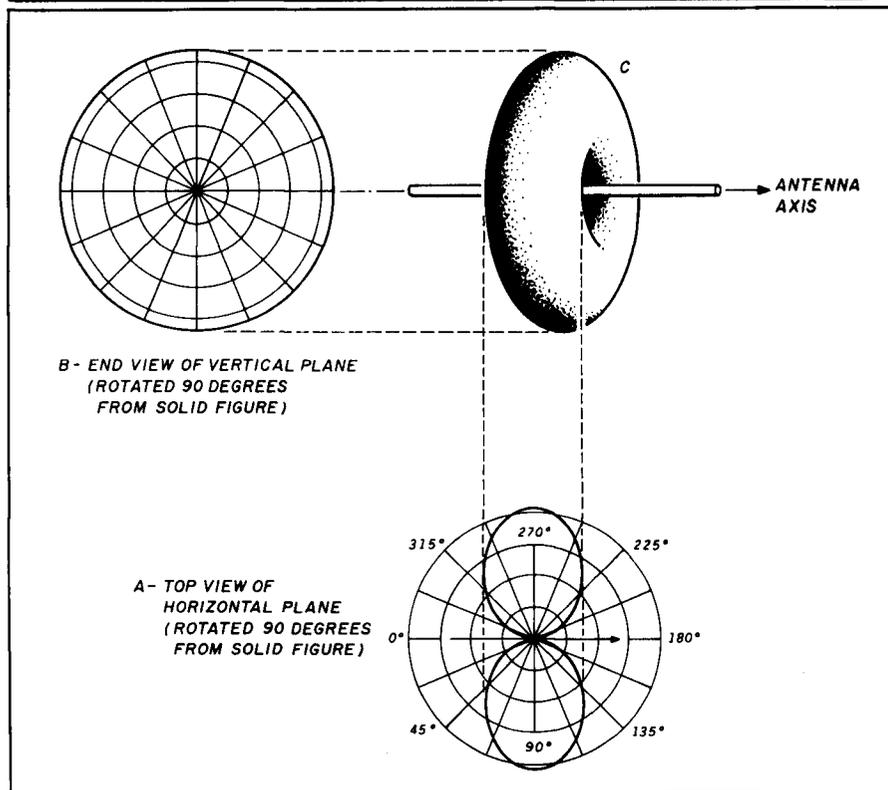
The third approach is to mount the antenna at a height (see Figure 3) where the expected radiation resistance crosses a standard coaxial cable characteristic impedance. The best height seems to be a half wavelength. The radiation resistance is close to the free space value of 72 ohms, and is a good match for 75-ohm coaxial cable (like RG-11/U or RG-59/U).

The dipole radiation pattern

When discussing antennas I keep returning to the concepts of *directivity* and *gain*, which are actually different expressions of the same fundamental concept. Antenna theory recognizes a point of reference called the *isotropic radiator*. This device is a theoretical construct consisting of a spherical point source of omnidirectional RF radiation. It creates an ever-expanding sphere as the RF wave front propagates outward. Antenna gain is a measure of how the antenna focuses available power away from a spherical wave front in a limited number of directions (two, for a dipole). This is how the concepts of directivity and gain are related.

Always remember that *directivity* and *gain* are specified in three dimensions. Many times authors (including me) simplify the topic too much by publishing only part of the radiation pattern (i.e., azimuth aspect as seen from above). You, in turn, wind up with a pattern viewed from above that shows the directivity in the horizontal plane. A signal doesn't propagate away from an antenna in an infinitely thin sheet, as such presentations seem to imply, but has an elevation extent in addition to the azimuth extent. Proper

FIGURE 4



Radiation pattern of dipole in free space as seen from two planes (A and B), and three dimensionally (C).

antenna evaluation takes both horizontal and vertical plane patterns into consideration.

Figure 4 shows the radiation pattern of a dipole antenna in free space "in the round." When the horizontal plane is viewed from above (Figure 4A), the pattern is a "figure eight" that exhibits bidirectional radiation. Two main "lobes" contain the RF power from the transmitter, with sharp nulls of little or no power off the ends of the antenna axis. This is the classic dipole pattern published in most antenna books.

I've also shown the vertical plane pattern for a dipole antenna in free space. Note that the radiation pattern is circular when sliced in this aspect (Figure 4B). When the two patterns are combined, you see a three-dimensional doughnut-shaped pattern (Figure 4C) that most nearly approximates the true pattern of an unobstructed dipole in free space.

When a dipole antenna is installed close to the ground and not in free space, as is the case at most stations, the pattern is distorted from that of Figure 4. You must take two effects into consideration. First and most important is that the signal from the antenna is reflected from the surface and

bounces back into space. This signal will be phase shifted by both the reflection and the time required for the transit to occur. At points where the reflected wave combines in phase with the radiated signal, the signal is reinforced; in places where it combines out of phase, the signal is attenuated. Thus the reflection of the signal from the ground alters the pattern from the antenna. The second factor to consider is that the ground is lossy, so not all of the signal is reflected; some of it heats the ground underneath the antenna. Consequently, the signal is attenuated at greater than inverse square law, further altering the expected pattern.

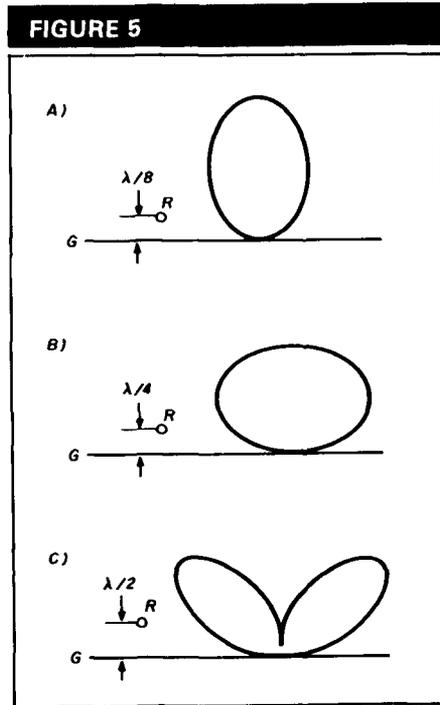
Figure 5 shows patterns typical of dipole antennas installed close to ground. The views in this illustration correspond to Figure 4B in that they are looking at the vertical plane from a line along the antenna axis. The antenna is represented by "R" in each case shown. Figure 5A shows the pattern for a dipole installed at one-eighth wavelength above ground. For this antenna, most of the RF energy is radiated almost straight up (now very useful). This type of antenna is basically limited to ground-wave and very short skip (when availa-

ble). The second case (Figure 5B) shows the pattern when the antenna is a quarter wavelength above the ground. Here the pattern is flattened, but still shows considerable vertically reflected energy (where it is useless). Now look at the pattern obtained when the antenna is installed a half wavelength above the surface. In Figure 5C, the pattern is best for long distance work because energy is redirected away from the vertical into lobes at relatively shallow angles.

Dipole construction and installation techniques

According to "conventional wisdom," the ideal dipole antenna should be installed at a very high altitude where its performance resembles the free space model. Unfortunately, complying with conventional wisdom is impossible — even for antennas in the higher end of the HF spectrum. Given that the dipole feedpoint impedance is a good match for 75-ohm coaxial cable, and that the pattern is ideal for long distance work when the antenna is installed at a height of a half wavelength above the surface, it's a good idea to

try installing the antenna at that height. Building and installing simple dipoles isn't terribly difficult. Figure 6

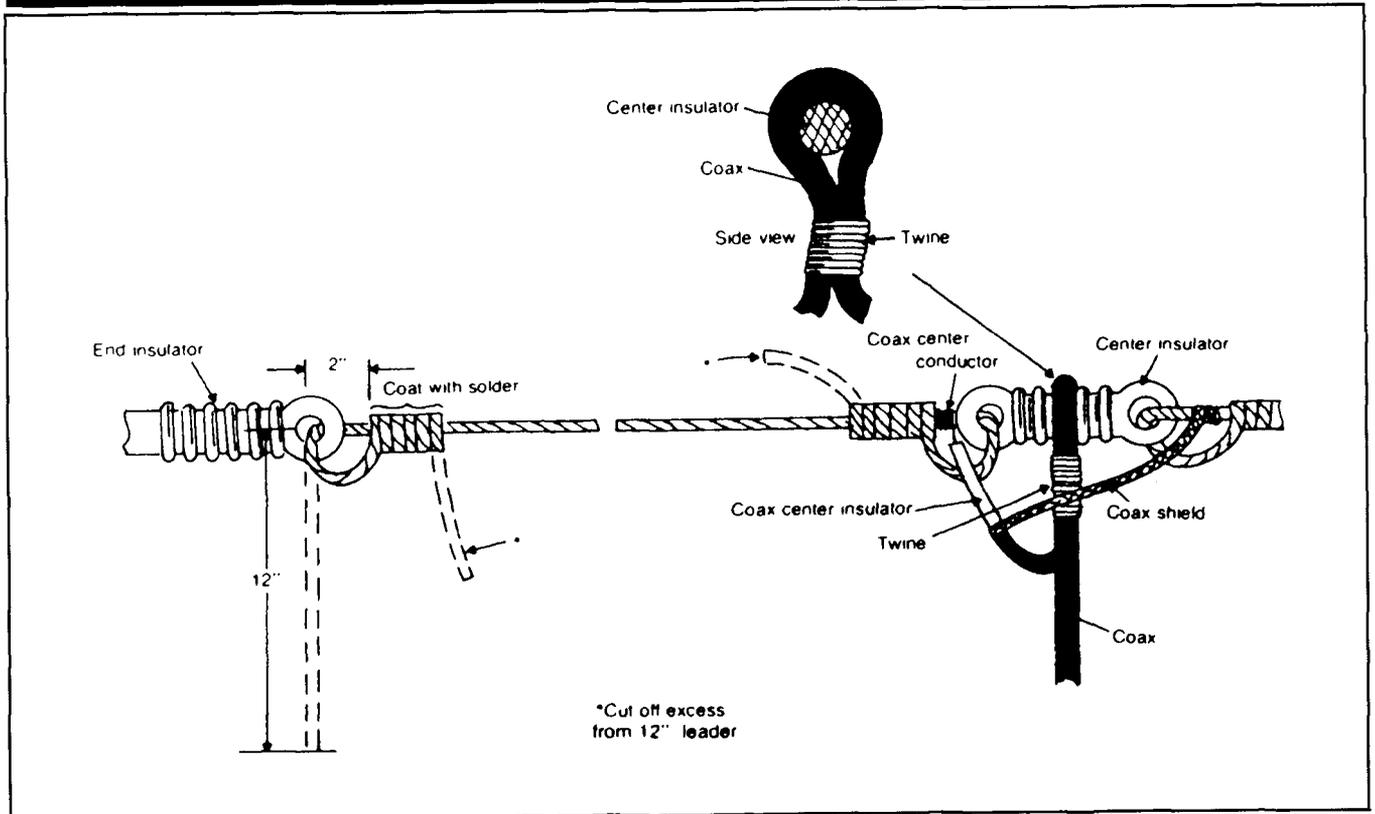


Vertical aspect radiation pattern of dipole close to earth's surface: (A) 1/8-wavelength, (B) 1/4-wavelength, and (C) 1/2-wavelength.

shows the method for building the antenna. First, cut the wire radiator elements to the approximate length indicated by Equation 2 plus an additional 12 to 24 inches; each element will finally be a quarter wavelength long. The wire can be either hard-drawn copper wire or Copperweld®. The latter is a special tough-service steel core antenna wire coated with copper. The RF resistance of this wire at frequencies above 1 MHz is the same as that of solid copper wire because of the "skin effect" (alternating currents like RF flow on the outer surface of the conductor only). At 160 meters the skin effect depth is only 50 microns (2 mils), while at 10 meters it's only 12 microns (0.5 mils). This means you have the advantage of copper conductivity along with the strength of steel wire.

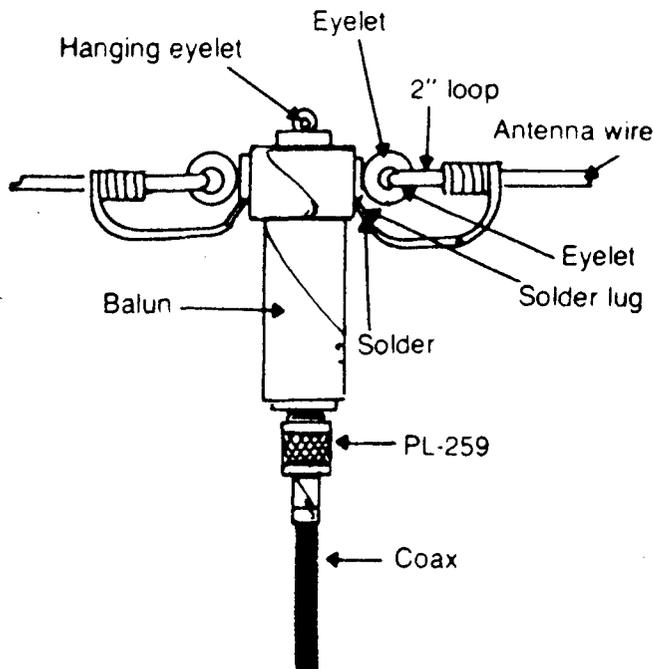
You'll need two end insulators, and both are assembled in the same way. Pass the wire through the hole in the insulator (see Figure 6) to a length of about 12 inches. Wrap the wire back on itself and wind it around the portion of the wire that's left on the other side of the insulator. Make this a permanent

FIGURE 6



Construction details of dipole antenna (from TAB Handbook of Radio Communications by J.J. Carr).

FIGURE 7



Use of a 1:1 balun transformer at the feedpoint (from *TAB Handbook of Radio Communications* by J.J. Carr).

connection by soldering it and clipping off the excess wire. The solder won't provide mechanical strength. Its purpose is to make a good electrical connection in the presence of corrosion.

Fix the antenna wires to the center insulator in the same way, unless you plan to use one of the special center insulators now on the market. Make these connections temporary until after you've tuned and tested the antenna. You may have to either lengthen or shorten the radiators when tuning your dipole.

Connect the transmission line (usually coaxial cable) to the antenna wire at the center insulator as shown. Attach the center conductor to one radiator element and the shield of the coax to the other. You need to provide strain relief for the coaxial cable; if you don't the cable will break after only a short period of service. The easiest strain relief method is shown in Figure 6. Simply wrap the cable once around the insulator and tie it off with twine.

Some commercial center insulators offer a strain relief hole or other mechanism. Many people prefer to use a 1:1 balun transformer at the dipole's feedpoint (see Figure 7). The transformer has a 1:1 impedance ratio, so it doesn't provide any matching. Instead, it's said to balance the currents flowing in the two radiators, and prevent radiation from reaching the feedline. While this claim has been controversial for some time, and the issue is still not resolved, the best evidence suggests that the pattern of a dipole close to ground is most nearly like the ideal pattern if a 1:1 balun transformer is used at the feedpoint. In Figure 7 the balun transformer also acts as the center insulator, so no other arrangement is needed.

Next month...

This month I looked at the basic resonant dipole. In part 2, I'll discuss tuning methods for the standard dipole, and some additional variations on the dipole theme. 

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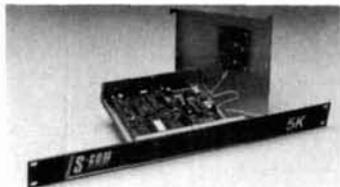
If you know how to solder, this problem is *easily* corrected by replacing the original, marginally effective [High VHF-Q] "parasitic suppressors" (?) with **Low VHF-Q Parasitic-Suppressors**. [For more information, see *QST Magazine*, Oct. 1988, page 36] All materials, components, instructions, diagrams and a 430°F silver-solder kit included, nothing else to roundup, **Low VHF-Q Parasitic-Suppressor™** amplifier retrofit-kits are now available from the author of this article. Prices start at \$12, delivered via First Class Mail®, for a (1) or (2) 3-500Z, or 3-400Z, amplifier retrofit-kit. For increased duty-cycle option, add \$2. **Parasitic-kits** are also available for HF-amplifiers that use 572B, 8122, 8873, 8874, 8875, 3CX800A7, 3-1000Z, 3CX1200A7 and 8877 tube types.

▶▶▶ Also available: Telephone RF Interference filter kits with diagram and instructions, four (4) for \$5, delivered via First Class Mail®.

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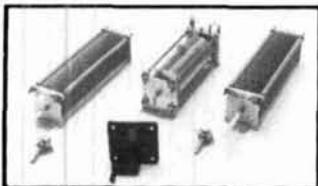
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PRODUCT REVIEW

Digital TWR-3 Weather Station

Like most New Englanders, I'm very interested in the weather. I wanted to own weather instruments that would provide good data, but found the cost of most systems prohibitive because of fancy features.

The Digital series of handheld weather stations meet my standards. The TWR-3, advertised as the world's smallest computer weather station, "packs a wallop" of information including wind speed (3 to 250 mph), wind direction (in degrees, two scales), wind gust record, temperature (-70 to +270 degrees F), high/low temperature record, and has an optional rainfall gauge at extra cost.

The TWR-3 reads in English or metric units and can be programmed to scan through its various functions. It operates on house current, 12-volt DC supply, or its own internal battery support. I installed it easily in less than an hour.

The model TWR-3 is made by Magnaphase Industries, Inc., and is available from Azimuth Weather Star, 11845 W. Olympic Blvd., Suite 1100, Los Angeles, California 90064, for \$159.95.

de N1GCF

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Mobile data unit

The TEMPO MPP1 TNC/printer combination is a compact unit for mobile or portable use. The processor portion of the unit is compatible with TAPR TNC 2 and makes use of the complete command set. With 32K ROM and 32K RAM it's possible to:

- store and print all messages received while operating.
- store all messages for printing within the ROM unit.
- store messages to a selected terminal for immediate or delayed printing.

Specifications:

Protocol

AX.25 level 2

Modem

AFSK (1200-2200 Hz)

Processor

Z80 software compatible

Memories

ROM 32K, RAM 32K (lithium battery backup)

Communication speed

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Connections Radio interface (5 pin)

Terminal interface (DB-9)

Power DC (2 pin)

The mobile data terminal is shipped complete with cables for connections to a transceiver, computer serial port (RS 232C), and DC power source. Also included are the installation and user's manual and a spare roll of thermal paper. An AC Adapter for supplying DC power from the 117 VAC line and a technical manual complete with schematic are optional. The technical manual is recommended for understanding the circuitry. It's very helpful if you're just getting started in packet, as the installation and user's manual is brief.

Instructions for connecting the unit to your VHF transceiver and computer are straightforward and well documented. You'll have to provide the proper radio and computer plugs to match your equipment, but the MPP1s are already mounted on their cables with the other ends left as flying leads.

I tested the unit with three different transceivers. ICOM-25A, Kenwood TM-221A, and Kenwood TM-621A. The ICOM and Kenwood units had different connections, so I made an adapter to interchange the units easily. I also used two different computers, an IBM PC-XT and a Radio Shack Model 100 (lap top) unit.

Initialization of the MPP1 was easy and well outlined in the manual provided. You must load the terminal's RAM with the proper defaults and your callsign on initial setup. Then your terminal is ready for base or remote use.

I did base station testing with the transceivers connected to a stacked pair of Yagis. I made contacts with stations as far away as Montreal, Canada by using digipeaters. The unit was very tolerant of audio level variations, and did not drop messages during periods of fading or when the audio level was intentionally varied using the gain control. My mobile testing included traveling a route that passed through known weak signal and multipath areas. I copied a couple of bulletin boards without fault and left messages for other users.

This unit is ideal for emergency communications. Amateurs involved in ARES activities should consider it for remote operation. The TNC portion of the unit is a complete processor in itself and directly controls the printer; it can be used as a receive only monitor for messages, with printout activated when convenient.

My only difficulties were due to "cockpit errors" because I have limited packet experience. It's well worth the extra cost to order the technical manual.

I'd like to thank Bill Burden, WB1BRE, for riding copilot and operator during the mobile testing.

The MPP1 sells for \$395 and is available from HENRY RADIO, 2050 S. Bundy Drive, Los Angeles, California 90025.

de WA1TKH

Circle #302 on Reader Service Card.

Cushcraft 124WB-element, 2-Meter boomer antenna

Here's a neat compact antenna that can be used in a number of applications. It's perfect for packing to the top of a mountain or high hill; fire it up on 2-meter SSB, or use it for repeater DXing. Apartment dwellers can sneak this antenna into almost any location and get the benefit of directivity and gain. For me, it simplified connecting into the local DX-spotting packet network.

In the past, I had been using either a horizontally polarized antenna or a 5/8-wave vertical. Unfortunately, anytime I rotated the tri-band beam, I lost the packet network. There was also a 20-dB signal loss between my horizontal beam and the packet cluster's vertically polarized antenna. The vertical, well, it never worked right.

Cushcraft's 4-element boomer is elegant in its simplicity. Construction is straightforward and takes just a few minutes. Because of the small size, this antenna can be shoehorned into almost any location.

The 124WB will tune 144-148 MHz with a less than 2:1 SWR. Cushcraft rates the forward gain at 10.2 dBd, with a front-to-back ratio of 19 dB. Assembled wind area is less than 6 inches and the antenna weighs less than three pounds. The retail price is \$60. If you want more gain, stacking instructions are included.

From the top of the tower to the attic (where mine is), installation is not a problem and takes just a few minutes. Later this spring, I'll move the antenna to the tower to gain a few additional vertical feet. Now that I'm using the 124WB, I can get into the packet network with ease. The forward gain and directivity gives me a better chance at connecting into the network, even during its busy times.

For further information contact Cushcraft Corp., P.O. Box 4680, Manchester, New Hampshire 03108.

de N1ACH

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Other features include: Panel-selectable RF preamp and attenuator, dual VFOs, noise blanker, RIT, semi-break in CW, selectable AGC, full-duty cycle, and optional narrow CW filter.

The suggested retail price of the IC-725 is \$949.

For more information contact ICOM America, Inc., 2380 116th Ave, N.E., PO Box C-90029, Bellevue, Washington 98009-9029.

Circle #304 on Reader Service Card.

Handheld frequency counter

Optoelectronics, Inc. introduces its new handheld frequency counter model 2210. It has low frequency coverage down to 10 Hz and microwave coverage up over 2.2 GHz. The counter runs on internal NiCd batteries and comes with a metal cabinet and precision quartz timebase oscillators. A full line of accessories includes antennas, probes, and carry case.



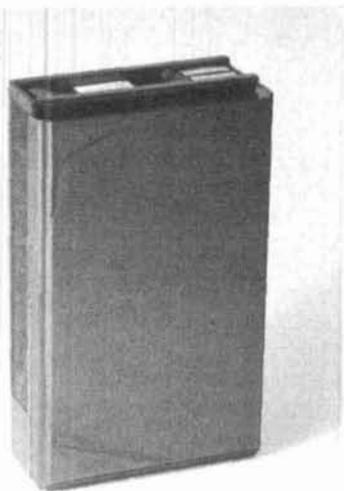
Input sensitivity is less than 10 mV from 10 Hz to 2 GHz; 3 mV is typical. Accuracy is 1 PPM with temperature compensated crystal oscillators. Resolution is 1 Hz below and 100 Hz above 12 MHz. A full 16-hour recharge yields 2 hours of battery operation. Use an AC adapter/charger to operate when recharging.

The model 2210 sells for \$189 complete with NiCd batteries and charger. The model TA-100S telescoping whip antenna is \$12 and the vinyl carry case is \$10. For more information contact Optoelectronics, Inc., 5821 N.E. 14th Avenue, Fort Lauderdale, Florida 33334.

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Periphex, Inc. offers battery packs that are compatible with the following Yaesu radios: FT-727R, 109RH, 209R/RH, 709R, 103R, 203R and 703R.



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For more information contact Periphex, Inc., 149 Palmer Road, Southbury, Connecticut 06488.

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WeFaxWorks is a program for the Apple Macintosh computer that allows reception of weather maps and charts on a standard Macintosh using Kantronics' KAM, all-mode interface. During live reception, the screen scrolls automatically. Synchronization is simple; point the mouse at the sync mark and click once. This causes the received and displayed signal to be aligned with the left edge of the Macintosh screen.

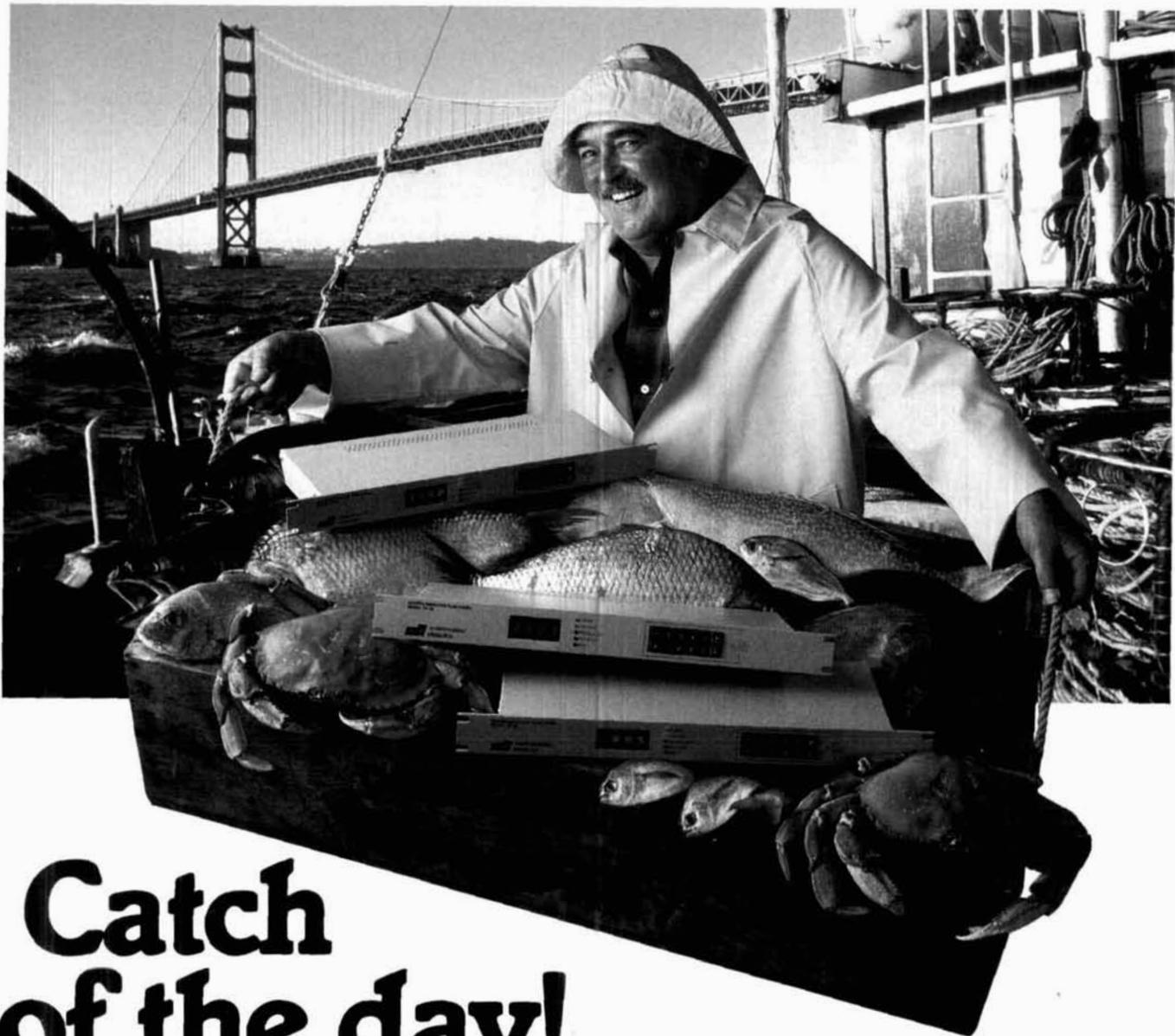
The entire picture is saved into a buffer as it's being received. Once you exit the receive mode, you can select the buffer display, and the captured picture will appear on the screen.

The buffer contents can be saved to disk by choosing "Save..." from the File menu, and a saved WeFaxWorks map can be recalled from disk with the "View File..." selection. The buffer contents are erased when you enter the "Receive Live" mode, or when a map is recalled from the disk.

Maps are printed with MacPaint format files. The portion you want to print is brought to the screen using the Display Buffer mode and scroll bar. These files are in MacPaint format and can be opened, touched up, and printed with MacPaint.

For more information contact Kantronics at 1202 E. 23rd Street, Lawrence, Kansas 66046.

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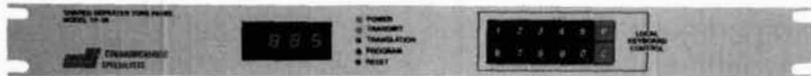


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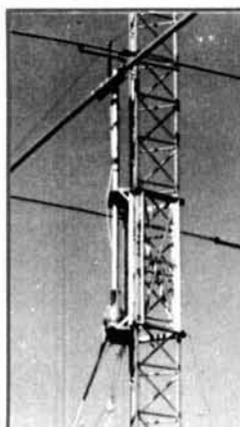
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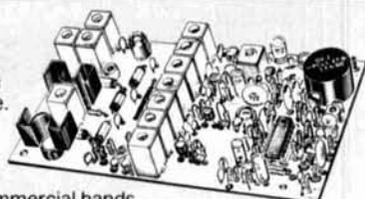
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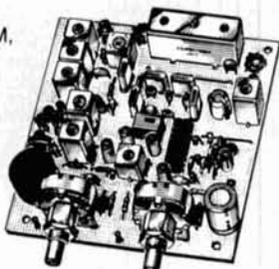
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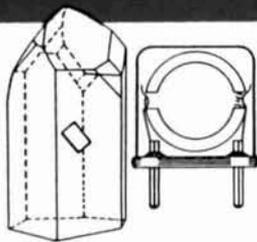
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DX Forecaster

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1988 PROPAGATION SUMMARY

September 1986 was definitely the month of sunspot minimum, with the smoothed sunspot number (SSN) equal to 12.3. That makes 1988 the second year of solar cycle 22. The solar flux minimum of 67.6 occurred in June 1986, but September of that year also brought the smoothed value minimum of 72.9, along with the sunspot minimum. In 1988, the sunspot number started out at 58 and ended at about 137 while the solar flux started at 108 and ended at 200. These numbers made it the steepest climbing year of the cycle as SSN and solar flux are expected to taper off to a maximum of 185 for SSN and 225 for the solar flux near the end of 1989. While the 1987 values were slightly above the highest in cycle 19, the 1988 values' slope crossed cycle 19's, so these forecast values are probably reasonable.

In 1988, midlatitude noontime maximum usable frequencies (MUFs) monthly median increased from 19 to 27 MHz for a 3000-km hop. The increase wasn't linear throughout the year because the summer F2 layer was 15 percent less than it was the rest of the time, with the equinoxal periods having the highest MUF. This increase in MUF levels off for SSNs greater than 150, as does the solar flux. Most solar flux to SSN conversion formulas don't take this into account. This also shows how the MUF follows solar flux values above 150 instead of the SSN; this will be the condition during 1989. The MUF formula for monthly medians is $MUF = 2.65 \times (0.0165 \times SSN + 8.4)$.

During 1988, propagation was affected adversely by several periods of geophysical events. In March I discussed how to forecast propagation conditions a day or two in advance using the trend in solar flux and geomagnetic A values. Now I'd like to show how MUFs correlated with flux and A indexes during several large events in 1988. The first occurred on



February 20th during a decreasing solar flux (107 to 102). A small solar flare started a high-latitude geomagnetic event which spread down to the midlatitudes by the 22nd. The A value went to 67 and decreased MUFs 48 percent to 10 MHz. The MUFs were 25 percent on the 23rd, 19 percent on the 24th, and 15 percent on the 25th before recovering. On March 25th, a small flare was probably the cause of a polar cap absorption and small geomagnetic disturbance. The MUF increased 33 percent over the first two days, then decreased 17 percent for the next two. The solar flux had just increased 7 units, but was level during the disturbance. A corresponding increase in MUF occasionally occurs as it did here, when the solar flux is on the increase or if the disturbance starts as the sun is rising on the propagation path. The next notable event began with a gradual disturbance of unknown cause on April 3rd through the 7th, which dropped the MUFs 46 percent during an A of 57. The solar flux was decreasing from 128 to 115.

Another large disturbance (A index = 63) began gradually near the end of May 5th and lasted one day. The MUF decreased 66 percent before it was over. The solar flux decreased 5 units during this period. The last significant event affecting propagation was on October 10th, caused by a small solar flare and solar flux burst. The geomagnetic disturbance measured 57 on the A index and the MUF decreased nearly 40 percent as the solar flux was going down. You'll notice that most of these events happened as the solar flux was decreasing and that the MUF decrease was greater at those times than it was during those disturbances when the solar flux was increasing. The MUF decrease averages 2 percent per A unit when the

solar flux is decreasing, as opposed to a 0.8-percent MUF decrease per A unit when the solar flux is level or increasing. The solar flux factor for the beginning of 1988 was 1 percent MUF change per unit of solar flux; later in the year it increased to 2 units of solar flux for a 1-percent MUF change.

You can use these factors during this phase of the SSN cycle to predict the best band for daily operation. There are more bands to jump to now and in July. Good luck!

Last-minute forecast

The higher frequency bands (10 to 30 meters) are expected to have the best openings the first and the last week and a half of May. These openings may include some 6-meter long skip when WWV solar flux values indicate the very peak of the 27-day solar cycle. Transequatorial (TE) single long-hop openings will probably be available towards the evening early in May; these openings are scarce during June, July, and August. Periods of disturbance around the 5th and 13th may enhance the possibility of TE openings. Another expected disturbance date, the 22nd, may come too late to help TE but may affect east-west paths on the lower bands. These lower bands should be best the second and third weeks of the month, when the solar flux is expected to be at minimum. The higher minimums restrict daytime DX distance from weak signals.

The full moon occurs on the 20th; the lunar perigee is on the 3rd and 31st. An Aquarid meteor shower (for meteor scatter and meteor burst DXers) peaks between May 4th and 6th, with rates of 10 and 25 per hour for the northern and southern hemispheres, respectively.

Band-by-band summary

Ten, 12, 15, 17 and 20 meters will support DX propagation from most areas of the world during daylight hours and into the evening, with long skip out to 2000 miles (3500 km) per hop. Signals on the upper three bands will be strongest from the southern countries and occur near local noon.

The propagation direction will follow the sun across the sky. It will be to the east in the morning, the south at mid-day, and the west in the evening. Sporadic-E short skip will be available at local noon on some days toward the end of the month.

Thirty, 40, 80, and 160 meters are the nighttime DXers' bands. The direction of propagation follows the darkness path across the sky: evening to the east, around midnight to the north and south, and toward the west in the predawn hours. Distances will generally decrease to 1000 miles (1600 km) for skip on these bands. Sporadic-E openings will be most frequently observed around sunrise and sunset toward the end of the month. 

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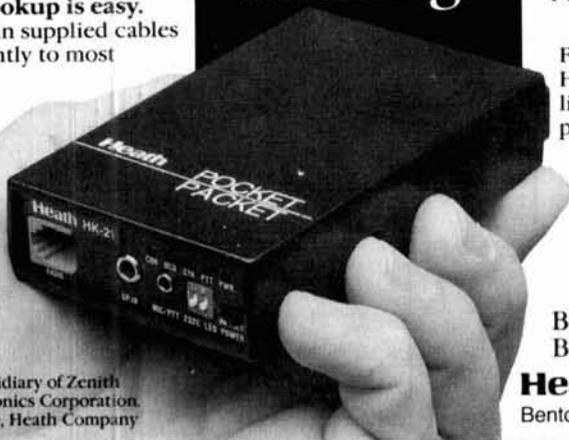
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AZDEN SERVICE by former factory technician. Rush service available. PCS 300 Nicad \$36.95. Southern Technologies Amateur Radio, Inc, 10715 SW 190 St, #9, Miami, FL 33157. (305) 238-3327.

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COMMUNICATIONS BATTERIES: Clone Packs! Ready-for-use. ICOM: BP5 \$44.95, BP-3S 2X cap. BP3 "Wall Chargeable" \$43.95, BP-7S 2X BP7 (SW only) or BP8S (BP8 + 50%) (Base Chg-only) \$67.95, YAESU: FNB2 \$22.95, FNB10S (10 + 60%) \$49.95, SANTEC 142 \$23.95. * Repair Inserts * ICOM: BP2 \$18.95, BP3 \$16.95, P5 \$24.95, BP7/BP8 \$29.95, KENWOOD: PB21 \$13.95, PB24 \$21.95, PB25/26 \$25.95, YAESU: FNB4/4A \$33.95, TEMPO: S1, 2, 4, 5, 15/450 \$23.95, AZDEN: 300 \$21.95. * E.P. Porta-Pac & Chgr 12V/5Ahr \$44.95 * * Rebuilding * Send pack—free estimate. * Antennas * Ducks/BNC \$8.95, 2mtr 5/8-Tel/BNC \$18.95. SASE Catalog, PA + 6%. \$3 Shipping/order. VISA/M/C + \$2. (814) 623-7000. CUNARD ASSOCIATES, Dept H, RD 6, Box 104, Bedford, PA 15522.

ENGINEER WANTED. With Ham experience to develop Ham products. Exciting proposition for the right person. Box 498, Greenville, NH 03048.

450 MHz SPECTRUM ANALYZER. Adapted from Nov 85 QST article by Al Helfrick, K2BLA. Use your low frequency scope for the display portion. Log output calibrated in 10 db steps. For complete kit, order #450-KIT \$459.95 plus \$4.50 s/h. Calif. residents add 6% sales tax. Foreign orders add 15% for shipping. For additional information send large SASE to: A & A ENGINEERING, 2521 W. LaPalma, #K, Anaheim, CA 92801 or call (714) 952-2114.

FOR SALE: Browning Golden Eagle Mark IV AM/SSB citizen's band receiver. Superb performance—tube-type double conversion, low noise, two tunable bands or crystal controlled. Use on CB, retune for 10 meters, or use as tunable IF for 2 meter, satellite or microwave receiver. Missing top cover, otherwise complete and working, with schematic. \$100.00. Peter Ferrand, WB2QLL, 65 Atherton Avenue, Nashua, NH 03060. (603) 889-1067.

WANTED: All types of Electron Tubes. Call toll free 1-800-421-9397 or 1-612-429-9397. R & N Electronics, Harold Bramstedt, 6104 Egg Lake Road, Hugo, MN 55038.

100 QSL CARDS \$8. \$3 thereafter. Grid square printed free. Shipped postpaid within two weeks. Guaranteed correct! Free samples. Shell Printing, KD9KW, Box 50B, Rockton, IL 61072.

SCHEMATICS. Devices, modules and components. Catalog \$1.00 refundable. Free flyer LSASE George Whitmore, 5746 Aberdeen Angus Way, Las Cruces, NM 88001.

N6SR/KX6: Anyone who has worked me as /KX6 has worked an illegal station. Anyone hearing that call please notify me and the FCC immediately.

WANTED: G.E. Mastr-Pro UHF 250 watt base station or amplifier with power supply or amplifier along. K8RUR, (313) 697-8886.

FOR SALE: IC-2AT 5/8 whip, rubber duck ant., mike, IC-BP4 battery pack, IC-DC1, two IC-BP3's with charger. \$200. Martin Hanft, POB 199, Gilsum, NH 03448.

UHF PARTS. GaAs Fets, mmics, chip caps, feedthrus, teflon pcb, high Q trimmers. Moonbounce quality preamps. Electronic sequencer boards. Send SASE for complete list or call (313) 753-4581 evenings. MICROWAVE COMPONENTS, PO Box 1697, Taylor, MI 48180.

COMMODORE-128 PROGRAM available to track the Amateur Satellites. Uses Keplerian data supplied by NASA free. Tracks up to 8 satellites simultaneously. Program also supports printing schedules and predictions for satellites. Use it to track MIR and talk to the Cosmonauts. SATRAK128, \$26.50 includes shipping. Other information on this or other programs for the C128, requires a business size SASE. Reid Bristor, WA4UPD, PO Box 0773, Melbourne, Florida 32936-0773.

WANT: 32S3 xmt, 250TL and 304TL tubes. KF6WM, 45300 Royal, King City, CA 93930.

DXERS—CUSTOMIZED PRINTOUT of antenna headings calculated for your location. List includes over 650 worldwide locations. Send Lat/Long coordinates, name, callsign, check for \$12.95 U.S. Brian Henderson, VE6ZS, 23 Deermoss Pl SE, Calgary, Alberta, Canada T2J 6P5. (403) 278-2084.

HANDICAPPED NOVICE needs HF equipment donated—anything please. KA3OUE, (412) 531-7443 anytime.

OFFICIAL MILITARY-TYPE ID TAGS. ("Dog Tags")! Customized with your Call Letters, etc. 5 seventeen space lines. 20" nickel plated chain included. \$4.29 postpaid. JPW ENTERPRISES, PO Box 353, Logan, Utah 84321

MAGAZINES WANTED: "Microwave Systems News" (MSN), "RF Design", "PCIM (Power Conversion & Intelligent Motion)" and "QEX" (1980-present). Call collect 519-742-4594 (Ontario) after 6 PM Eastern time.

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.

BACK ISSUES OF HAM RADIO. Have most issues from 1969 to 1974. Mint condition. \$3.00 for single issues. WN0G, 319-377-3563.

HAM TRADER YELLOW SHEETS. In our 27th year. Buy, swap, sell ham radio gear. Published twice a month. Ads quickly circulate—no long wait for results. Send NO. 10 SASE for sample copy. \$13 for one year (24 issues). PO Box 2057, Glen Ellyn, IL 60138-2057 or PO Box 15142, Dept HR, Seattle, WA 98115.

VHF-UHF-SHF. Large SASE. West Coast VHFer, POB 685, Holbrook, AZ 86025.

CHASSIS & CABINET KITS. SASE. K3IWK, 5120 Harmony Grove Rd, Dover, PA 17315.

ANALOG AND RF CONSULTING for the San Francisco Bay area. Commercial and military circuits and systems. James Long, Ph.D., N6YB (408) 733-8329.

RTTY JOURNAL—Now in our 36th year. Read about RTTY, AMTOR, PACKET, MSO'S, RTTY CONTESTING, RTTY DX and much more. Year's subscription to RTTY JOURNAL \$10.00, foreign slightly higher. Order from: RTTY JOURNAL, 9085 La Casita Ave., Fountain Valley, CA 92708.

RUBBER STAMPS: 3 links \$5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birdlake Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

"HAMLOG" COMPUTER PROGRAM. Full features, 17 modules. Auto-logs, 7-band WAS/DXCC. Apple \$19.95. IBM, CP/M. KAYPRO, Tandy, C128 \$24.95. HR-KA1AWH, POB 2015, Peabody, MA 01960.

WANTED: ARC-5 and SCR-274 equipment, parts and accessories, any condition. Ken, WB9OZR, 362 Echo Valley, Kinross, NJ 07405. (201) 492-9319.

WANTED: Ham equipment and other property. The Radio Club of Junior High School 22 NYC, Inc. is a nonprofit organization, granted 501(C)(3) status by the IRS, incorporated with the goal of using the theme of Ham Radio to further and enhance the education of young people. Your property donation or financial support would be greatly appreciated and acknowledged with a receipt for your tax deductible contribution. In Dayton, meet the crew from 22 and relax at our flea market tables, check in on 144.30 simplex. Please write us at: PO Box 1052, New York, NY 10002. Or call our round the clock hotline: (516) 674-4072. Thank you!

WANTED: Drake Linear Amp Model MN4439-1000W (2000 PEP), 1.8-30 MHz. Call Bruno Molino, VE2FLB, 26 Rue Des Anciens, Gatineau, Quebec J8T 3T2. (819) 561-3689.

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Announces The Next Generation

The 10 amplifiers you have been waiting for!

Designed For Quality And Value!

Every effort has been made in the design of these amplifiers to offer the highest specifications possible, provide the ultimate in reliability, and still keep prices affordable. Compare these amps with all others on the market! You'll be glad you waited for the N E X T generation of solid-state amplifiers from MIRAGE/KLM!

144 MHz Amplifiers
B-1016-G 10W in = 160W out
B-3016-G 30W in = 160W out
B-215-G 2W in = 150W out

1-6

13.8 vDC

220 MHz Amplifiers
C-1012-G 10W in = 120W out
C-3012-G 120W out
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New protection circuitry automatically reduces the output power to prevent damage to output transistors and even returns the amplifier to full power automatically when problem is cleared!

New GaAs-FET pre-amp designs provide gain of over 25 dB and a noise figure of less than .6 dB!

Picture this . . . You know your station . . . You are at home with your gear . . . all the knobs, switches, meters . . . QSY's are no big deal, you could do them in your sleep (and you probably have!).

Now picture this . . . It's contest time, multi-op . . . do you worry about your gear? . . . NO! At least not your amps . . . your station amps are bullet-proof. Point and shoot, no tune, no touch. From 160 meters to 70 cm . . .

Your Amps Are Mirage!

Each of the four following amplifiers provide . . .

Bullet-proof, thermal shutdown . . . VSWR shutdown . . . over-current shutdown . . .

120% ICAS duty cycle . . . air-cooled . . . fan hood available . . .

Active cooling kit available for 100% key-down duty cycle

TWO 144-MHz Amplifiers

Finally, a ruggedized high-speed RF switching relay that takes the punishment SSB-op's demand . . . 5mS or less typical switch-time . . . Dual-gate GaAs MES-FET pre-amp . . . 22 dB typical gain . . . Wide, dynamic range for overload protection . . . 1 dB compression > +4 dBm

7

30W in - 300W out

(Linear curve: 1W - 30W, 45W max.)
13.8 vDC 32 amps max.
440 watts (DC) 68% efficiency

8

30W in - 600W out

24v DC

TWO H-F Amplifiers All Solid State (Waiting FCC Type-Acceptance)

The Band-Pass filter allows wideband performance while meeting FCC specifications . . . 1.8 - 4.0 MHz, 4.0 - 9.0 MHz, 9 - 15 MHz, 15-30 MHz . . . Typical harmonic - 50 dB

9

50W in - 800W out

13.8 vDC
1,215 watts (DC)
88 amps
Available with power supply

10

50W in - 1,500W out

48.0 vDC
110/220 - 50/60 cycles
Auto-Band switch Vacuum Relay
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Power Supply included

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COMING EVENTS

Activities — "Places to go . . ."

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

May 5-6: NEBRASKA: Hamboree #11 sponsored by the 3900 Club and the Sooland ARA, Marina Inn, South Sioux City. For reservations write Al Smith, W0PEX, 3529 Douglas Street, Sioux City, IA 51104.

May 7: NEW JERSEY: Annual Indoor Hamfest/Flea Market, sponsored by the Tri-County Radio Association, Passaic Township Community Center, Stirling, 8AM to 2PM. For information/reservations Dick Franklin, PO Box 182, Westfield, NJ 07090. (201) 232-5955.

May 6: NEW YORK: 30th annual Southern Tier Hamfest sponsored by the Southern Tier ARCS (STARC), Tioga County Fairgrounds, Rt 17c, Owego. For information/tickets SASE to STARC, PO Box 7082, Endicott, NY 13760.

May 6: KANSAS: Tailgate Swapfest sponsored by the Flint Hills ARC, Augusta City Park, 20 min east of Wichita. 8AM to 3PM. For information SASE to Zack Wilkerson, K0DZY, Rt 1, Box 90, El Dorado, KS 67042.

May 6: WISCONSIN: The Ozaukee Radio Club will sponsor its 11th annual Swapfest, Circle B Recreation Center, Highway 60, Cedarburg, 8 AM to 1 PM. For information send business SASE to ORC Swapfest, N5415 Crystal Springs Court, Fredonia, WI 53021.

May 6: NEW YORK: The Putnam Emergency Amateur and Radio League will have their PEARLFest at the John F. Kennedy Elementary School, Foggintown Road, Brewster. 9 AM to 4 PM rain or shine. For registration contact Terri Culm, N2GWF, 40 Mile Hill Road, Highland, NY 12528 or Jim Morgan, KA2FIQ, 39 Overlook Road, Ossining, NY 10562.

May 6-7: MARYLAND: Capital Fest sponsored by the Timex Sinclair Computer Club, Howard Johnson's, Rt 95 & 450, New Carrollton. For information Audrey Curnutt, 10400 Truston Road, Adelphi, MD 20783.

May 5-7: ARIZONA: The Cochise Amateur Radio Association's annual Hamfest, Club training facility, Sierra Vista, tailgating, Handi facilities. For information N7JNK (602) 378-3155 after 6 PM or write CARA, PO Box 1855, Sierra Vista, AZ 85636.

May 7: NEW JERSEY: Spring Hamfest sponsored by the Bergen ARA, Bergen Community College, 400 Paramus Rd, Paramus. 8-3. For information Jim Joyce, K2ZO, 286 Ridgewood Blvd, No. Westwood, NJ 07675. (201) 664-6725.

May 6-7: SOUTH CAROLINA: 50th annual Greenville Hamfest sponsored by the Blue Ridge Amateur Radio Society, American Legion Fairgrounds, Greenville. Saturday 8-5; Sunday 8-3. For advanced tickets or information SASE to Blue Ridge ARS, POB 6751, Greenville, SC 29606.

May 13: WISCONSIN: Lakeshore Hamfest sponsored by Manarad Radio Club, Manitowoc County Expo Center, Hwy 42-151 and I-43 on County Hwy R. Starts 8 AM. Contact: Manarad Radio Club, PO Box 204, Manitowoc, WI 54220.

May 14: OHIO: Medina County Hamfest sponsored by the Medina 2 Meter Group, Medina County Community Center, 735 Lafayette Road, Medina. 8AM to 2PM. For information/tickets SASE to Medina Hamfest Committee, PO Box 452, Medina, OH 44258. (216) 769-3033 or 725-4492. 10AM to 5PM.

May 14: OHIO: 10th annual Hamfest sponsored by the Athens County ARA, City Recreation Center, Athens. 8AM to 3PM. For information Carl J. Denbow, KA8JXG, 63 Morris Avenue, Athens, OH 45701.

May 19-21: OKLAHOMA: Green Country Hamfest, Expo Square Pavilion, 17th and Louisville, Tulsa. Registration Green Country Hamfest, POB 4283, Tulsa, OK 74135. For information (918) 272-3081.

May 19-21: NEW HAMPSHIRE: The 15th annual Eastern VHF/UHF/SHF Conference, sponsored by the Northeast VHF Association, Rivier College, Nashua. Registration chairman David Knight, KA1DT, 15 Oakdale Ave, Nashua, NH 03062.

May 20: MICHIGAN: Swap and Shop sponsored by the Waxauke ARA, Cadillac Middle School, 500 Chestnut Street, Cadillac. 8:30AM to 2:30PM. Contact John Craddock, KX8Z (616) 797-5491 or Waxauke ARA, PO Box 163, Cadillac, MI 49601.

May 20: ARKANSAS: Ozark Hamboree sponsored by the Northwest Arkansas ARC, Rodeo Community Center, Springdale. 8AM to 3PM. For information Randall Spear, WA5QGH, (501) 846-3210.

May 20: ILLINOIS: 3rd annual Hamfest and Electronic Flea Market sponsored by the Lewis & Clark Radio Club, Lewis & Clark Community College, Godfrey. 8AM to 3PM. For information/tickets: Lewis & Clark Radio Club, PO Box 553, Godfrey, IL 62035. (618) 466-1909.

May 20-21: WASHINGTON: Hamfest '89 sponsored by the Yakima ARC, Central Washington State Fairgrounds, Yakima. Contact Dick Umberger, N7HHU (509) 453-8632 days or 453-3580 evenings. Early bird special Yakima ARC, W7AQ, PO Box 9211 Yakima, WA 98909.

May 20: TEXAS: 4th Annual Armed Forces Day Hamfest, sponsored by the Key City ARC, Abilene Civic Center, Pine St, Abilene. 8AM to 5PM. Wheelchair Accessible. For information Bill Jones, N5DOX (915) 698-4606 or KCARC, PO Box 2:722, Abilene, TX 79604.

May 20: MINNESOTA: Swapfest '89 sponsored by the Arrowhead Radio Amateur Club, First United Methodist Church, 230 East Skyline Parkway, Duluth. 10AM to 3PM. For information/registration John Crow, KA0SYN, 1365 Roland Road, Cloquet, MN 55720. (218) 879-5356.

May 20: GEORGIA: 10th annual Lake Hartwell Hamfest sponsored by the Anderson, Hartwell and Toccoa ARCS, Lake Hartwell Group Camp, Hwy 29, 4 miles north of Hartwell. For information George C. Haddock, KB4HCB, Rt 1, Box 52, Martin, GA 30557 or Carl Davis, College Avenue, Hartwell, GA 30603.

May 20: PENNSYLVANIA: Lancaster County Hamfest, sponsored by the Ephrata Area Repeater Society, Ephrata Senior High School, 803 Oak Blvd, Ephrata. Starting 8AM. For information/reservations Tom Youngberg, K3RZF (215) 267-2514 or EARS, 906 Clearview Ave, Ephrata, PA 17522.

May 20: RHODE ISLAND: Spring Flea Market and Auction sponsored by the RI Amateur FM Repeater Service, VFW Post 6342, Main Street, Forestdale (No. Smithfield). Noon to 5 PM. For information contact Rick Fairweather, K1KYI, Box 591, Harrisville, RI 02830. (401) 568-0566 from 7-9 PM.

May 20: COLORADO: 1989 Swapfest sponsored by the Pikes Peak Radio Amateur Association, Rustic Hills Mall, Palmer Park and academy Blvd, Colorado Springs. For information/reservations Al, N0CWM (719) 473-1660 or write PPRAA Swapfest, PO Box 16521, Colorado Springs, CO 80935.

May 21: PENNSYLVANIA: 15th annual Hamfest sponsored by the Warminster ARC, Middletown Grace Fairgrounds, Penns Park Road, Wrightstown. Gates open 8 AM. For information/registration Bill Cusick, W3GJC, Apt 804, Garner House, Hatboro, PA 19040. (215) 441-8048.

May 21: ILLINOIS: Knox County Hamfest, sponsored by the Knox County Radio Club, Knox County Fairgrounds, Knoxville. Starts 8AM. For tickets/information Keith L. Watson, WB9KHL, 119 South Cherry Street #3, Galesburg, IL 61401-4527. (309) 342-3885 evenings.

May 21: WEST VIRGINIA: The 11th annual TSRAC Wheeling Hamfest/Computer Fair, Wheeling Park. 8 AM to 3 PM. To reserve space contact Sandi Williams, WC8P, 9 East High Street, Flushing, OH 43977 (614) 968-3652. For tickets TSRAC, Box 240, RD 1, Adena, OH 43901 (614) 5546-3930.

May 21: ILLINOIS: Hamfest sponsored by the Kankakee Area Radio Society, Will County Fairgrounds, Peotone. 8-3. For information write KARS c/o Frank DalCanton, KA9PWW, RR 1, Box 361, Chebanse, IL 60922. (815) 932-6703 after 4 PM or (815) 937-2452 before 4 PM CST.

May 21: ILLINOIS: Mini-Hamfest sponsored by the Chicago ARC, North Park Village, 5801 N. Pulaski, Chicago. 9AM to 3PM. For information contact CARC, 5631 W. Irving Park Road, Chicago, IL 60634. (312) 545-3622.

May 21: CALIFORNIA: HAMSAP, sponsored by the North Hills Radio Club, Folsom Community Clubhouse, Folsom. 8 AM to 3 PM. Contact NHRC, PO Box 41635, Sacramento, CA 95841 or call Bob, WA6ULL (916) 983-2776.

May 27: NORTH CAROLINA: 10th annual Durhamfest 1989, sponsored by the Durham FM Association, lower rear deck South Square Mall, Durham, rain or shine. 8AM to 4PM. For information Mick Rankin, W4ZUS, 1001 Wedgewood Lane, Durham, NC 27712.

May 28: MARYLAND: Memorial Day Hamfest sponsored by the Maryland FM Association, Howard County Fairgrounds, Rt 144, West Friendship. 8AM to 3PM. For information/reservations Mike Cresap, 1294 Dorothy Road, Crownsville, MD 21032. (301) 923-3829.

June 3: NEW HAMPSHIRE: The Hoesstraders Flea Market is back at the Deerfield Fairgrounds. Admission \$5 per person. Wheelchair accessible. Questions or map SASE to WA1IVB, RFD Box 57, West Baldwin, ME 04091.

June 4: MICHIGAN: Swap 'N Shop sponsored by the Chelsea ARC. For information Robert Schantz, 416 Wilkinson Street, Chelsea, MI 48118. (313) 475-1795.

June 4: NEW YORK: Lancaster Hamfest sponsored by the Lancaster ARC, Depew Grove, 271 Columbia at French Rd, Depew. 8AM to 5PM. For information WA2CJJ (716) 681-6410 or KE2FM (716) 681-3512.

June 4: PENNSYLVANIA: 35th annual Hamfest sponsored by the Breeze Shooters, White Swan Amusement Park, Rt 60 (Parkway West) near greater Pittsburgh International Airport. For information John Colbert, K3SDL, 1851 Highland Ave, Irwin, PA 15642. (412) 863-5167.

June 4: NEW YORK: Hall of Science Hamfest, sponsored by the Hall of Science ARC, Hall of Science parking lot, Flushing Meadow Park, 47-01-111 Street, Queens. Starts 9 AM. For information Steve Greenbaum, WB2KDG (718) 898-5599 or Arnie Schiffman, WB2VXB (718) 343-0172.

June 10: MICHIGAN: 15th annual Hamfest sponsored by the Central Michigan Amateur Repeater Association (CMARA), Midland Community Center, Midland. 8AM-1PM. For information SASE to CMARA Hamfest, PO Box 67, Midland, MI 48640 June 10: MAINE: 3rd annual Outdoor Hamfest sponsored by the Pine State ARC, Hammond Street Campground, near 195, Bangor. Dawn to 5 PM. For information Ed Richardson, N91L, 825-4417; Howie Soule, K1CZ, 848-3397.

June 10: ONTARIO: Central Ontario Amateur Radio Flea Market, Bingham Park, Kitchener, Ont. Contact Ray Jennings, VE3CZE, 61 Ottawa Crescent, Guelph, Ont. N1E 2A8. (519) 822-8342.

OPERATING EVENTS

"Things to do . . ."

April 30: The Clairmont Repeater Assoc. will operate W6FZZ, SAMS DAY, to honor Samuel F.B. Morse. Samuel F.B. Morse III will be operating from this station. For QSL CLARA, Box 7675, Huntington Beach, CA 92615.

May 13-14: Nevada QSO Party sponsored by the Frontier ARA, Las Vegas. 000Z May 13 to 0600Z May 14. 6-160m, CW/SSB/FM/RTTY/Packets/SSTV. Mail logs to Jim Frye, NW70, 4120 Oakhill Ave, Las Vegas, NV 89121.

May 20: The Maryland Mobiles ARC will operate WA3PJO aboard the Submarine U.S.S. Torsk (SSK-423), 1400Z to 2100Z. For certificate send legal SASE to MMARC, POB 784, Severna Park, MD 21146.

May 20: 40th annual ARMED FORCES DAY. In recognition the ARS W4ODR located Northside aboard Naval Air Station Memphis, Millington, TN, will operate from 1300Z to 2300Z. For additional information W4ODR/Navys—Marine Corps MARS Station NNN0NIF, Bldg N-100, NAS Memphis. (901) 873-5134.

May 20: Special event station KM3I will be on the air to commemorate the 145th anniversary of the telegraph message "What Hath God Wrought?", transmitted on an experimental line from Washington, DC to Baltimore, MD. For a commemorative certificate, Amateurs send QSL card, SWL's send QSO details with large SASE to The Bay Area ARS, PO Box 805, Pasadena, MD 21122-0805.

May 20-21: The St. Charles ARC will operate WB0HSI from 1300Z to 2100Z to commemorate Lewis and Clark Rendezvous Days. For certificate send large SASE to St. Charles ARC, PO Box 1429, St. Charles, MO 63302-1429.

May 21-27: Special event station WA4ZIO will be operating from the Alabama Reunion Train. Sponsored by the Heart of Dixie Railroad Museum, Birmingham ARC and ARRL AL section. 80-10m phone and CW. For certificate send QSL and 9x12 SASE to Birmingham ARC, POB 603, Birmingham, AL 35201.

June 3: The Conemaugh Valley ARC will operate WA3WGN to commemorate the centennial of the flood of 1889 in Johnstown, PA. Lower General phone bands, 20, 40m. Novice phone 10m. For QSL send #10 SASE to Conemaugh Valley ARC, 194 Barron Ave, Johnstown, PA 15906.

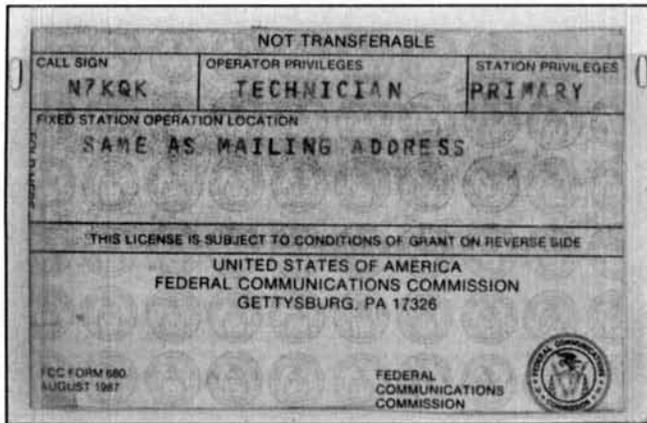
June 4: The Wireless Institute of Northern Ohio (WINO) an organization sponsored by the Lake County ARA will operate special event station KO6O from a winery in Madison, Ohio, to commemorate Ohio Wine Month. 1500Z to 1900Z 14235 and 21310 kHz. For QSL send legal SASE to KO6O, WINO Weekend, 10418 Briar Hill, Kirtland, OH 44094.

NORTH COAST ARC 1989 LICENSE EXAMS. 12:30 PM, Saturdays February 11, April 15, June 10, August 12, October 14, December 9. N. Olmsted Community Cabin, S. of Lorain on W. Park. Novice thru Extra. Walk-ins allowed. Talk in 145.2n repeater. For information Dan Sara, KB8A, 15591 Rademaker Blvd, Brookpark, Ohio 44142. 267-5083 or Pauline Wells, KA8FOE, Rick Wells, K8SCI, 777-9460/779-8999.

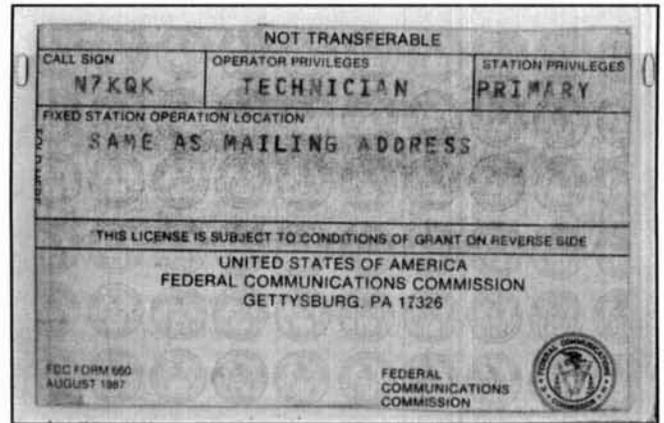
AMATEUR RADIO CLASSES: For those people interested in obtaining a Novice (basic level) Ham license or upgrading to Tech/General, the Chelsea Civil Defense, in cooperation with QRA Radio Club, will sponsor Amateur Radio Communications classes evenings at Chelsea High School starting MARCH 7, 1989. For more information write Frank Masucci, K1BPN, 136 Grove Street, Chelsea, MA 02150. Please enclose your telephone number.

THE MIT UHF REPEATER ASSOCIATION and the MIT Radio Society offer monthly Ham EXAMS. All classes Novice to Extra. Wednesday, MAY 24, 7 PM, MIT Room 1-150, 77 Mass Avenue, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann at (617) 484-2098. Exam fee \$4.50. Bring a copy of your current license (if any), two forms of picture ID, and a completed form 610 available from the FCC in Quincy, MA (617) 770-4023.

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Model R-T and LT EMP Series Arc-Plug® cartridges are designed to protect against nuclear electromagnetic pulse (EMP), as well as lightning surge voltages.

The EMP Series design is based on the National Communications System Technical Information Bulletin 85-10 covering EMP protection for communication equipment.

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See Data Sheet for surge limitations.

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by Jim Rafferty N6RJ

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This disk contains four different contest programs; ARRL Sweepstakes, Field Day, Universal WW Contest log, plus a dupe checking routine. Automatically enters date, time, band and serial number for each contact. When the contest is over, the program will print your results listing all duped and scored contacts in serial sequence with all the necessary information as well as completed score at the bottom of the page.

HD-CL (For C-64) \$24.95

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Particularly helpful in determining long path and grey line openings. Super fast speed and dazzling graphics make this program a treat to use. The MS-DOS version also includes a close up (zoom) feature for detailed examination, a MUF calculator and a great circle bearing routine. All call sign prefixes and country names are built into the data base for easy pinpointing of locations. MS-DOS version also color compatible. Requires 2 disk drives, 348K of memory, Hercules, CGA or EGA graphics and DOS 2.1 or later.

XN-DOS (IBM or compatible computers) \$34.95

XN-C64 (C-64 computer) \$34.95

XN-DX (slide rule version) \$22.95

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NEW PRODUCTS

Electronic temperature-control soldering station

The Elenco electronic temperature-control soldering station has a circuit which lets you change tip temperature from 300°F (150°C) to 900°F (480°C) without changing the tip or heating element. A temperature sensor located near the tip offers rapid response and little temperature variation. The tip of the unit is isolated from the AC line by a transformer. Low voltage (24 volts) powers the heating element. Completely electronic switching protects voltage and current-sensitive components. This unit has a linear LED array readout which accurately indicates tip temperature. It is priced at \$169.

Contact Elenco Electronics, Inc., 150 W. Carpenter Avenue, Wheeling, Illinois 60090 for details.

Circle #309 on Reader Service Card.

Transverter from R N Electronics

R N Electronics of Essex, England announces the new 2 to 6-meter transverter. It can be used with your existing 2-meter transceiver and has 25 watts PEP output. For more information write R N Electronics, 37 Long Ridings Avenue, Hutton, Brentwood, Essex CM131EE, UK.

ICOM's new IC-765 HF transceiver

ICOM announces the new ICOM IC-765 HF transceiver which features:

- Direct Digital Synthesizer (DDS).
- Band stacking registers.
- 99 fully tunable memories.
- CW pitch control.
- Maximum operating flexibility.
- Built-in AC supply.
- Automatic antenna tuner with built-in CPU and memory.
- 10-Hz readout.

The IC-765 is priced at \$3,149, and comes with narrow 500-Hz CW filters. The 250-Hz FL-53A and FL-101 are optional filters.



For details contact ICOM America at 2380 116th Avenue N.E., PO Box C-90029, Bellevue, Washington 98009-9029.

Circle #310 on Reader Service Card.

New manuals from Kantronics

Kantronics, Inc. announces its new manual set. The three manuals included are: *Installation Manual*, *Operation Manual*, and *Command Manual*. This set includes instructions for KAM, KPC-2, KPC-2400, and KPC-4.

For more information contact Kantronics, Inc., 1202 E. 23rd Street, Lawrence, Kansas 66046.

Circle #311 on Reader Service Card.

Jensen Tool catalog

A new catalog is offered free by Jensen Tools. Illustrated in full color, the 160-page catalog describes Jensen's full line of over 40 specialty tool kits for field service, plus a new line of products of fiber optics and wire/cable systems. Also included are hand and power tools in English and metric sizes, test equipment, soldering/desoldering stations, static control, lighting/optical aids, carrying cases, shipping containers, and more.

For a copy write Jensen Tools, Inc., 7815 S. 46th Street, Phoenix AZ 85044, or call (602) 968-6231.

Circle #312 on Reader Service Card.

VOICE-ID™ digital voice annunciator

VOICE-ID™ can store and reproduce voice messages and/or CW in any logical combination with various delays. It's an add-on device for repeaters and Amateur Radio stations, and of interest to the DX contester.

High-quality, non-robotic voice reproduction is achieved through voice compression algorithms encoded in a non-volatile EPROM. Voice messages are stored in the EPROM, so no re-recording is necessary after a power failure.

The VOICE-ID™ is field installable, and is suitable for use in remote applications. It may also be battery operated in case of emergency.

For more information contact Time Domain Systems, 5003 Cowell Boulevard, Davis, California 95616.

Circle #313 on Reader Service Card.

Isolator line expanded

Electronic Specialists expands their patented isolator line to include remote power switching, power fail interrupt, and 20-A options. Suppressor performance of all units has been expanded to 39,000 surge amperes for added equipment protection. Isolators, with wideband high attenuation channel filters, are available in commercial, industrial and laboratory grades. Expanded isolator performance and options are available. Prices start at \$100.



For details contact Electronic Specialists, Inc., 171 South Main Street, Natick, Massachusetts 01760.

Circle #314 on Reader Service Card.

M-5000 autoranging multimeter

The Elenco M-5000 is a handheld 3-1/2 digit autoranging multimeter. VOM functions, Hi-Low ohms, diode check, 10-A AC/DC current ranges, and audible continuity check are standard. Other features include: data hold, memory, and manual or autoranging.

The M-5000 comes complete with operator's manual, test leads, and battery. It weighs under 7 ounces and is priced at \$69.95.

For details contact Elenco Electronics, Inc., 150 West Carpenter Avenue, Wheeling, Illinois 60090.

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TS-140S AFFORDABLE DX-ing!

- HF Transceiver With General Coverage Receiver
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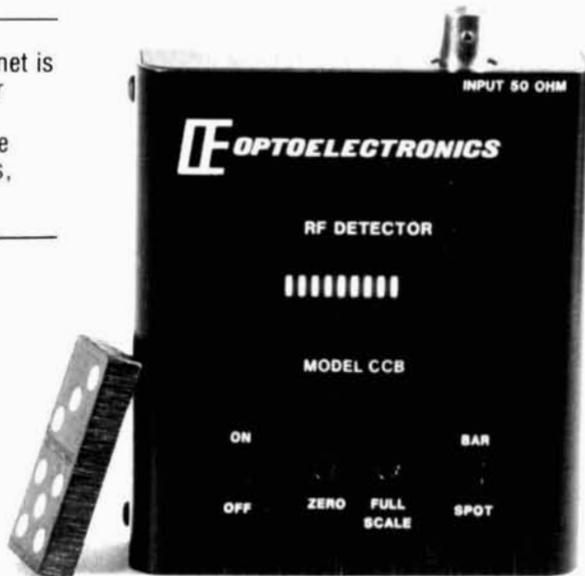
MODEL	2210 <i>New</i>	1300H/A	2400H	CCA	CCB <i>New</i>
RANGE: FROM	10 Hz	1 MHz	10 MHz	10 MHz	10 MHz
TO	2.2 GHz	1.3 GHz	2.4 GHz	550 MHz	1.8 GHz
APPLICATIONS	GENERAL PURPOSE AUDIO-MICROWAVE	RF	MICROWAVE	SECURITY	SECURITY
PRICE	\$199	\$169	\$249	\$299	\$99
SENSITIVITY					
1 KHz	< 5 mv	NA	NA	NA	NA
100 MHz	< 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

ACCURACY ALL HAVE +/- 1 PPM TCXO TIME BASE.

All counters have 8 digit red .28" LED displays. Aluminum cabinet is 3.9" H x 3.5" W x 1". Internal Ni-Cad batteries provide 2-5 hour portable operation with continuous operation from AC line charger/power supply supplied. Model CCB uses a 9 volt alkaline battery. One year parts and labor guarantee. A full line of probes, antennas, and accessories is available.



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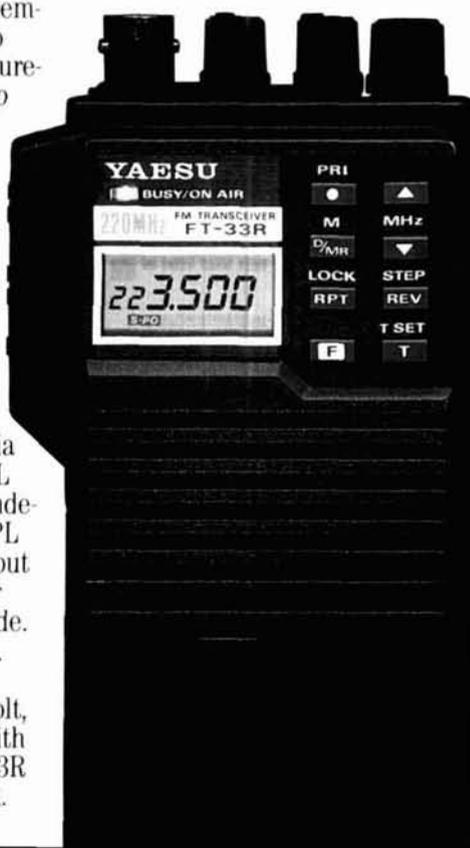
To begin with, you'll find a model that's right on your wavelength. The 2-meter FT-23R. The 220-MHz FT-33R. Or the 440-MHz FT-73R.

Whichever you choose, you benefit from incredibly small packaging. (Take a look at the actual size photo.) Aluminum-alloy cases that prove themselves reliable in a one-meter drop test onto solid concrete. And moisture-resistant seals that really help keep the rain out.

But perhaps best of all, each radio blends sophisticated, micro-processor-controlled performance with surprisingly simple operation. In fact, it takes only minutes to master all these features:

Ten memories that store frequency, offset and PL tone. Memory scan at 2 frequencies per second. Tx offset storage. Priority channel scan. Channel selection via tuning knob or up/down buttons. PL tone board (optional). PL display. Independent PL memory per channel. PL encode and decode. LCD power output and "S" meter display. Battery-saver circuit. Push-button squelch override. Eight-key control pad. Keypad lock. High/low power switch.

The FT-23R comes with a 7.2-volt, 2.5-watt battery pack. The FT-73R with a 7.2-volt, 2-watt pack. And the FT-33R with a powerful 12-volt, 5-watt pack.



You can choose the miniature 7.2-volt, 2-watt pack shown in the photo below. And all battery packs are interchangeable, too.

And consider these options: Dry cell battery case for 6 AAA-size cells. Dry cell battery case for 6 AA-size cells. DC car adapter/charger. Programmable CTCSS (PL tone) encoder/decoder. DTMF keypad encoder. Mobile hanger bracket. External speaker/microphone. And more.

Check out the FT-23R Series at your Yaesu dealer today. Because although we can tell you about their incredible performance, toughness and small size, seeing is really believing.



YAESU

Yaesu USA 17216 Edwards Road, Cerritos, CA 90701 (213) 404-2700. Repair Service: (213) 404-4884. Parts: (213) 404-4847

Prices and specifications subject to change without notice. PL is a registered trademark of Motorola, Inc. FT-33R shown with optional FNB-9 battery pack.

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TH-75A

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The new TH-75A Dual Band HT from Kenwood is here now! Many of the award-winning features in our dual band mobile transceivers are designed into one hand-held package.

- Dual Watch function allows you to monitor both bands at the same time.
- One watt on 2 meters and 70cm: **5 watts when operated on 12 VDC (or PB-8 battery pack).**
- Large dual multi-function LCD display.
- 10 memory channels for each band stores frequency, CTCSS, repeater offset, frequency step information, and reverse. A lithium battery backs up memories. Two memories for "odd split" operation.
- Selectable full duplex operation.
- Extended receiver range: 141-163.995 and 438-449.995 MHz; transmit on Amateur band only. (Modifiable for MARS and CAP. Permits required. Specifications guaranteed on Amateur bands only.)
- Uses the same accessories as the TH-25AT (except soft cases).
- Volume and balance controls, plus separate squelch controls on top panel.
- Super easy-to-use! For example, to recall memory channel, just push the channel number!
- CTCSS encode/decode built-in!
- Automatic Band Change (ABC). Automatically switches between main and sub band when signal is present.
- Automatic offset selection on 2 meters.
- Tone alert system for quiet monitoring. When CTCSS decode is on, the tone alert will function only when a signal with the proper tone is received.
- Four ways to scan, including **dual memory scan**, with time operated or carrier operated scan stop modes, and priority alert.
- Automatic battery saver circuit extends battery life.



- Supplied accessories: Dual band rubber-flex antenna, PB-6 battery pack, wall charger, belt hook, wrist strap, water resistant dust caps.

Optional Accessories

- PB-5 7.2 V, 200 mAh NiCd pack for 1.0 W output
- PB-6 7.2 V, 600 mAh NiCd pack
- PB-7 7.2 V, 1100 mAh NiCd pack
- PB-8 12 V, 600 mAh NiCd for 5 W output
- PB-9 7.2 V, 600 mAh NiCd with built-in charger
- BC-10 Compact charger
- BC-11 Rapid charger

- BT-6 6-cell AA battery case
- DC-1/PG-2V DC adapter
- HMC-2 Headset with VOX and PTT
- SC-22 and SC-23 Soft case
- SMC-30/31 Speaker mics.
- WR-1 Water resistant bag.

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