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ham radio

magazine

MAY 1978

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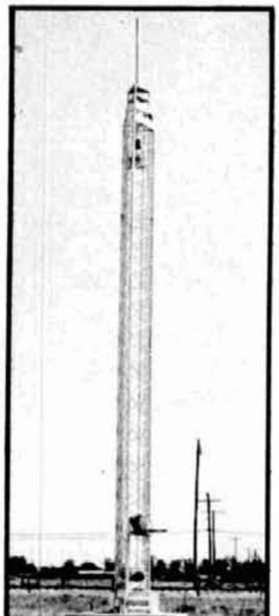
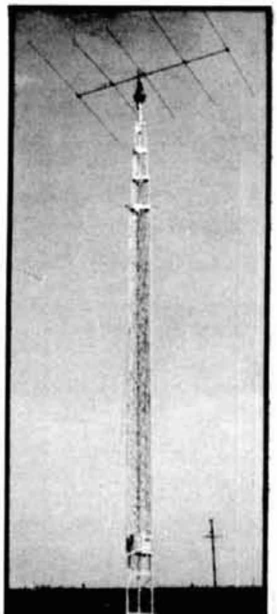
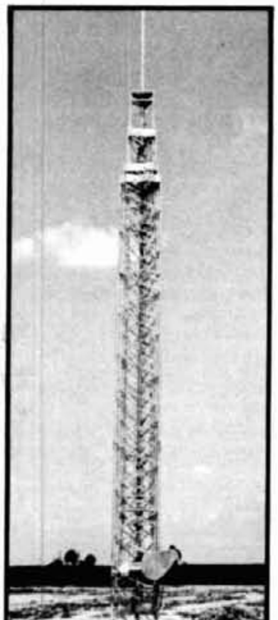
for HAM operators at basic prices!



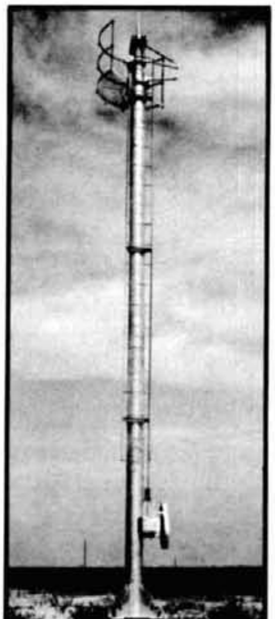
W-51



MW-65



LM-470D



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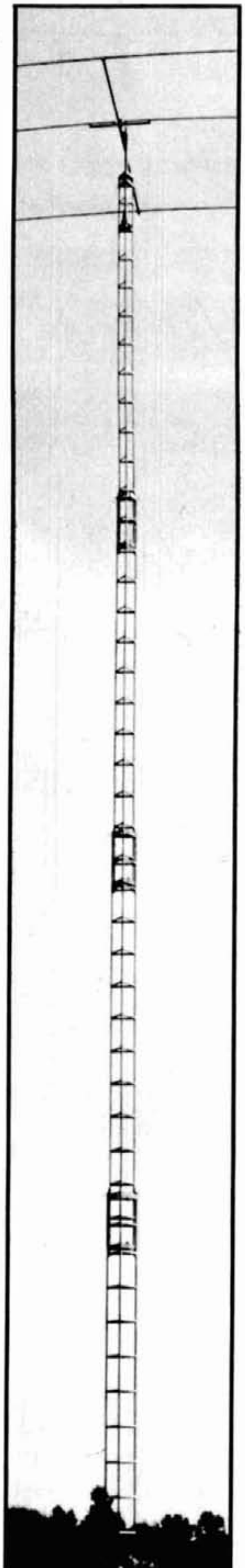

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An aerodynamic tower designed to hold 9 square feet in a 50 mph wind. Six models at different heights.

MW Series

Self-supporting when attached at first section — will hold normal Tri-Band beam. Six models.

LM Series

A 'W' brace motorized tower. Holds large antenna loads up to 70 feet high. Super buy.

TM Series

Features tubular construction for really big antenna loads. Up to 100 feet. Free-standing, with motors to raise and lower.

THD Series

Very popular. Low Cost. Holds Tri-Band antennas. Eight models — all support 7 square feet of antenna at full height in 70 mph winds. Guyed.

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Tri-Ex®

TOWER CORPORATION

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More Details? CHECK — OFF Page 150

This NEW MFJ Versa Tuner II . . .

has SWR and dual range wattmeter, antenna switch, efficient airwound inductor, built in balún. Up to 300 watts RF output. Matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines.

BRAND NEW

$79.95

Antenna matching capacitor. 208 pf. 1000 volt spacing.

Sets power range, 300 and 30 watts. Pull for SWR.

Meter reads SWR and RF watts in 2 ranges.

Efficient airwound inductor gives more watts out and less losses.

Transmitter matching capacitor. 208 pf. 1000 volt spacing.

Only MFJ gives you this MFJ-941 Versa Tuner II with all these features at this price:

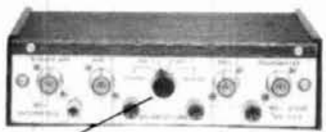
A SWR and dual range wattmeter (300 and 30 watts full scale) lets you measure RF power output for simplified tuning.

An antenna switch lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

A new efficient airwound inductor (12 positions) gives you less losses than a tapped toroid for more watts out.

A 1:4 balun for balance lines. 1000 volt capacitor spacing. Mounting brackets for mobile installations (not shown).

With the NEW MFJ Versa Tuner II you can run your full transceiver power output — up to 300 watts RF power output — and match your transmitter to **any** feedline from 160 thru 10 Meters whether you have coax cable, balance line, or random wire.



ANTENNA SWITCH lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

You can tune out the SWR on your dipole, inverted vee, random wire, vertical, mobile whip, beam, quad, or whatever you have.

You can even operate all bands with just one existing antenna. No need to put up separate antennas for each band.

Increase the usable bandwidth of your mobile whip by tuning out the SWR **from inside your car.** Works great with all solid state rigs (like the Atlas) and with all tube type rigs.

It travels well, too. Its ultra compact size 5x2x6 inches fits easily in a small corner of your suitcase.

This beautiful little tuner is housed in a deluxe eggshell white Ten-Tec enclosure with walnut grain sides.

SO-239 coax connectors are provided for transmitter input and coax fed antennas. Quality five way binding posts are used for the balance line inputs (2), random wire input (1), and ground (1).

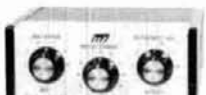
$59.95

BRAND NEW

MFJ-901 VERSA TUNER

New efficient air wound coil for more watts out.

Only MFJ uses an efficient air wound inductor (12 positions) in this class of tuners to give you more watts out and less losses than a tapped toroid. Matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines. Up to 200 watts RF output. 1:4 balun for balance lines. Tune out the SWR of your mobile whip **from inside your car.** Works with all rigs. Ultra compact 5x2x6 inches. SO-239 connectors. 5 way binding posts. Ten Tec enclosure.



$49.95

BRAND NEW

MFJ-900 ECONO TUNER

Same as MFJ-901 Versa Tuner, but does not have built-in balun for balance lines. Tunes coax lines and random lines.

$39.95

MFJ-16010 RANDOM WIRE TUNER

Operate 160 thru 10 Meters. Up to 200 watts RF output. Matches high and low impedances. 12 position inductor. SO-239 connectors. 2x3x4 inches. Matches 25 to 200 ohms at 1.8 MHz.

$49.95

BRAND NEW

MFJ-202 RF NOISE BRIDGE

This MFJ RF Noise Bridge lets you adjust your antenna quickly for maximum performance. Measure resonant frequency, radiation resistance and reactance. Exclusive range extender and expanded capacitance range (± 150 pf) gives you much extended measuring range.

Tells resonant frequency and whether to shorten or lengthen your antenna for minimum SWR. Adjust your single or multiband dipole, inverted vee, beam, vertical, mobile whip or random system for maximum performance. 1 to 100 MHz. SO-239 connectors. 2x3x4 inches. 9 volt battery.

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More Details? CHECK—OFF Page 150

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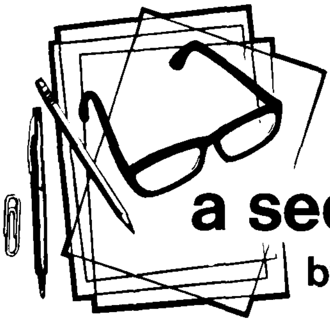

a second look

by Jim Fisk

Let's say you have an Extra or Advanced-class license and are bored with working DX, handling traffic, or trying out new SSTV circuits. What next? Why not shift gears and put your station and expertise to work helping out the Novice operators?

The other day I was tuning across the 40-meter Novice band and heard a cool, crisp, CW signal sending CQ. The band was full of shortwave-broadcast signals, and the CQ signal wasn't very strong. But it sounded like a tape recording — slow, steady, and with good spacing between characters. I listened. Nothing much but foreign-broadcast QRM and a few weak CW stations working each other. Then, down in the noise, I heard a fellow calling and answering the CQ. He was a WB9, sending very slowly with many errors, pauses, and lots of stops and starts. The station calling CQ answered the WB9, slowly and patiently.

What ensured was one of the more beautiful things that happen in amateur radio. The station calling CQ was an old timer, who decided to put his rig on the Novice 40-meter band. The Novice who answered his call, obviously brand new and unsure of himself, was glad to hear *any* station. I listened to the QSO. What happened renewed my faith in ham radio. The old timer was obviously ready to use his high-speed keyer — you could tell at the end of his transmission. But the old timer kept his cool and pounded away on his straight key, slow and steady.

It all ended with the usual amenities. The Novice wanted another QSO, but above all, he wanted a QSL card to confirm the QSO. No problem. Addresses and names were exchanged. Hopefully, both exchanged QSL cards and both followed up to build a lasting friendship.

If you're an experienced ham operator and are interested in expanding your horizons, why not consider putting your rig on the Novice bands? The 40-meter Novice band is a good place to start. Many new hams obtain gear that is band limited, and most Novices start out on the low-frequency ham bands — 40 and 80 meters. Equipment is easy to get working on these bands, as most old timers will realize. Whatever Novice band you choose, bear in mind the following facts.

Most Novices have dipole antennas and rather unsophisticated transmit-receive facilities. Many use knife switches to transfer between receive and transmit. So if you decide to put your rig on the Novice bands, with all the new features, remember that the Novice is unaware of most modern developments. He's interested in receiving your transmission, simply and without flair.

Patience is the watchword when working Novices. Patience requires a certain discipline that pays off in genuine self satisfaction when you've completed a good QSO.

If you're an overseas amateur, your signal is more than welcome in the U.S. Novice bands. You'll find Novices that can handle Morse pretty well — they're almost ready for the next step — the exam for the General-class license. But don't overlook the vast majority of U.S. Novices. All would like a QSO with a DX station. Slow down from time-to-time and give the new fellows a chance.

Some additional tips if you're an old timer and want to work Novices: If you call a Novice station and a reply is not immediately forthcoming, listen for at least two minutes. Perhaps it's his first QSO. The new amateur who puts his rig on the air for the first time is usually nervous, anxious, and maybe somewhat confused. If you can remember your first QSO, I'm sure you know the feeling (the first time I attempted to answer another station I forgot to turn on the B + supply to the transmitter, but that's another story). So if the Novice you call doesn't come back right away, tune around and wait. Chances are he'll answer after a few minutes. If not, call him again. He's probably overwhelmed that someone answered his call.

It's a great feeling to be able to help a newcomer. All you need is a little patience and perseverance. If you do it right, you'll be surprised. You'll have a pile-up of Novices trying to work you — it's almost like being a rare DX. Try it, you'll like it . . . but be patient.

Jim Fisk, W1HR
editor-in-chief

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ICOM's superior LSI technology introduces the most advanced transceiver in the HF world, the **IC-701!** Now ICOM's famous One/Two Team of **single knob** frequency selection and **dual VFO's** leaps to the forefront of HF with an extremely compact, all solid state, fully synthesized, 100 W CONTINUOUS OUTPUT Maximizer of all modes and all bands, from **160-10M.** The **IC-701** is the ICOM breakthrough you've been waiting for: the future in HF.

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Sold complete, with base mic and AC power/speaker, as shown.

IC-701 meets FCC Part 97.73.

All ICOM radios significantly exceed FCC specifications limiting spurious emissons.

Specifications: □ Frequency Coverage: 1.8 MHz — 2.0 MHz; 3.5 MHz — 4.0 MHz; 7.0 MHz — 7.5 MHz; 14.0 MHz — 15.2 MHz; 21.0 MHz — 21.5 MHz; 28.0 MHz — 30.0 MHz □ Frequency Control: LSI based 100 Hz step Digital PLL synthesizer. Independent Transmit-Receive duplex on same band, standard with every radio. □ Frequency Readout: 6 digit LED 100 Hz readout □ Power Supply Requirements: DC 13.6 V ± 15% Negative ground current drain, 18 A max at 100 W output; AC power supply, speaker console for AC operation □ Antenna Impedance: 50 ohms unbalanced, VSWR 2.0:1 □ Weight: 7.3 Kg □ Size: (transceiver unit only) 111mm (h) × 241mm (w) × 311mm (d) □ RF Power Output: CW (A1), RTTY (F1), 100 W; SSB (A3J), 100 W PEP; Continuously adjustable 0-100W □ Emission Modes: A1, CW; A3J, SSB; F1, RTTY □ Harmonic and Spurious Output: more than 60 dB below peak power (meets FCC 97.73) □ Carrier Suppression: more than 40 dB down □ Unwanted Sideband: more than 40 dB down at 1000 Hz AF input □ Microphone Impedance: 600 ohms □ Receiving System: triple conversion, super heterdyne, with continuous bandwidth control (100 Hz — 2.4 KHz) □ Receiving Modes: A1, A3J (USB/LSB), F1 □ IF Frequencies: 1st & 3rd, 9.0115 MHz; 2nd, 10.7015 MHz; with continuous bandwidth control □ Sensitivity: better than 0.25 microvolts for 10 dB S+N/N □ Selectivity: SSB, RTTY, ± 1.1 KHz at −6 dB (adjustable to ± 0.5 KHz min), ± 2.0 KHz at −60 dB; CW, ± 250 Hz at −6 dB ± 700 Hz at −60 dB; CN-N, ± 100 Hz at −6 dB, ± 500 Hz at −60 dB (with Audio Filter) □ Spurious Response Rejection Ratio: better than 60 dB

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OSCAR 8 IS UP AFTER A flawless launch that lifted off the launch pad right on schedule. Listeners to the AMSAT 20-meter net heard the actual countdown and blastoff relayed by WA3NAN. The later ejection of OSCAR 8 from the launch vehicle was so perfect that the 435.095-MHz beacon signal showed the satellite was tumbling far less than had been expected. This permitted extension of the 10-meter antenna on the fifth orbit, instead of waiting several days as had originally been planned, and on the following orbit Mode A was turned on as OSCAR 8 came over the southern horizon and its 29.402 MHz beacon laid a strong signal across the entire U.S.

Preliminary Tests Show the new bird to be a fine performer, with good Mode J signals and Mode A performance apparently even better than OSCAR 6 was.

OSCAR 8's Orbit came very close to predictions, with refined calculations showing it nearly circular with a 909.583 km mean altitude, a period of 103.23162 minutes, and an inclination of 98.99° resulting in a 25.80867° progression at the equator.

Orbit Information booklets for OSCAR 8 will be available the end of this month from Skip Reymann, W6PAJ, Box 374, San Dimas, California 91773. Costing $5 ($3 for AMSAT members or free for AMSAT Life members), the booklet will have orbit information through the end of this year.

OSCAR 8 Is Available for general use, operating Mode A weekdays (Wednesday, experimental only) and Mode J Saturday and Sunday (GMT). Both transponders are working so well (20W ERP produces S9 and 1W S6 on Mode A, and Mode J appears to be about 6 dB better!) that linears are definitely not needed.

THE AMATEUR CALLSIGN structure has been completely overhauled, effective March 24. No special callsigns or special events callsigns will be issued, and licenses for personal secondary stations will no longer be renewed. Call areas will remain, and will determine the prefix of a callsign issued to an address in that area, but an individual can retain his callsign indefinitely if he wishes, even if his permanent station location changes call area.

Under The New Policy, which divides callsigns into four "groups," 2x1 and Nx3 callsigns will soon be heard. Group A includes 1x2, 2x1, and 2x2 (AA-AL prefixes), available to Extras only. Group B includes 2x2s, other than AA-AL, and will be available to Advanced Class license holders. Group C is 1x3 callsigns, available to Generals and Technicians, while Group D is 2x3s for Novices. In Group A, 2x1s will be the first calls issued, while Nx3s will be the initial Group C issue.

The Initial Extra Class callsign (Group A), assigned in each call area will be AAxA, but 2x2s for Advanced Class (Group B) will begin with KA1AA in the first call area but KBxAA in all others (since KAx2 callsigns are currently in use by U.S. personnel in Japan). Novices are also going to start a new callsign group (Group D), with KAx3 such as KA3AAB.

NOTE: Only Extras and other class licensees who are upgrading or applying for their first Amateur license are currently eligible for the new calls. Requests are made by specifying on line 13A of Form 610, "Please assign a Group_ callsign" when filing the application. But, for any Extra who has received a 2x2, the FCC will process a request for change, if received before October 1, 1978, to a 2x1 callsign. Applications from others will be returned!

EFFECTIVE MAY 15, Novice licenses now have five year terms and may be renewed just like any other Amateur license as a result of the Commission's action on Docket 20282. At the same time they also extended Technician's privileges to all Amateur frequencies above 50 MHz, giving Techs all of 6 as well as 2 meters.

Also, Effective May 15, the new 2-meter repeater sub-band will be 144.5-145.5 MHz. In addition to deleting WR callsigns, repeaters will not be permitted use of 220.0-220.5, 431-433, or 435-438 MHz, to permit weak signal and satellite communications.

RICHARD COOPER HAS BEEN CHARGED with misrepresentation by the California State Attorney General's Office. In a suit filed in Los Angeles Superior Court, the state asked that Cooper and his Communications Attorney Service be ordered to stop disseminating false information about himself and his service.

Among Cooper's Claims Challenged in the suit were his statements that he and his partners were licensed to practice law, that he holds MS and PhD degrees, that Communications Attorney Service has associate attorneys in Atlanta, New York, and Chicago, and that his office has "one of the most extensive libraries of FCC reports in the country." Also branded as misleading or false in the suit were claims about the number of subscribers, size of his attorney staff, and the money that CBers could save by being a subscriber.

NEW PREFIXES FOR U.S. Pacific and Caribbean Islands are: KH1, Canton; KH2, Guam; KH3, Johnston; KH4, Midway; KH5K, Kingman; KH5, Palmyra; KH6, Hawaii; KH7, Kure; KH8, Samoa; and KH9, Wake. Also KP1, Navassa; KP2, Virgin Islands; KP3, Serrana Bank, and KP4, Puerto Rico. Other spots, such as the Marshall Islands and Guantanamo Bay, are not FCC administered and so remain unchanged.

KENWOOD

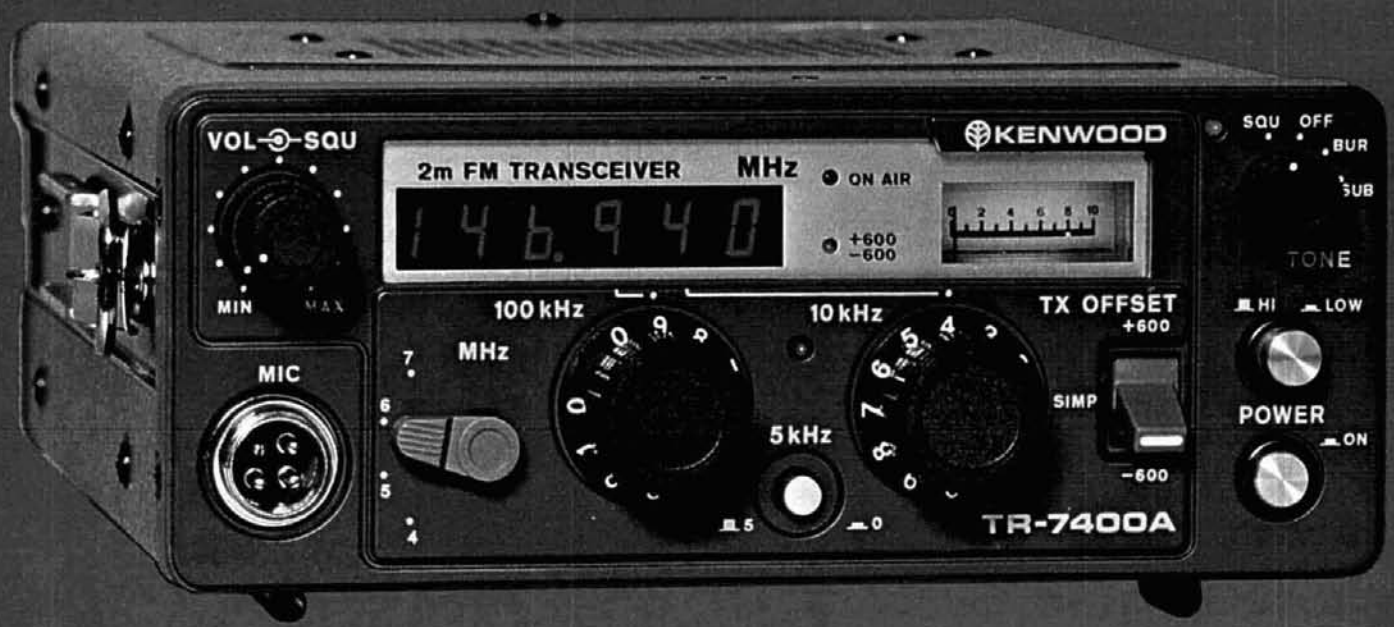
...pacesetter in amateur radio

Kenwood's exciting 2-meter transceiver . . . still the most powerful. 800 channels, repeater offset over all 4 MHz (144-148 MHz), dual frequency readout, easy to read 6 digit display, Kenwood's unique continuous tone coded squelch system and outstanding receiver performance. All in a rugged, compact package.

The TR-7400A lets you go anyplace on the 2-meter band . . . covers the entire band without compromise. It exceeds all FCC emission requirements for amateur transceivers. Its RF output is factory spec'd at 25 watts . . . but is typically over 30! It offers a dual frequency readout with large easy to read 6 digit LED display plus a functional dial readout system, fully synthesized 800 channel operation and repeater offset over all 4 MHz (144-148 MHz). The unique Continuous Tone Coded Squelch system is a Kenwood exclusive.

Outstanding sensitivity, large-sized helical resonators with High Q to minimize undesirable out-of-band interferance, and give a 2-pole 10.7 MHz monolithic crystal filter combine to give your TR-7400A outstanding receiver performance. Intermodulation characteristics (Better than 66dB), spurious (Better than −60dB), image rejection (Better than −70dB), and a versatile squelch system make the TR-7400A tops in its class.
(Active filters and Tone Burst Modules optional)

TR-7400A



The TR-7400A is shown with its furnished hand mike and the PS-8 DC power supply (optional). Take your TR-7400A out of the car and you can use it as a powerful base station. The PS-8 is rated at 8 Amps and is among the most rugged, well-regulated supplies available for VHF transceivers requiring 12V DC.

TR-7400A Specifications

Range: 144.00 MHz to 147.995 MHz
Mode: FM
800 Channels: 5 KHz spaced
Sensitivity: Better than 0.4 uV for 20 dB quieting
Better than 1 uV for 30 dB S/N
Squelch Sensitivity: Better than 0.25 uV
Selectivity: 12 KHz at −6 dB down
40 KHz at −70 dB down
Image Rejection: Better than −70 dB
Spurious Interference: Better than −60 dB
Intermodulation: Better than 66 dB
Receive System: Double conversion
First IF: 10.7 MHz
Second IF: 455 KHz
Audio Output: More than 1.5 Watts (8 ohm load)
RF Output Power: 25 Watts (High) 5-15 Watts (Low-adjustable)
Antenna Impedance: 50 ohms
Frequency Deviation: ±5 KHz
Spurious Response: Better than −60 dB
Microphone: Dynamic, with PTT switch, 500 ohms
Current Drain: Less than 1A in receive (no input signal)
Current Drain: Less than 8A in transmit

Windom antennas

Some facts and fiction about the Windom antenna, and how it measures up in terms of bandwidth and input impedance

One of the simplest and most economical antennas suitable for multiband amateur use is the single-wire, off-center-fed antenna, often called the Windom antenna.

Although this antenna in its present form was first described in the professional literature in 1929 by Everitt and Byrne,[1] and in the amateur literature by Loren Windom, W8GZ/W8ZG,[2] very little new or original theoretical or design data has been published since that time. To the best of my knowledge, no information on impedance or bandwidth measurements on the antenna has ever been published.

In this article I will begin by giving the history of the antenna, then a theoretical discussion of the antenna and single-wire transmission line, followed by a method of adjusting the antenna, and finally, the results of measurements I have made on an actual antenna.

history

The single-wire feedline has been credited to Frank Conrad, 8XK, of Westinghouse who used it in the broadcast band to feed a quarter-wavelength grounded (Marconi) antenna. The next step was taken by V. D. and E. B. Landon, 8VN, who connected the single feedwire to the junction of the antenna and counterpoise.[3] In a later article by Howard M. Williams, 9BXQ, the counterpoise was stretched out and made a part of the antenna with the feedpoint still off-center.[4] **Fig. 1** shows a summary of these early developments.

In the middle 1920s Loren Windom operated 8GZ/8ZG in the Columbus, Ohio, area. Windom ran high power for that time (250 watts) and was considered the technical bellwether of Columbus hams, as he was an active experimenter, working with antennas among other things.

In July, 1926, the technical editor of *QST*, Robert S. Kruse, published an article which gave a roundup of various methods of feeding an antenna as they were understood at that time.[5] A discussion by Windom on how to adjust the off-center-fed antenna was a part of that article; the procedure consisted of placing a light bulb in the center of the antenna (**fig. 2**) and adjusting the feedpoint for maximum lamp brilliance (maximum current). This procedure was also published in the first three editions of the ARRL *Radio Amateur's Handbook*.

John Byrne, then 8DKZ, became associated with Windom; 8GZ/8ZG QSL cards, circa 1925, carried the names of both Windom and Byrne. Both were students at Ohio State University — Byrne in electrical engineering and Windom in law.

It was customary in those days for senior engineering students to do a thesis for graduation; Byrne and his thesis partner, E. F. Brooke, 8DEM, chose the single-wire transmission line as their topic. They carried out a considerable amount of research on the

By John J. Nagle, K4KJ, 12330 Lawyers Road, Herndon, Virginia 22070

subject under the guidance of their faculty advisor, W. L. Everitt, who was then Associate Professor of Electrical Engineering. Everitt will be remembered for his book, *Communication Engineering*, which was the standard college textbook on the subject in the 1930s and early 1940s.

Windom, although not formally associated with the project, assisted from time to time in rigging or making measurements on his lunch hour or after classes, so he was up-to-date on the progress of the work. In fact, he incorporated many of the more promising facets at his own station.

Byrne's investigation was incomplete when he graduated in June, 1927; however, he returned as a graduate student for the 1927-28 academic year and continued work on the single-wire feeder as his Master's thesis with A. B. Crawford as his partner. The work thus accomplished was published in the *Proceedings of the IRE* in October, 1929, with Everitt and Byrne listed as authors.[1] It was standard practice then, as now, that when a student's research work is formally published, the faculty advisor is listed as senior author.

The work of Everitt, Byrne and Brooke showed that the Windom procedure of 1926 was incorrect. Windom agreed in his 1929 article, and stated that his earlier method should not be used; he went on to describe the Everitt-Byrne method that I will explain later. Beginning with the seventh edition of the *ARRL Handbook*, the old, Windom graphs for the single-wire feeder were presented which give the length of the antenna as a solid line and the location of the tap as a dashed line for each amateur band. The Everitt-Byrne procedure has not been published in any of the ARRL handbooks or antenna manuals and is not widely known.

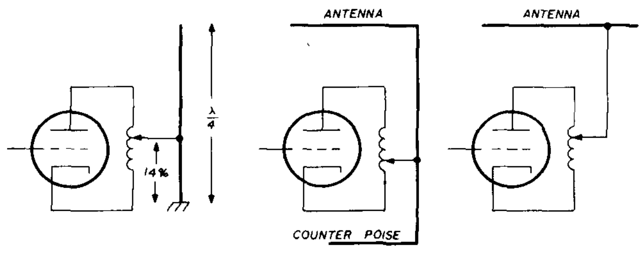


fig. 1. Early development of the off-center fed antenna beginning with the single-wire feedline used on the broadcast band by Frank Conrad, 8XK, to feed a quarter-wavelength vertical (left). Later the Landon brothers, 8VN, connected a single feedline to the junction of the antenna and counterpoise (center). Howard Williams, 9BXQ, stretched out the counterpoise and made it part of the antenna with the feedpoint still off-center (right).

Windom recognized the utility of the method to the amateur community and encouraged Byrne to write it up for *QST*. Byrne declined, suggesting that Windom write it up himself as Windom was familiar with the work. Windom's article was published in the September, 1929, issue of *QST*,[2] a month before the Everitt-Byrne article. The delay in the Everitt-Byrne paper is ascribed to the more extensive editorial review required by the *Proceedings of the IRE*; Windom, to his credit, stated in an early paragraph that he had not done any of the work himself — that he was only reporting the work of others. The fact that Windom's article was published a month before the Everitt-Byrne paper did cause considerable consternation, however.

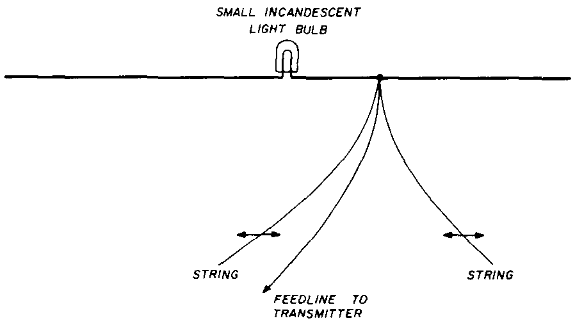


fig. 2. Windom's 1926 procedure for adjusting the feedpoint of an antenna as described in *QST*. The feedline was moved along the antenna until the bulb glowed the brightest. Later experiments at Ohio State University showed this method to be in error.

Byrne, having completed his studies, turned the project over to John Ryder, W8DQZ. Ryder was later to become Professor of Electrical Engineering at Michigan State University and the author of numerous electronics textbooks (he is not the Rider of *Rider's Manuals* fame). When Byrne left Ohio State University, he went to work at Bell Telephone Laboratories and gave up amateur radio. I cannot help but feel that amateur radio lost a valuable member by this decision.

The first use of the name "Windom" appears to be by the Wireless Institute of Australia which in 1930 published an article which was substantially a reprint of Windom's *QST* article. Shortly after World War II, the Radio Society of Great Britain ran an article, "Why Not a Windom?" It thus appears that the name Windom was imported into the United States from overseas. Rightly or wrongly, the name has stuck because it is much simpler than the more technical term, "off-center-fed Hertz antenna."

antenna theory

One of the most important characteristics of an antenna, when trying to couple energy into it, is its input impedance which, of course, is composed of both resistance and reactance. The resistance and reactance both depend on a large number of factors

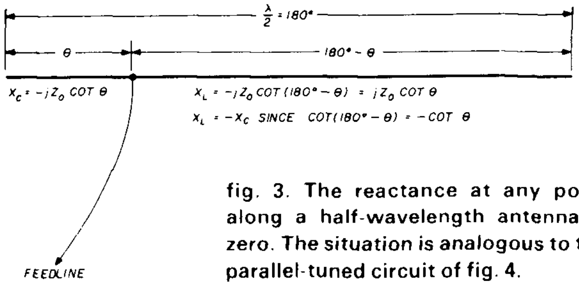


fig. 3. The reactance at any point along a half-wavelength antenna is zero. The situation is analogous to the parallel-tuned circuit of fig. 4.

such as height above ground, diameter of the antenna, nearby obstructions, and length of the antenna compared to a half-wavelength. While the input impedance can be calculated in the ideal case, the feedpoint impedance of an actual antenna must usually be measured, especially under amateur conditions.

Let's first talk about the antenna's reactance; to simplify things, we'll assume the antenna is exactly one-half-wavelength long or, as the old timers would say, "operating on its fundamental wave."

As is well known, the reactance looking into the center of a half-wavelength antenna is zero. Not as well known is the fact that the reactance at any other point on a half-wave antenna is also zero. Consider **fig. 3** where the antenna feedpoint is off-center. Looking into the short end, we have a transmission line with the far end open-circuited; the reactance looking into an open-circuited transmission line is given by

$$X_C = -jZ_o \cot \theta \qquad (1)$$

Where

$\theta =$ the electrical length of the transmission line in degrees

$Z_o =$ the characteristic impedance of the line.

If the antenna is exactly one-half-wavelength long, the length of the long end will be $180° - \theta$ and its reactance will be

$$X_L = -jZ_o \cot (180° - \theta) \qquad (2)$$

From trigonometry it can be shown that

$$\cot (180° - \theta) = -\cot \theta \qquad (3)$$

regardless of the value of θ. Combining **eqs. 1, 2,** and **3** gives

$$X_L = -X_C \qquad (4)$$

As the reactances are equal in magnitude but opposite in sign, the input reactance is zero. Since this is true regardless of the length of the line (value of θ), the reactance will be zero regardless of where the tap is located, provided only that the antenna is exactly a half-wavelength long. This situation is analogous to the parallel-tuned circuit shown in **fig. 4.** At frequencies below resonance, the inductive reactance will be less than the capacitive reactance so the current in the inductive branch will be greater than the current in the capacitive branch. Similarly, at frequencies above resonance, the current in the capacitive branch will be greater. At resonance the two currents will be equal. This fact provides an excellent means of determining the resonant frequency of any antenna that is an integral number of half-waves long, as discussed later.

We will now discuss the resistive component of impedance. For a half-wavelength dipole, the resistive component at the center is usually considered to be 72 ohms. This is true when the antenna is in free space or at certain heights above ground. The input resistance of the antenna at any location along its length can be easily determined in terms of its input resistance at the center. When the antenna is very thin and not terminated (as in a rhombic antenna), the current distribution along the antenna is essen-

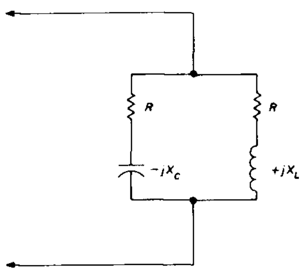


fig. 4. The parallel-tuned circuit analog of the off-center fed antenna.

tially cosinusoidal, as shown in **fig. 5**. The power applied to the center of the antenna is

$$P_o = I_o^2 R_o \qquad (5)$$

Where

P_o = power applied to center of antenna

I_o = rms current at center

R_o = radiation resistance at center

The power given by **eq. 5** must be equal to the power at any other point x along the antenna; therefore

$$I_o^2 R_o = I_x^2 R_x \qquad (6)$$

Since the current distribution is assumed to be cosinusoidal, $I_x = I_o \cos \theta$ where θ is the distance along the antenna in electrical degrees from the center. Substituting **eq. 6**

$$I_o^2 R_o = (I_o \cos\theta)^2 R_x$$

so that

$$R_x = \frac{R_o}{(\cos\theta)^2} \qquad (7)$$

From **eq. 7** we see that *theoretically* the input resistance of a half-wavelength antenna goes from a nominal 72 ohms at the center to infinite ohms at the ends. In practice, the antenna current does not drop to zero at the ends, so the resistance does not become infinite; the resistance can become very high, however. If the characteristic impedance of the single-wire feeder is between 72 ohms and "very high," we should be able to find a point on the antenna that will match the feedline characteristic impedance.

transmission line theory

When discussing a transmission line of any type, probably the first question to be asked is what its characteristic impedance is (and how it is measured or calculated). With conventional two-conductor transmission lines, such as coaxial cable, one method to determine the characteristic impedance is to measure the impedance seen looking into a length of the line with the far terminals both open- and short-circuited; the square root of the product of these two measurements is the characteristic impedance of the line. However, in the case of the single-wire line, what do you short-circuit it to?

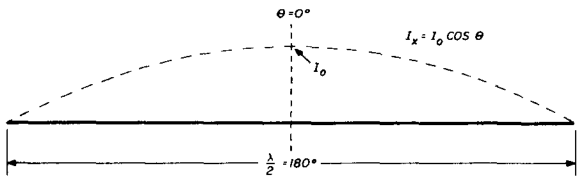


fig. 5. The current distribution along a half-wavelength antenna is essentially cosinusoidal as shown here.

All right, you say, if you can't measure it, can you calculate it? The usual way to calculate the characteristic impedance of a transmission line is to calculate its inductance per unit length and the capacitance between the conductors per unit length; then divide the inductance by the capacitance and take the square root:

$$Z_o = \sqrt{\frac{L \text{ per unit length}}{C \text{ per unit length}}} \qquad (8)$$

It is not difficult to calculate the inductance per unit length of a single-wire line, but the capacitance per unit length is another matter. The capacitance to what? There is no other conductor, so this leaves ground. However, the capacitance of the single-feed wire to ground can be expected to vary over a fairly wide range as the feedline leaves the station — relatively close to ground — and winds its way up to the antenna high in the air. And as the capacitance varies, so does the characteristic impedance of the line. *Hence, the calculation approach is not practical.*

table 1. Characteristic impedance of long wires.

conductor diameter mm	inches	AWG	characteristic impedance, Z_o (ohms) 3.5 MHz	7.0 MHz	14.0 MHz	28.0 MHz	56 MHz
12.5	0.500	—	560	518	475	435	393
6.5	0.250	—	600	560	518	475	435
3.3	0.128	no. 8	641	600	559	516	474
2.6	0.102	no. 10	654	613	572	530	490
2.1	0.080	no. 12	669	628	586	545	503
1.6	0.064	no. 14	684	643	600	560	517

This can be seen more graphically by taking a slightly different approach. Consider **fig. 6** which shows the single-wire feeder and its ground image; this converts the single-wire transmission line into a balanced line. This might be used to calculate the characteristic impedance of the line, but half the distance between the actual line and its image is in the ground whose dielectric constant and other electrical parameters are not accurately known. This makes it difficult, if not impossible, to accurately calculate the characteristic impedance.

Using the representation of **fig. 6** does point up the fact that the single-wire feedline can be explained as one-half of a variable spaced balanced line, with two wires of the balanced line being closer together at the bottom and gradually increasing in spacing as they approach the antenna. Tranmission lines of this types are known as "tapered lines." Since the wire diameter is the same all along its length, the characteristic impedance will gradually increase as the line approaches the antenna. This also means that, assuming the feedline is matched to the antenna, the current measured along the line will not be constant, but will slowly decrease as one moves toward the antenna. Therefore, the current measured at the input end of the feedline should be greater than the current measured at the antenna, even when the system is matched.

After explaining so carefully why it is so difficult to calculate the characteristic impedance of a single-wire feedline, I was rather nonplussed to find the following equation in an old publication:[7]

$$Z_o = 138 \left[\log \frac{0.56\lambda}{2\pi r}\right] \qquad (9)$$

Where

Z_0 = characteristic impedance of single-wire line

r = radius of conductor

λ = operating wavelength

Table 1 gives several values of characteristic impedance, taken from reference 8. As no sources or references are given, I do not know what approximations or assumptions were made in this formula's derivation and can not attest to its accuracy. Since wavelength is included in the calculation, note that characteristic impedance varies with frequency!

The fact that it is not practical either to measure or calculate the characteristic impedance of the single-wire feedline does not mean that the line does not have a characteristic impedance. It does — we just don't know what it is. While this will certainly affect the procedure used to match the feedline to the antenna, it should not stop us from obtaining a match.

bandwidth considerations

Although the bandwidth of the Windom antenna has always been assumed to be large, I have never seen it discussed in the literature. My own experience indicates that while an off-center-fed antenna can be used over a relatively wide range of frequencies, it operates as a true Windom only over a very narrow bandwidth.

As shown in **fig. 4**, the Windom antenna is analogous to a parallel-tuned circuit. The Q of a parallel-tuned circuit can be defined as the ratio of either the inductive or capacitive branch current (at resonance, the two are equal) to the line current. Therefore, the Q of a Windom can be easily measured by using the circuit of **fig. 7**.

If the antenna operates as a true Windom over a relatively narrow bandwidth, why does the antenna work so well as a wide band antenna? Let's look at **fig. 8A**. As the frequency is increased above the *Windom frequency*, the current in the long end decreases while the short-end current increases. The long end loses its effectiveness as a radiator and the Windom antenna degenerates to a random-length, single-wire antenna with the single-wire feedline operating as part of the radiating system. Below the

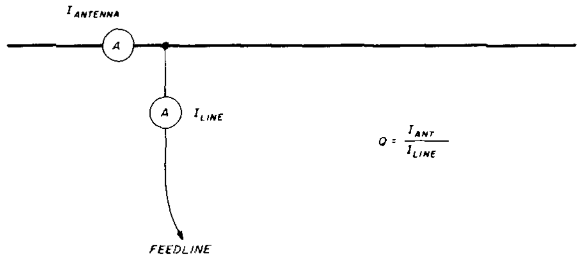


fig. 7. Test circuit for measuring the Q of an off-center-fed Windom antenna.

Windom frequency the current in the long end predominates and the antenna appears as in **fig. 8B**. *Again the feedline will radiate.*

harmonic operation

The Windom antenna has a theoretical advantage over the balanced, center-fed dipole because the Windom will resonate on even-order harmonics or, more accurately, on *approximate* harmonics. Because of the *end effect*, harmonic resonant frequencies are not integral multiples of the half-wave resonant frequency. An antenna that is half-wave resonant at 3.525 MHz, for example, will have harmonic resonant frequencies at 7.235, 14.656, 22.077, and 29.498 MHz. Note that some of these frequencies are not in an amateur band. Thus, if multiband operation is desired, the antenna must be operated off-resonance on some bands and might not load up well.

As mentioned in the bandwidth discussion, the antenna may operate satisfactorily at other than its Windom frequency, but as a random-length antenna. It is also doubtful whether the feedline tie point will be sufficiently accurate on the harmonic bands to provide a good match. My own measurements indicate that it will not.

Those amateurs who are contemplating harmonic operation of the Windom antenna should read Wrigley's excellent article on harmonic operation of dipoles.[8] My own experience with the Windom, though limited, bears out Wrigley's comments. The antenna will operate on harmonics, probably not as a Windom, but rather as a random-length antenna.

efficiency and radiation

The final factors I will discuss are the efficiency

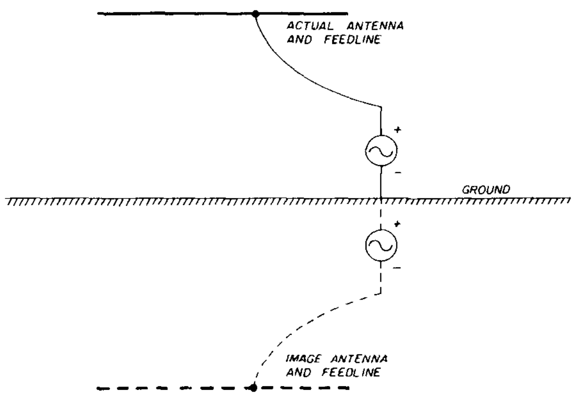


fig. 6. A single-wire feedline and its ground image.

and radiation of the single-wire feedline. I have not attempted measurements of this type myself, but will mention the results from reference 1 which describes the work done by J. D. Ryder and E. D. Shipley. Ryder and Shipley report that for a 365-meter (1200-foot) feedline driving a 15-meter (50-foot) antenna,

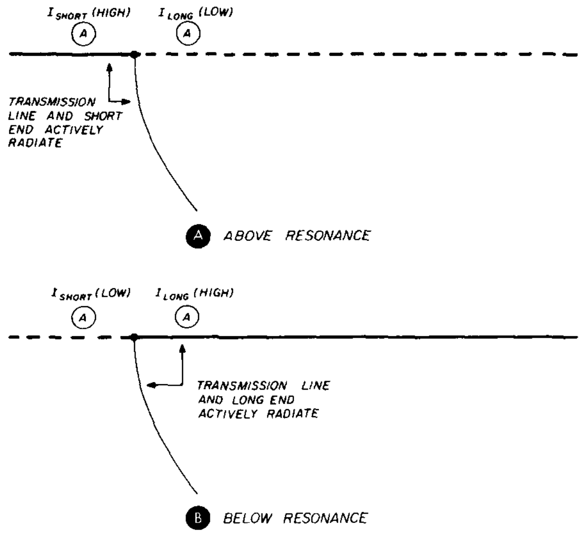


fig. 8. Operation of the Windom antenna above resonance, A, and below resonance, B. Above resonance the current in the longer leg loses its effectiveness as a radiator because rf current decreases and the antenna degenerates to an end-fed single-wire antenna with the feedline as part of the radiator. Below resonance the rf current in the short end decreases and the antenna again operates as a random-length end-fed antenna.

all the measured losses in the system could be accounted for by the I^2R loss of the feedline. Radiation losses were too small to measure. The accuracy of the power measurements was estimated to be within 4 per cent. Therefore, it appears that the radiation from a single-wire feedline operating at its matched (Windom) frequency is negligible, while at other frequencies feedline radiation may be appreciable.

adjusting the antenna

The biggest problem I found with Windom's *QST* article was his failure to recognize the very complex variation of antenna impedance with height and local ground conditions; he assumed that the feedline tap could be placed on the basis of distance measurements alone. This myth has been perpetuated by the ARRL *Handbooks* and *Antenna Manuals*, and by the *Radio Handbook* as well.

With this in mind, the most important question facing the amateur who installs a Windom is "How do I adjust the antenna?" Basically, there are two adjustments to make: 1) the antenna must be cut to the desired frequency, and 2) the single-wire feedline must be connected to the proper place on the antenna to provide a good match for the transmission line.

The first problem must be solved first — the antenna must be cut to the desired frequency (or the actual resonant frequency of the antenna must be accurately known). This is necessary to insure that the antenna will present a pure resistive load.

The method for finding resonance recommended by Everitt-Byrne proves to be very simple, yet very exact. As discussed earlier for the analog with a parallel-tuned circuit, when the tuned circuit is at resonance, the currents in the inductive and capacitive branches are equal. In the Windom the rf currents can be measured by placing rf ammeters in the short and the long sides of the antenna as shown in **fig. 9**. The antenna is resonant at the frequency where the rf currents in the two ends are equal.

If resonance at a specific frequency is desired, begin by cutting the antenna about 1 meter (3-1/2 feet) too long (on 80 meters). Set the transmitter to the desired frequency and prune the ends of the antenna until the current in the two sides is equal. I found I had a strong psychological urge to cut the antenna length from the end with the larger current; actually, the antenna may be cut from whichever end is more convenient.

If a knowledge of the actual resonant frequency of the antenna is all that you want, connect the ammeters as shown in **fig. 9** and vary the transmitter frequency until the two currents are equal.

From a practical point of view, the situation is not as simple as depicted in **fig. 9**. The stresses in a wire antenna are considerable and would probably pull most meter cases apart. Therefore, a means must be devised to take the mechanical stresses off the meters. I mounted the two meters in a piece of plexiglass and suspended the plexiglass from the antenna with spring clips of the type used at the end of dog leashes. The antenna is broken with an insulator and the single-wire feedline is supported by a second insulator. The photographs show the plexiglass meter bracket and how it is suspended from the antenna. I

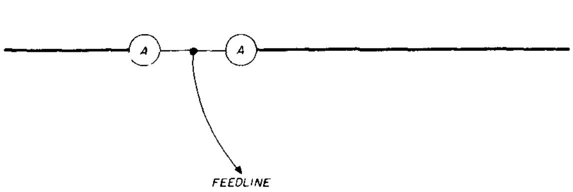
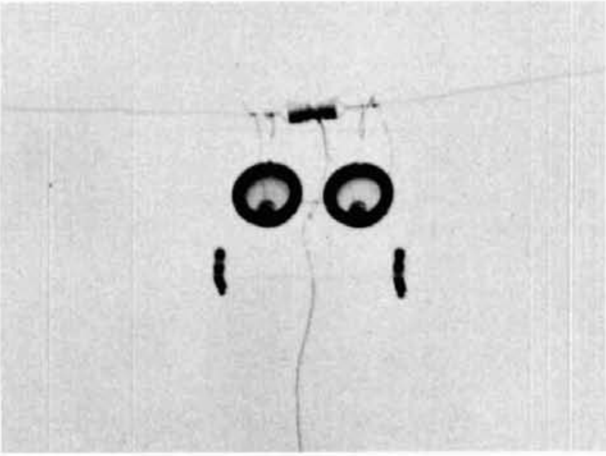


fig. 9. Test circuit for determining resonance of an off-center-fed antenna. When the rf current in the two ends is equal, the antenna is resonant. If the current in the long end is higher than in the short side, the antenna is too long; if the short side current is greater, the antenna is too short.



The test meters installed in the antenna as seen from the ground. The antenna is not quite at resonance since the currents are not equal. This was taken with a 200mm lens.

used 1-ampere meters with an output power of about 90 watts. The resonant currents were 0.76 ampere on 80 meters. This value will depend on how far off-center the feedline is initially attached.

When measuring the rf current, if the current in the short side of the new antenna is lower than in the long side, the antenna is too short; if the long side current is greater, the antenna is too long. See **fig. 10**.

I highly recommend that the transmitter power be brought up slowly and that you have a friend watch the meters to make sure they don't go off scale; thermoammeters are not noted for their tolerance to overloads. The meters, of course, must be read from a distance. I used a 25-power spotting telescope. Higher magnification might be better, depending on how high your antenna is — try to pick a windless day so the meters don't bounce around while you try to focus on them!

Another possibility is a telephoto lens on a 35 mm single-lens reflex camera. However, I found my 135 mm lens was much too short to do any good; I estimate a lens 300 mm or longer would be needed if your antenna is exceptionally high.

Using this method of adjusting the antenna length, I was able to notice a change of as little as 5 to 7 cm (2 to 3 inches) near the resonant point or a frequency change of less than 10 kHz (on 80 meters). The primary factor in limiting the accuracy of this method is the problem of reading the meters at great distances. Much greater accuracy could be obtained by using remote thermocouples — mount the thermocouples on the antenna and run leads to the meters on the ground — or better yet, at your operating position. Unfortunately, remote-reading rf ammeters are very expensive and hard to find!

It is interesting to note that the two-meter method of determining antenna resonance is not limited to the Windom antenna; it can be applied to other dipoles as well. Simply short-circuit the center insulator, insert a second insulator at some convenient place, install an off-center feedline and the meters, and check. The method can also be used to determine harmonic resonant frequencies.

I am surprised that the antenna manual publishers have not presented this procedure before; it is relatively simple and straightforward and has been available since 1929. To the best of my knowledge, the only discussion of this technique to appear in the amateur literature, since Windom's 1929 article, was written by Paul Rockwell, W3AFM, in 1963.[9]

adjusting the feedline

After the antenna has been cut to the proper length, the correct feedpoint can be determined. At first glance this appears to be a formidable problem since we only approximately know the characteristic impedance of the feedline and only approximately know how the radiation resistance varies along the antenna. However, we do know certain characteristics of transmission lines which are helpful; namely, that when a transmission line is terminated in its characteristic impedance, 1) the input impedance of the line is a pure resistance, and 2) the current or voltage is constant along the line (neglecting the impedance taper effect along the line). Either of these facts may be used to find the proper feedpoint.

It is absolutely necessary to make all feedline tests

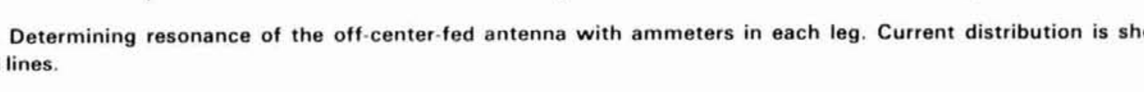


fig. 10. Determining resonance of the off-center-fed antenna with ammeters in each leg. Current distribution is shown by dashed lines.

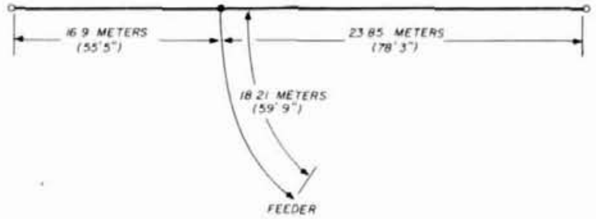


fig. 11. Dimensions of the experimental Windom antenna built by K4KJ. Resonant plots of this antenna for 80, 40, and 20 meters are shown in figs. 12, 13, and 14, respectively.

at the resonant frequency of the antenna. Only at this frequency will the input resistance of the antenna be purely resistive.

For my own tests I used a General Radio 916A rf impedance bridge to measure the input impedance of the line.[10] This measurement could also be made with an RX noise bridge such as the one described in reference 11. Simply measure the input impedance of the line at the antenna resonant frequency, and adjust the feedline tap along the antenna until the reactance component of the input impedance is zero.

If an rf impedance bridge is not available, try the method recommended by the old timers: if the feedline is on the order of a half-wavelength long, insert four rf ammeters in the lower half of the line and adjust the tap for identical current on all four meters. In this case, it will be assumed that the impedance taper along the lower half of the line is negligible so that line current will be constant.

If rf ammeters are not readily available, try soldering neon bulbs along the feeder and adjust the tap point for a constant brightness. This is best done at night, but be prepared for startled neighbors!

results

My Windom antenna was configured as shown in **fig. 11**. With this arrangement the results of my 80-meter measurements are plotted in **fig. 12** which

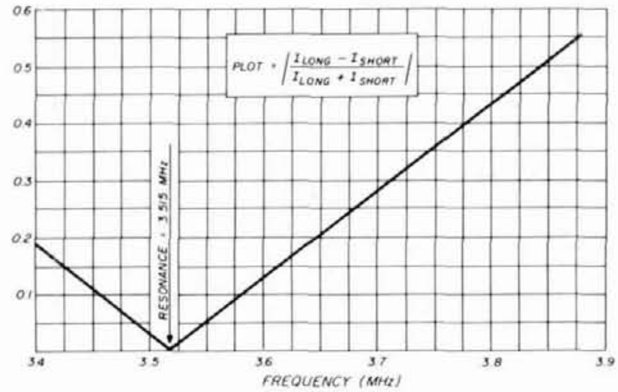
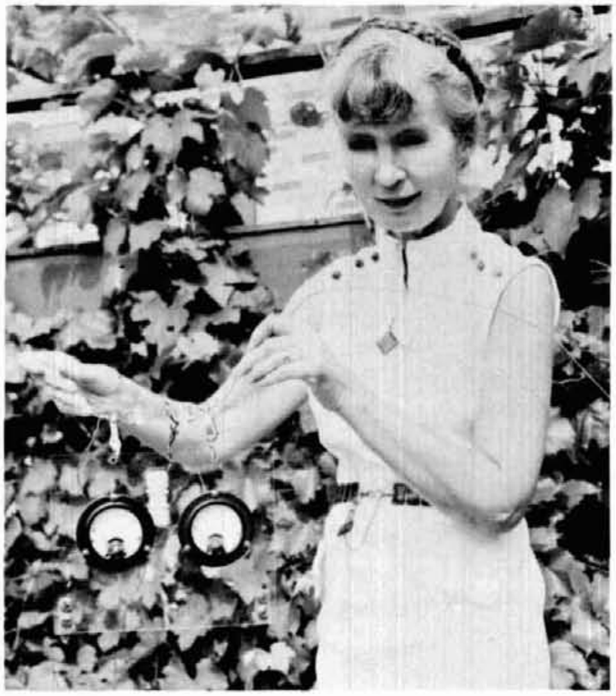


fig. 12. Resonance of the 80-meter Windom antenna of fig. 11, as determined by the method described by Everitt and Byrne (see text).

shows the difference between the antenna current in the two legs. To remove any variations due to changes in rf power input, I referenced everything to the sum of the two currents. Taking the absolute magnitude of the differences eliminates the problem of algebraic sign.

$$plot = \left| \frac{I_{long} - I_{short}}{I_{long} + I_{short}} \right| \quad (10)$$

The resonant frequency is where the plot equals zero. **Figs. 13** and **14** show the same factor for the same antenna on the 40- and 20-meter bands, respectively. Notice that the resonant 20-meter frequency is above the amateur band.



K4KJ's wife holding the plexiglass meter panel as installed in the antenna. The feed-line runs out the bottom.

The input impedance to the feedline is shown in **fig. 15** for 80 meters; the input impedance is anything but constant. The reactance is zero a little below 3.525 MHz at which point the input resistance is 590 ohms. **Fig. 16** shows the same factor over the 40-meter band. Note that the zero reactance point falls above the band limits at which frequency the input resistance is more than 1000 ohms. This indicates to me that the feedline is not matched at any frequency. The input impedance on 20 meters is graphed in **fig. 17** and the same general comments apply.

conclusions

As a result of my experiments, I have reached the following conclusions concerning the off-center-fed antenna:

1. The Windom antenna is very simple and economical to build and, if the proper procedures are used, relatively simple to adjust.

2. The bandwidth as a true Windom antenna is relatively narrow. The antenna will operate relatively well, over a much wider frequency range as a random-length antenna, however.

3. Based on the above data, it is my opinion the Windom antenna can be properly matched for only one amateur band. On other bands, the comments of item 2, above, apply. After reviewing the comments of Windom and Ryder to an early draft of this article, however, I am not convinced that the tap is at the optimum point. It may be possible to improve the match on 40 meters.

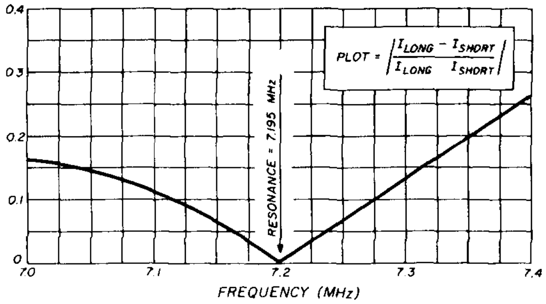


fig. 13. Resonant curve of the Windom antenna of fig. 11 on 40 meters.

4. Those amateurs whose antenna requirements are modest and whose circumstances dictate a simple and inexpensive antenna will continue to find that the off-center-fed antenna will serve their needs.

acknowledgement

Some of the historical data on the Windom antenna was obtained from Paul Rockwell, W3AFM, who very kindly made his files available to me. I would like to thank Messers Everitt, Ryder, Byrne, and Windom, who very kindly made extensive comments on a draft of this article. In addition, Professor Ryder provided previously unpublished technical data which space does not permit me to include.

postscript

The Windom antenna, one of the oldest antennas developed for amateur use, has had a complicated history, one as interesting as the theory of the antenna itself. And the people who have contributed to its development are equally intriguing:

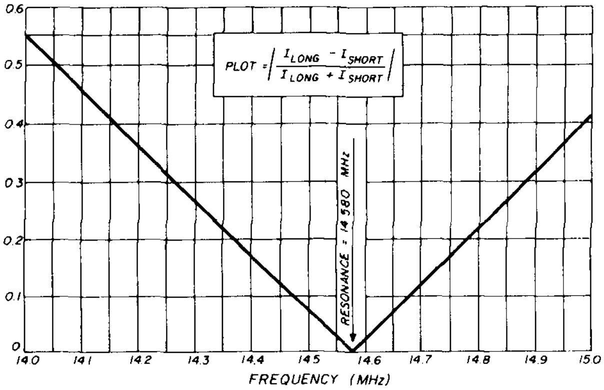


fig. 14. Resonant curve of the Windom antenna of fig. 11 on 20 meters. Note that the antenna is resonant above the amateur band.

Frank Conrad, 8XK, who started it, is called the father of broadcasting. His amateur station became KDKA in Pittsburgh, the first broadcasting station in the country.

Vernon D. Landon, one-half of 8VN, went on to become an eminent scientist with RCA. He has contributed widely to the advancment of electronics with his many very readable papers published in the *Proceedings of the IRE* and in the *RCA Review*.

Loren "Windy" Windom, 8ZG, for whom the antenna was named, did not, oddly enough, pursue electronics professionally, but became a lawyer in Ohio. He is still an active amateur.

John Byrne, who with his thesis partners, E. F. Brooke and A. B. Crawford, did much of the actual work on developing the antenna, became an outstanding research engineer and educator.

John Ryder, now K4IHX, who also worked on the antenna as a thesis project, was to become a well-

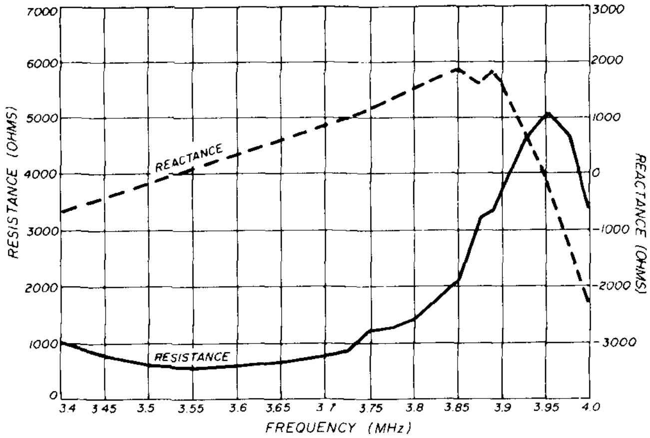


fig. 15. Input impedance of the single-wire feedline to the Windom of fig. 11 on 80 meters, as measured with a General Radio 916A rf impedance bridge. The reactance is zero at 3.520 MHz.

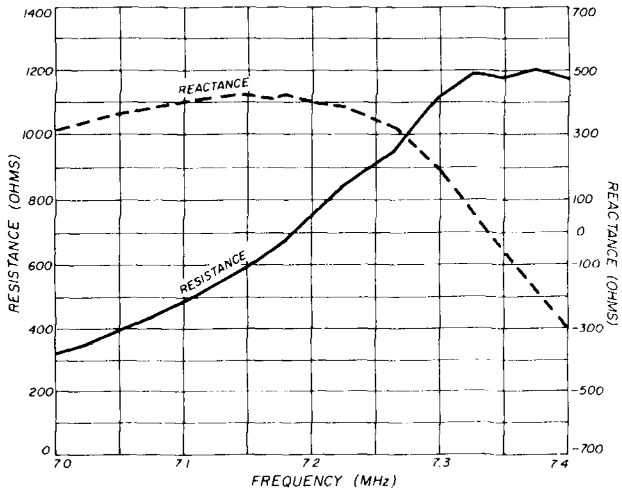


fig. 16. Feedline input impedance on 40 meters as measured with an rf impedance bridge. The reactance is zero at about 7.338 MHz; the resistance at this frequency is more than 1000 ohms.

known educator, an author of several electronics textbooks, and a president of the IRE. He is an active amateur in Florida.

William L. Everitt, as faculty advisor to the students who worked on the antenna, contributed much to its development. Everitt began his amateur career in 1914 as 2ABI; in 1921 he became 8CRI. When he decided to go into communications professionally, he dropped amateur radio because he did not want to have the same vocation and avocation. Everitt was later to become a prominent author and educator. He retired as Dean of Engineering at the University of Illinois and is now Dean Emeritus at that university.

Unfortunately, Dr. Everitt has been ill-treated by the amateur community. His work on the off-center-fed antenna is largely unknown and certainly unrecognized. Unhappily, the same thing happened to him a second time. Everitt was the first to describe the use of a pi network as a coupling device. He published this work in the *Proceedings of the IRE* in 1931[12] and in *Communications*;[13] the pi network is also described in his book, *Communication Engineering*.[14] Arthur Collins, W9CXX, of Cedar Rapids, Iowa, recognized the advantages of the pi network to couple the output stage of a transmitter to a transmission line, and used it in his transmitters. This application played an important part in establishing the reputation that Collins equipment will load up to "anything." Collins described the pi network to the amateur community in a *QST* article[15] and in a similar article *Radio*,[16] and the network became known by old-timers as the "Collins Coupler" instead of, perhaps, the "Everitt Easy Loader."

I believe these are excellent examples of how people who have made notable contributions to the advancement of electronics developed their interest in electronics through amateur radio. It would be interesting to be able to look into a crystal ball to see how the many young people who are today joining the ranks of amateur radio through high school science classes, or as CBers, will go on to make significant contributions to future electronics.

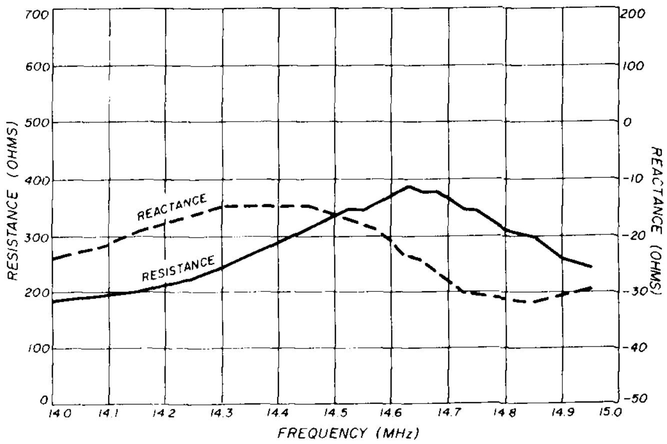


fig. 17. Feedline input impedance on 20 meters as measured with rf impedance bridge. The reactance never falls to zero over the frequency range from 14 to 15 MHz.

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ham radio

selective receiving antennas: a progress report

An active receiving antenna system useful from 75 through 10 meters capable of directing a null toward the interference source

By **Henry S. Keen, W5TRS,** Fox, Arkansas, 72051

The concept that man-made interference is a receiving problem that can best be handled by improved receiving techniques rather than increased power, voice processing, and similar transmitting enhancements, can hardly be disputed. While the high-power advocate may seem to be solving his own difficulties, he can become a major part of the problem to the rest of us.

What to do? One can't always outshout the opposition. No amount of racket on the frequency can prevent signals from point A reaching point B. Only conditions of propagation can do that. What such racket does do, however, is drown out and prevent copy of the signals from A, unless means can be devised at B to make the interference self-destruct. The signal *strength* produced by a given receiving antenna is of little importance compared to the signal-to-noise ratio it delivers. And you'd better believe that all forms of man-made interference must realistically be classified as noise!

Therefore, under interference conditions, it behooves us to consider the use of a separate specialized receiving antenna system. If its signal level production *should be less than that of the transmitting* antenna, the difference can usually be made up by suitable amplification, without loss of signal quality.

This article describes an active receiving antenna system, useful from 75 through 10 meters, with the ability to direct a null toward the source of the interference. The rejection of unwanted signals that can be obtained can often make the difference between solid copy and losing the battle with the interference.

design consideration

What sort of antenna is indicated for this application? Obviously the major requirement is not forward gain but the exact opposite: a broad coverage with a deep null in the response pattern, which can be directed toward the source of interference. Furthermore, you can get many more dB difference in a null than in forward gain. Thus, our problem is to devise a simple inexpensive system having just those characteristics.

The familiar phased array with quarter-wavelength spacing, although capable of an excellent null, is far too cumbersome as a specialized receiving antenna for the lower frequency bands. If its elements should be brought close together for compactness, mutual coupling between the elements increases to a point where phasing and power distribution go haywire and complicate the problem beyond reason.

What to do? In the first place, coupling between the antenna pickup elements should be minimized by making the elements nonresonant. Secondly, an isolating preamplifier following the antenna probe will ensure independence of each unit. Thus, each pickup element acts as a probe in the electrostatic field of the passing wave front.[1]

In an earlier antenna study, a wide range phase control (phasor) was developed for pattern control of a pair of vertical antennas.[2] This phasor, which provided 180° of continuous phase control, was later incorporated in a feasibility study directed originally at the 40-meter foreign broadcast problem.[3] The resulting conclusion was that a useful degree of rejection was possible and that further study would be justified.

All-band system. In this effort, signals from two small, fixed, active antennas with close spacing had been combined in a quadrature hybrid, with control of phase and amplitude. It soon became apparent that, because of the close spacing, 180° phase control was overkill, and the single-band frequency limitations imposed by the phasor and hybrid were undesirable. Therefore, attention was directed toward the development of an all-band receiving antenna system without these restrictions.

Preamplifiers. The first requirement was the development of improved preamplifiers to be used with the antenna probes. They must be wide band, have low noise response, and be stable. In addition to having high gain, they must also have very high input impedance, compatible with the tiny antenna probes with which they would be used (2 foot or 0.6m whips).

To obtain this high impedance, an MPF102 fet is employed as a source follower, with feedback to the bias network. Although this type of input stage may seem to lose signal, the actual power gain is considerable as the impedance level is reduced to several hundred ohms. This stage in turn drives a bipolar transistor stage, which is coupled to the coaxial output line through a 3:1 stepdown transformer trifilar wound on a T50-6 toroid core.

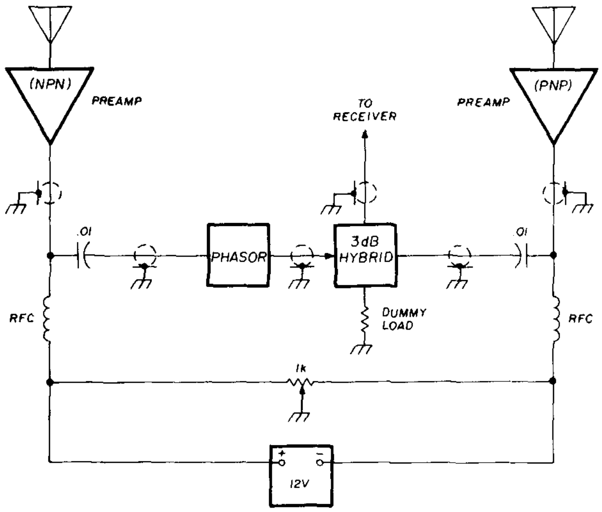


fig. 1. Method of balancing gain in the signal preamplifiers. Balance potentiometer bridges the dc supply.

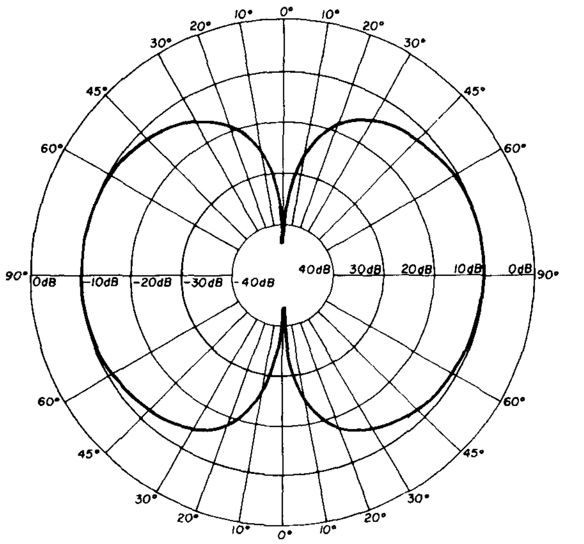


fig. 2. Antenna Figure-8 pattern, calculated for 0.05-wavelength spacing between antennas.

Various means were tried to balance the signal amplitude of one preamplifier against the other, but the method finally chosen was that which had been used in reference 3; that is, by controlling the dc voltages applied to each one (**fig. 1**). In this method the two preamplifiers are in series, one being fed a positive voltage to its npn output stage while the other gets a negative voltage to its pnp output stage.

Identical lengths of coaxial cable were used between the antenna modules and the signal balancing device, so that phasing was obtained mechanically by rotating the boom upon which the modules were mounted. This eliminated the phasor and resulted in a Figure-8 pattern, (**fig. 2**).

Antenna pattern. The very sharpness of the null offers both advantages and disadvantages. The rejection of a local amateur signal, for example, was a walloping 60 dB, assuming 6 dB per S-unit. In a crowded area where one has problems from "the ham down the street" this arrangement may offer much-needed relief. Rejection of skip signals was less effective, as the multipath propagation effects accompanying such signals introduced variations in both phase and amplitude relationships of the signals. This had the effect, particularly on distant signals, of making the *direction* of arrival appear to vary.

It would seem that a reduction in the sharpness of the null would improve this situation. Analysis of the Figure-8 pattern indicates that signals that deviate from the direction of the null will vary as the sine of the angle of deviation. A cardioid pattern, on the other hand, (**fig. 3**) should give a response based upon the delay between the time when a signal strikes the first antenna and when it reaches the second. This parameter should vary as the cosine of the angle of deviation, and is therefore less critical.

Signal combiner. Many different methods of combining signals from the two preamplifiers were also tried and compared. For example, a CA3028 differential amplifier was built with a balanced output transformer, but was found to be unsatisfactory. Spurious signals caused by IMD made it clear that signals from the antenna modules would have to be balanced out before further amplification could be used. Too much continuous amplification, particularly in a very broad spectrum system such as this, pushes the final stage beyond linearity. Signal reduction inherent in the phasing-out process keeps the amplitudes within limits.

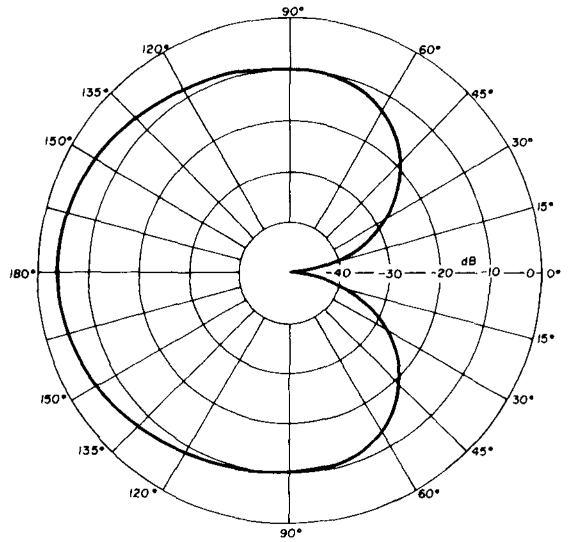


fig. 3. Cardioid pattern, calculated for 0.05-wavelength antenna spacing.

The most effective combiner, other than the frequency-limited quadrature hybrid previously used, was a toroid input transformer with separated primary and secondary windings. A bifilar primary was used for improved balance. The five-turn secondary was close wound and separated as far as possible from the primary (**fig. 4**).

Each coaxial line from the antenna modules is terminated by a 51-ohm resistor to reduce mismatch reflections that might affect the phase of signals presented to the differential transformer.

Common amplifier. The amplifier that follows the differential transformer was originally a broadband unit similar to the preamplifiers. It was found that spurious IMD signals were present, particularly on 20

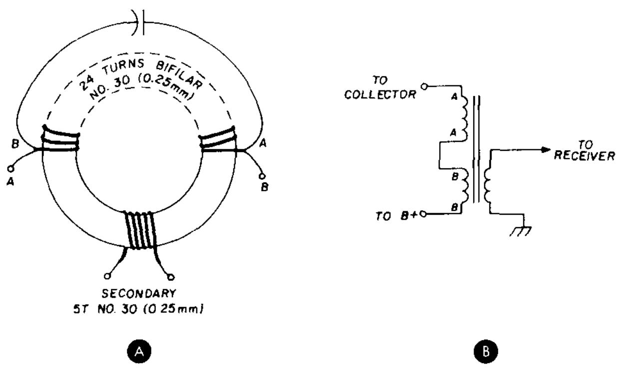
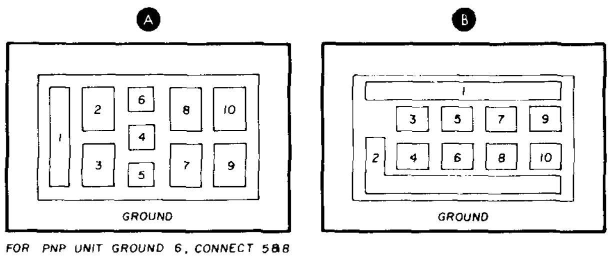


fig. 4. Method of constructing transformer T2 (A), and schematic of output transformer T3 (B).

meters, caused by out-of-band commercial stations causing overloading. At the same time, the overall gain left something to be desired. The cure to both problems was to place a tuned circuit ahead of the common amplifier. The Q of the tuned circuit raised the gain to an acceptable level, while at the same time rejected signals responsible for the IMD.

For operation on different bands some means must be provided for bandswitching this tuned circuit. This is what you must do with your transceiver, so it doesn't represent any additional operating hardships.

Phasing method. The boom upon which the antenna modules were mounted was approximately two meters in length — an arbitrary choice. To obtain a cardioid pattern, the coaxial line from one module to the common amplifier should be longer than the other by approximately 2/3 of the spacing between the antenna probes (allowing for the velocity of propagation in the coaxial lines). Once adjusted, phasing should be independent of frequency, right? Well, not exactly, for allowance is made for the fact that signals arrive at different angles from the ionosphere, and a minor compensation should be provided. Either the phasing extension of the coaxial line can be shortened, or the boom can be tilted to correspond to the angle of arrival of the signal. A compromise angle of 25 degrees was chosen.



FOR PNP UNIT GROUND 6, CONNECT 5&8
FOR NPN UNIT GROUND 5, CONNECT 6&8

fig. 6. PC-board layout for the preamplifiers (A), and common amplifier (B).

Strictly speaking, signals will arrive at different angles on different frequency bands, with different heights and with different conditions of propagation. However, the exact angle does not seem to be too critical. Phase differences in the two preamplifiers will likely have a greater effect. With the Figure-8 pattern, there should be two nulls 180° apart. If not, the change to the cardioid pattern will not be very effective.

To be able to use either pattern, the extension should be an extra piece, which can be added to either of the feed lines. The angle of signal arrival does not affect the Figure-8 pattern.

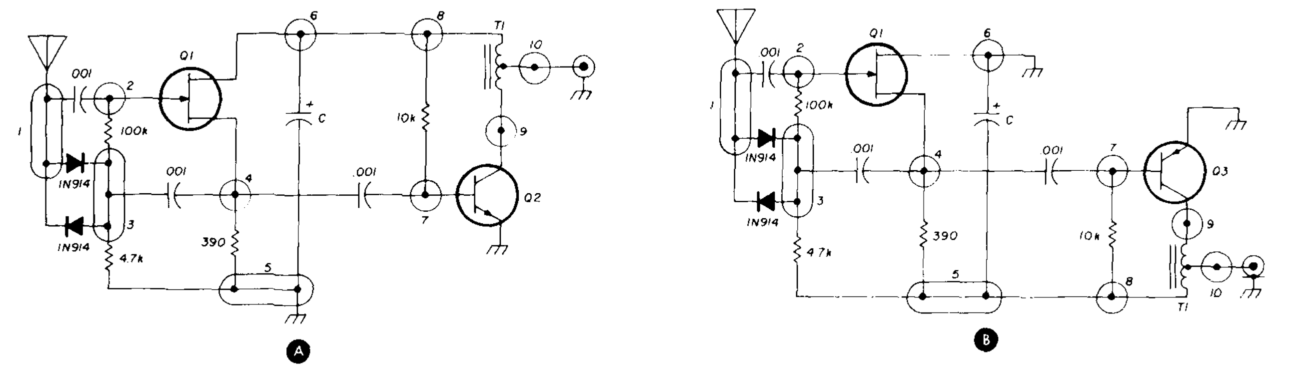
As with all antennas, the higher and more in the clear you can get it, the better. Other antennas or structures in the vicinity can distort the pattern obtained.

construction

The amplifier modules were built on PC boards as shown in **figs. 5** and **6**. Transistor sockets were used to mount the fets, which were inserted after all connections were completed. Although sockets may be dispensed with, if care is used in the assembly process, the ability to quickly select and substitute individual fets to help balance the gain of each module is worth the extra trouble. Where balance is as important as it is in this application, bargain-basement transistors are not recommended. Procedures in making PC boards have been well covered in the literature and are not repeated here. Perf board may make an equally effective substitute if you're not too fussy.

The preamplifier transformers use 22 turns, trifilar wound, of no. 30 (0.25mm) enameled wire on a T50-6 core; the windings are connected in series. Toroids are conveniently mounted by inserting a small rolled cylinder of paper through the center and applying a drop or two of cement in the right places. The preamplifiers are mounted in Miniboxes with a porcelain feedthrough antenna terminal on one end and a UG-625/U BNC connector at the other. Because of their light weight they can be supported by the antenna and output leads.

The common amplifier floats across the power supply, operating from the full 12 volts, neither terminal of which is grounded. Note that the output transformer of this amplifier has an isolated secondary winding, rather than being an autotransformer, as used in the preamplifiers.



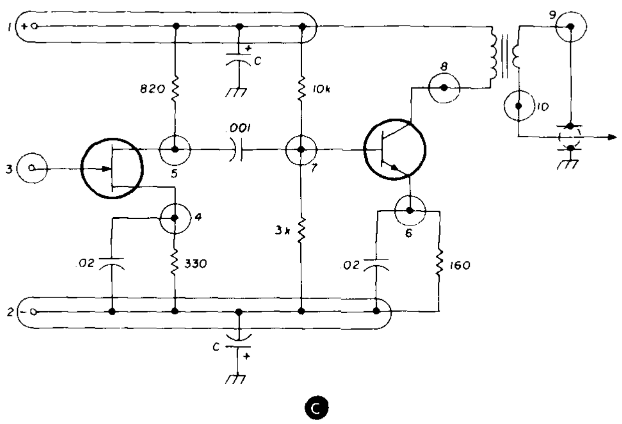


fig. 5. Amplifier schematics with PC-board mounting keys. The npn preamp and pnp preamp are shown in (A) and (B); the common amplifier is shown in (C).

The differential input transformer and the tuned circuit combination is placed in a shield box, made by soldering together bits of scrap PC board. At present, operation on 20 and 40 meters is covered by the tuning range of the trimmer capacitor, although a complete bandswitching arrangement is scheduled for the near future. The coil could be tapped, or shunt capacitors or inductors switched in for operation on the other bands. The overall schematic of the entire system is given in **fig. 7**.

preliminary tests

The common amplifier is first checked out by connecting a receiver to its output terminal and a signal generator to first one input terminal and then the

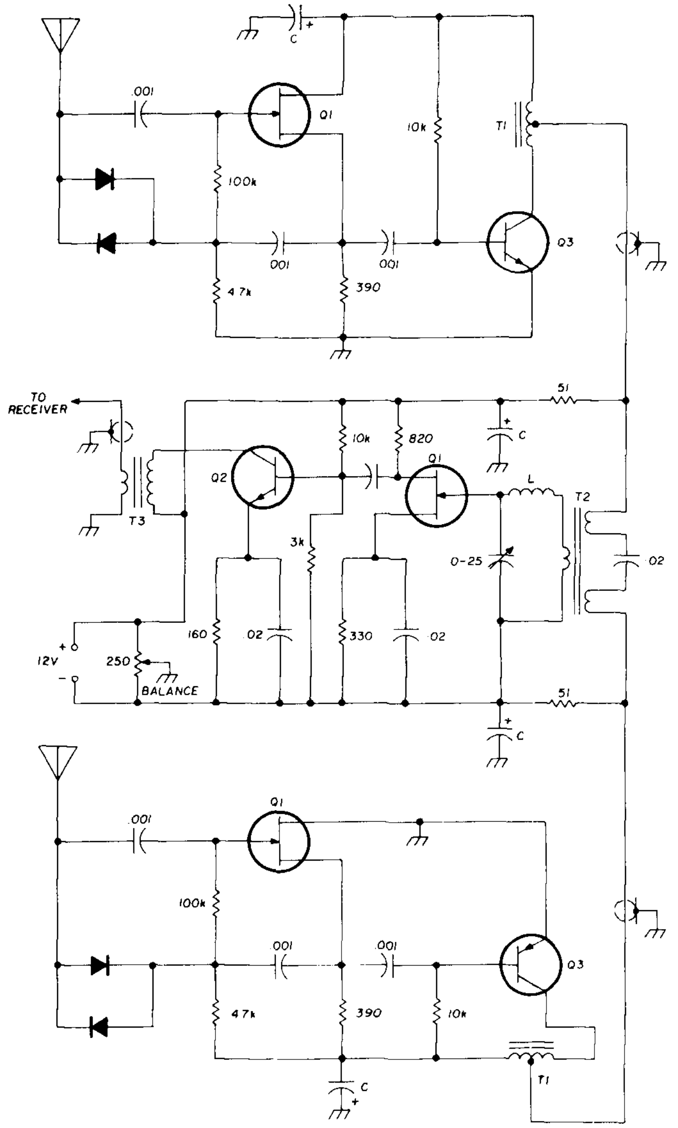


fig. 7. Overall schematic of the selective receiving system.

other. Use a series capacitor to avoid shorting the dc on the input connectors. The tuned circuits should be checked to ensure peaking on each band.

Dc voltages on the input connectors should be checked as the balance potentiometer is rotated. And don't forget to mark the polarity at each connector so that the right preamplifier will be connected to each.

An antenna preamplifier is now connected to the proper input terminal with a random length of coaxial cable. A short piece of wire clipped to the antenna connector should pick up strong signals. Repeat the test with the other preamplifier.

When both antenna modules have passed this preliminary test, you are ready for the final checkout. Both antenna modules are connected to the common amplifier with equal lengths of coaxial cable. Their Miniboxes should be clipped together and the antenna terminals fed through a capacitive tee junction, as in **fig. 8**. The two antenna terminals should not be directly connected together because of the dc potential difference between the npn and pnp preamplifiers.

The signal thus injected into each preamplifier is identical in both phase and amplitude, and the balance potentiometer is adjusted for minimum signal. To get a comparison first adjust for minimum signal then disconnect one of the preamplifiers. A signal increase of from 40 to 50 dB should be obtained. If not, you will have to do some troubleshooting. Ideally the dc voltages at balance should be equal on each preamplifier, but 2:1 may be par for the course.

This null depth test should be made on all bands where the antenna system will be used. If you can't get a deep null at this stage, you are not likely to get it later.

Having passed the null depth test, the system is ready to be installed. Mark the coaxial lines and amplifier terminals, if it hasn't already been done, so that there will be no polarity mixup. I found during the null depth tests that rejection was better if the outer conductors of the coaxial lines were tied together at intervals of several feet (about 1 meter). All these things are better done on the ground before the assembly is raised.

operation

Tuning up and adjusting a system of this kind can be a frustrating experience if you don't go about it the right way. For example, you tune in a good strong signal and start making adjustments only to find that it's in an ssb net, operating VOX, and the signals bounce back and forth so fast that you don't know which one you are hearing. Or else rapid fade makes you think you have a null when you haven't. The foreign broadcasters are particularly troublesome in this respect. If you try watching the S-meter while tuning the antenna, that's no answer either. The trick is to reduce the receiver rf gain enough to disable the agc, and then you can do it by ear. A CW or SSTV station may be a better choice. You may also find that a test signal fed into a separate antenna is a useful aid in adjustment. A TV birdie sometimes provides a good test signal.

The CW man can buy a great deal of interference relief with appropriate filter systems, but the ssb operator, with his wider bandwidth requirement must resort to other means. Some little operating tricks are helpful. For example many 40-meter hams, operating lower sideband among the broadcasters, will operate several hundred to a thousand Hz

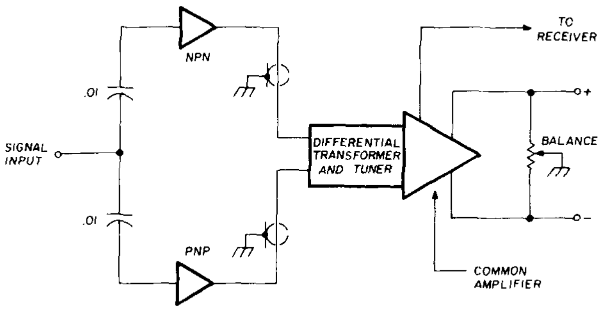


fig. 8. Setup for the null-depth test.

below beat, letting the steep skirt of the crystal filter take out the heterodyne and much of the monkey chatter. If the broadcaster is not too strong, zero-beat works fairly well, particularly with the receiver's rf gain turned down so that the agc doesn't fight you. Often overlooked in interference situations is the simple expedient of switching sidebands. After all, what is so sacred about lower sideband on 75 and 40 or upper sideband on the higher frequency bands? Frequently it's not only easier to switch sidebands than to hunt for a clear frequency but there's a lot less chance of losing your contact in the process.

closing remarks

In our battle for better radio contacts much development work remains to be done. The directional receiving antenna, of which this system is only one possible variation, appears to be a promising area for study. It is still in the experimental stage and there are probably better ways of doing it. Perhaps we owe it to ourselves to investigate some of them.

The major problem in this approach is the method by which we balance out one signal against the other. Alternative methods, such as the phasor-hybrid combination of reference 3, worked very well on a single band. Another method that worked quite well was to obtain a phase reversal in one of the preamplifiers by means of a reversed winding on one of the output transformers, coupled with a differential capacitor ahead of the common amplifier (**fig. 9**). This circuit acted as an amplitude control and eliminated the need for the npn-pnp combination of preamplifiers with the balancing potentiometer. It is probably of equal merit with the system described in this article, but most of us don't have differential capacitors in our junk boxes.

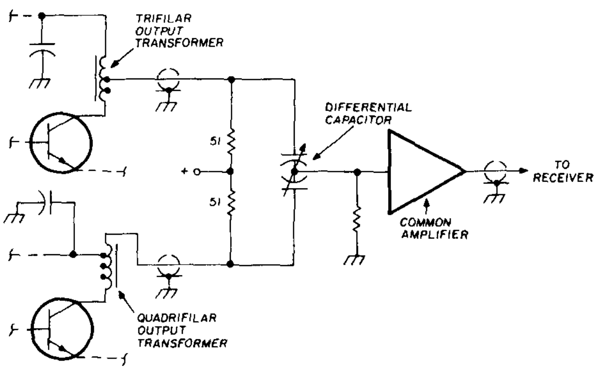


fig. 9. Alternative balancing method, which has been tested. The balancing potentiometer is replaced by a differential capacitor.

Yet another possibility, which has not been tested, is shown in **fig. 10**. In-phase components would be out of phase when arriving at resistor R, while out-of-phase components would add, as in a lattice filter. It would seem that the value of R would have to be quite low for good rejection, but the tuning network should retrieve a useful part of the loss.

No single piece of equipment can solve all of our problems, and this is no exception. There are times when it is very impressive, as well as times when conditions of propagation make it less effective. Advantages of as much as 30 dB have been obtained when conditions were favorable. This has the effect of reducing kilowatts to watts, a worthwhile gain. During a contact, when someone opens up too close to your frequency, it's most gratifying to turn the receiving antenna and hear the interfering signal fade into the background. It doesn't always happen that way as the interference may be in line with your contact; however, it's a satisfying feeling when it does.

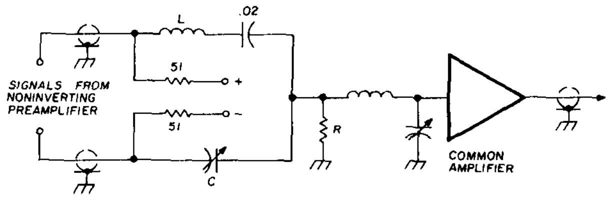


fig. 10. Another alternative balancing system, which has not been tested. In-phase signals would be out of phase at resistor R, while out-of-phase signals would add. The tuning network should compensate for signal loss because of the low value of R.

The forthcoming WARC confab is no guarantee that we'll be any better off when it's over. Can we afford to just sit around and wait?

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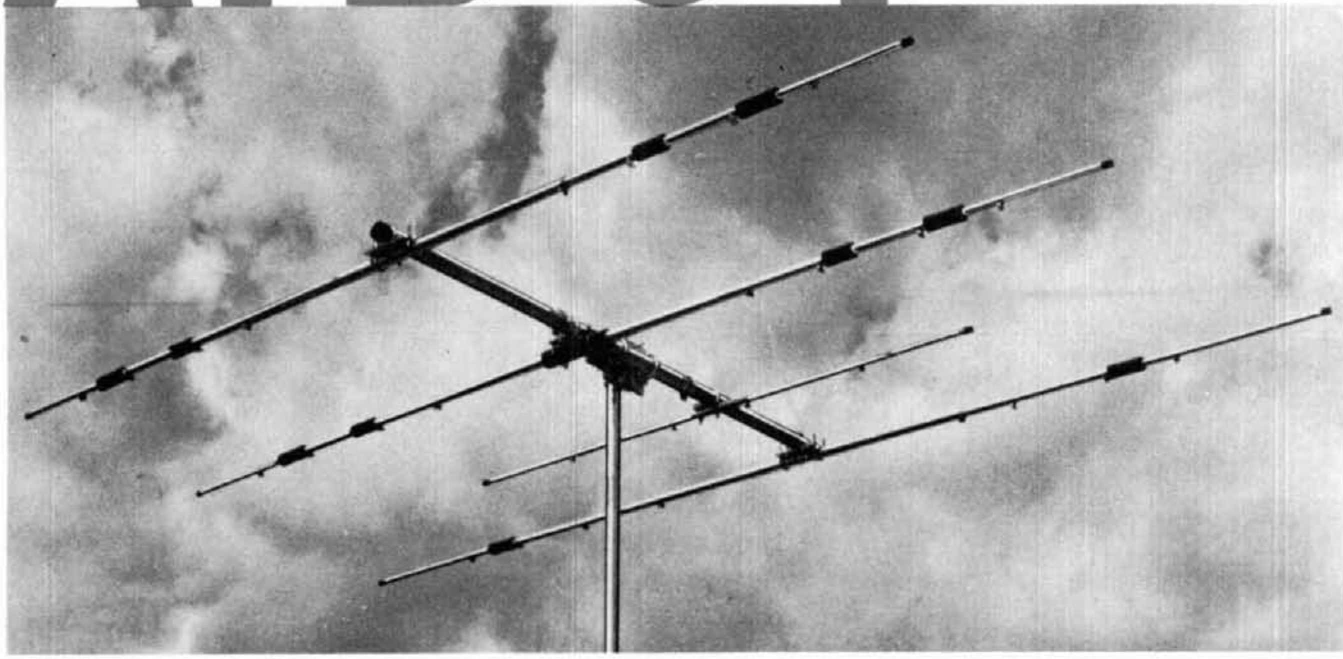

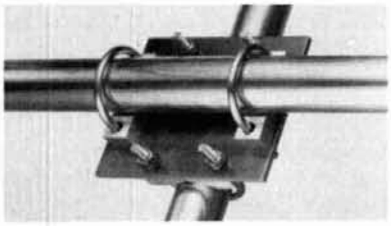

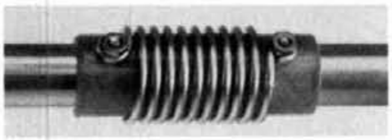

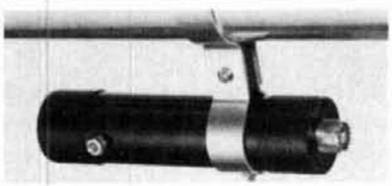

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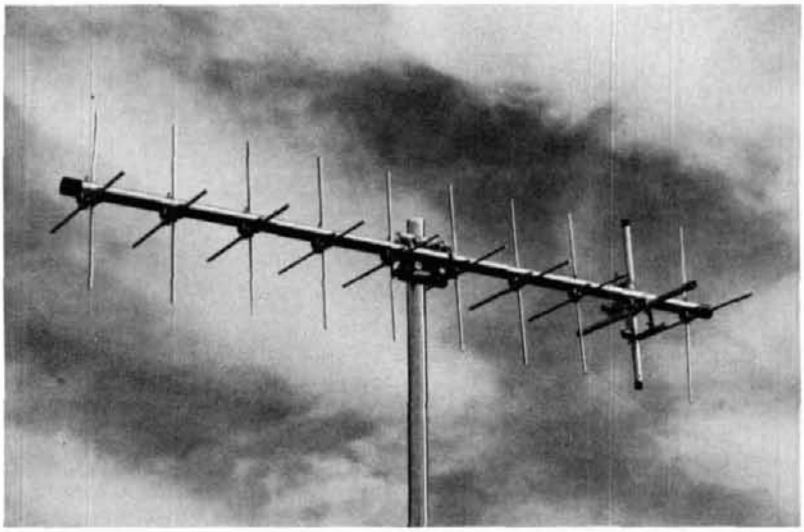

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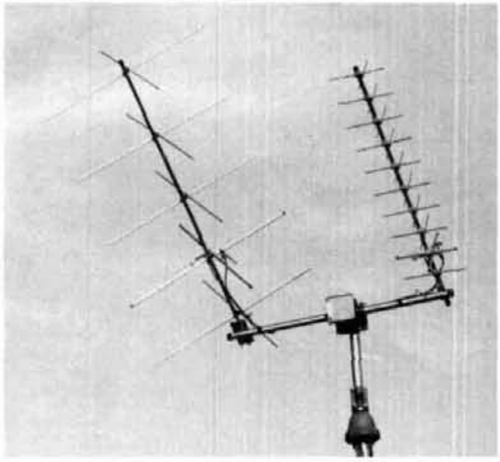

A144-10T $34.95 **A432-20T $49.95**
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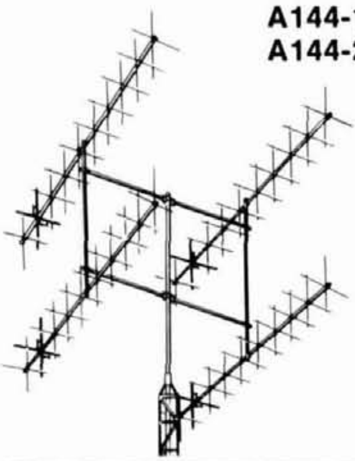

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SPECIFICATIONS

Model	A147-20T	A144-10T	A144-20T	A432-20T
Center Freq. (MHz)	144.5/146.5	145.9	145.9	432
No. Elements	10/10	10	20	20
Weight (lbs.)	6	3.5	6	3.5
Wind Surf. Area (ft.²)	1.42	.74	1.42	.37
Mounting	Center	Rear	Center	Rear
Dimensions (Inches)	40x40x140	40x40x70	40x40x140	14x14x57

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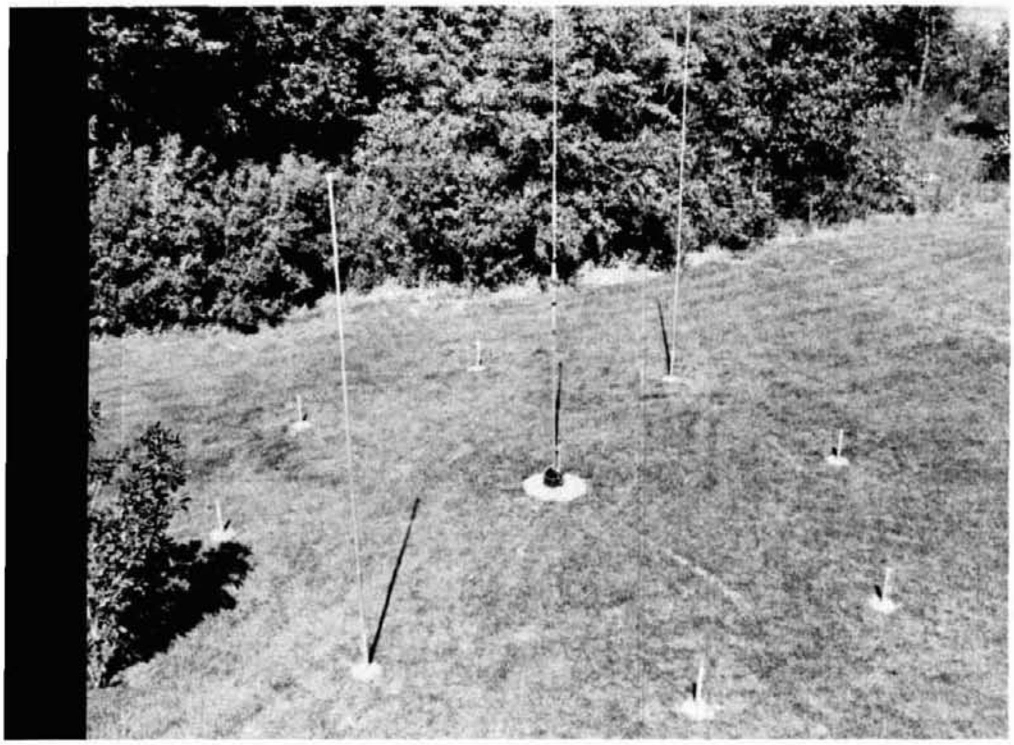
More Details? CHECK—OFF Page 150



multiband vertical antenna system

A good ground system and parasitic elements enhance the performance of a commercial trapped multiband vertical

This project started *out as a simple installation of a* Hy-Gain 18AVT/WB vertical antenna. The 18AVT/WB is a trapped vertical for 80-10 meters that uses top-loading to achieve 80-meter operation. The ground system began as a single 2-1/2 meter (8 foot) ground rod driven in at the base of the antenna with the base of the antenna mounted 30 cm (12 inches) above the ground. The performance was good enough to arouse my curiosity to see what improvements I could experiment with to make the system work even better.

Since a vertical antenna is no better than its grounding system, that was a good starting point. I began by researching the *ARRL Handbook*[1] and *QST* for grounding information. K4ERO's article on ground-

By Ladd Seaberg, WØNCU, Route 1, Atchison, Kansas 66002

ing systems for vertical antennas[2] provided the information I was looking for.

The ground system grew slowly, from four radials, each 10.7 meters (35 feet) long, to the final system of 72 radials shown in **fig. 1**. Theoretically this configuration has a 2 dB power loss on 80 meters at 0.125 wavelength for 24 radials; a 1 dB loss on 40 meters at 0.2 wavelength for 60 radials, and slightly less than 1 dB loss on 20, 15, and 10 meters. As explained by K4ERO in his article, it does little good to increase the radial length unless you also increase the number of radials.

The radial system was surveyed using a 1910 vintage transit with a large magnetic compass. In my location, magnetic north is 8 degrees east of true north. Eight degrees was subtracted from magnetic headings so the cardinal radials are on true compass headings, 45 degrees apart.

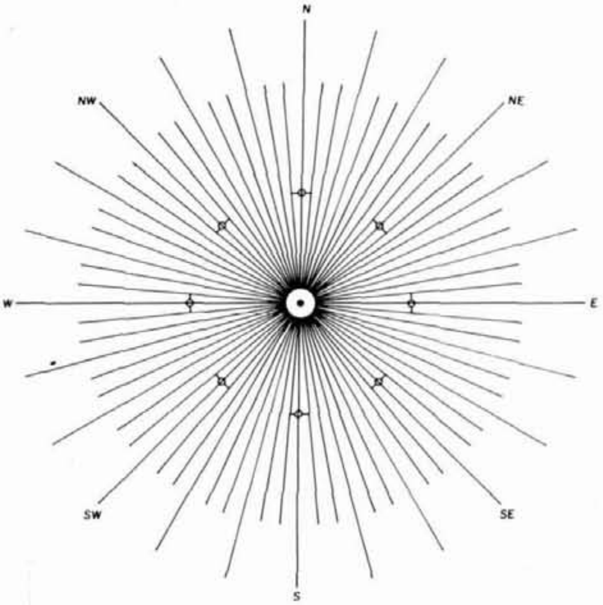


fig. 1. Layout of the radial system used with WØNCU's vertical antenna system for the five high-frequency amateur bands. The system uses a total of 72 radials: 24 are 11 meters (35 feet) long; the other 48 are 8.5 meters (28 feet) long.

Long before I finished the ground system, the idea of making the antenna system directional kept popping into my mind. I looked into both phasing[3,4] and parasitic arrays.[5] I wanted to keep the system simple but I didn't want to lose its multiband capabilities. I finally settled on a three-element parasitic array; it would be directional on 20, 15, and 10 meters and built in such a way that I could remove the director and reflector for omnidirectional use on 80 and 40 meters.

construction

The ground system was built from scrap no. 14

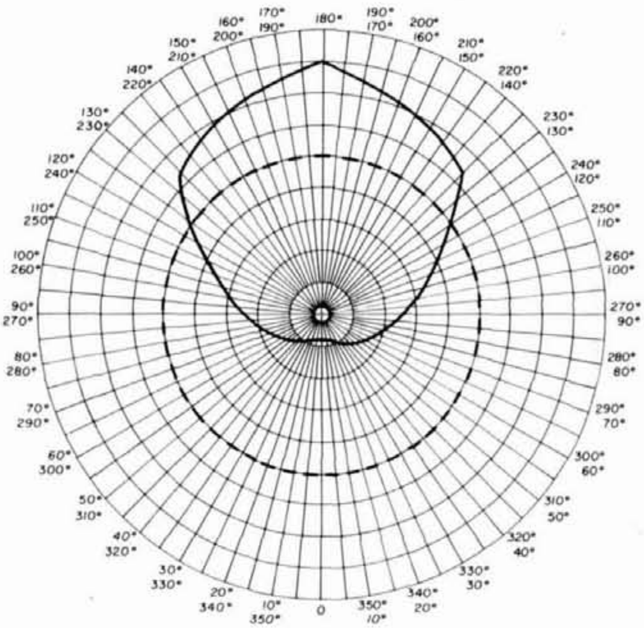
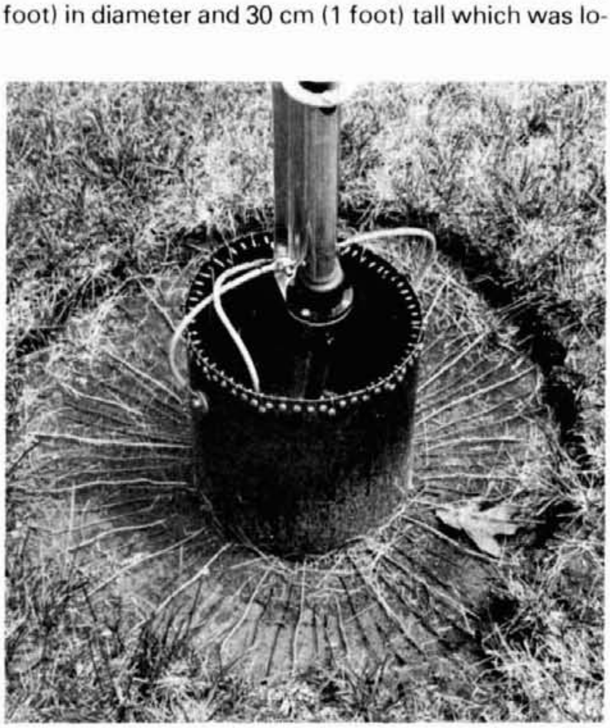


fig. 2. Relative field strength of the three-element vertical beam on 20 meters. The dotted line shows the radiation of the center vertical by itself. Front-to-back ratio is 20 dB; gain is 4.1 dB.

(1.6mm) and no. 16 (1.3mm) copper wire. An edging spade was used to cut a small slit in the grass; then the wire was pushed down into the slit. Using this method, I didn't have to dig up the whole yard. All the wires were terminated on a copper pipe, 30 cm (1 foot) in diameter and 30 cm (1 foot) tall which was lo-



Base of the driven element showing the length of large diameter copper pipe used for terminating the radials. Narrow slits were cut in the lawn to place 72 radials around the base of the antenna.

After the antenna was installed and tuned, the base was covered with plastic sheeting and crushed stone to give it a nice appearance.

cated at the base of the antenna. (The copper pipe was a scrap piece from an old vodka still at the distillery where I work as a chemical engineer; perhaps this antenna should be called the "moonshine vertical!")

The reflectors and directors are built of 25 mm (1 inch) aluminum conduit with 19 mm (3/4 inch) aluminum conduit slipped inside to make a nice fit. The eight ground array supports are made of 3.2 cm (1-1/4 inch) galvanized conduit. They are 1.2 meter (4 feet) long with 75 cm (30 inches) driven below ground. A 6.5 mm (1/4 inch) hole is drilled in the supports at ground level to drain out water. The array elements slip easily into the ground supports; good electrical connection is made with a 6.5 mm (1/4 inch) bolt in the elements which slips into a spade lug on the ground support (see photograph). A wing nut is used to quickly tighten the connection. The spade lug on the ground support is in turn wired to the radial running immediately below it and to the two adjacent radials. The connections below ground are wrapped and soldered.

The RG-8/U coaxial feedline is in a conduit which runs underground approximately 11 meters (35 feet) from the shack and comes up through the copper ground-wire termination pipe.

tuning and measurements

The central driven element was tuned for minimum swr on all bands starting with 28 MHz and working down to the lower frequencies, one band at a time. The antenna dimensions are quite different than those recommended by Hy-Gain because the grounding system changes the resonant characteristics of the antenna.

After this radiator was tuned up, the 20-meter elements were installed on the ground supports. The director was raised and lowered for maximum gain as

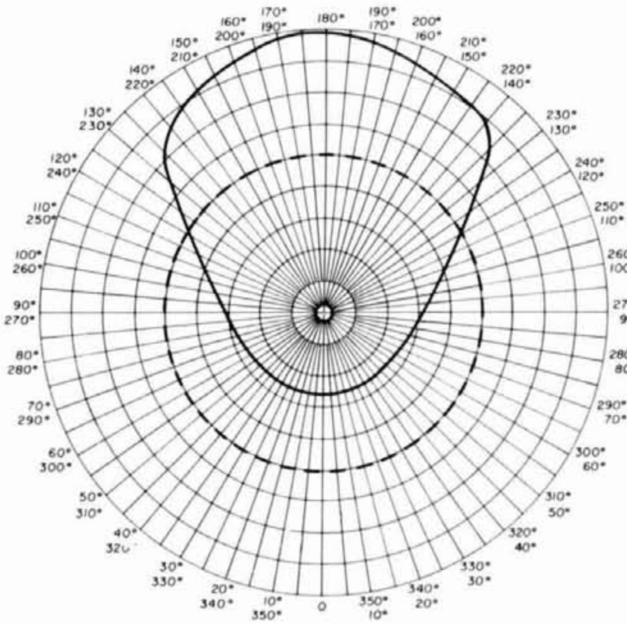


fig. 3. Radiation pattern of the three-element vertical beam on 15 meters, as determined by field strength measurements. The dotted line shows the radiation pattern of the driven element by itself. Front-to-back ratio is 11 dB; gain is approximately 5 dB.

indicated by a field-strength meter; the reflector was raised and lowered for maximum null. The front-to-back ratio was 20 dB as indicated by the field-strength measurements, with 4 dB gain over the vertical without the parasitic elements. This was later confirmed by W9HF when I repeatedly turned the

table 1. Dimensions of the multiband vertical beam and its performance characteristics with the radial ground system shown in fig. 1. The base of the driven element is 30 cm (12 inches) above ground. The height of the antenna on 40 meters is 5.11 meters or 16 feet, 9 inches (from antenna base to bottom of the top hat); the height of the antenna on 80 meters is 6.41 meters or 21 feet, 1/2 inch (from base of antenna to antenna tip).

band	director length	driven element	reflector length	element spacing	front-to-back ratio	gain
20 meters	4.67 m (15′4″)	3.71 m (12′2-1/4″)	5.59 m (18′4″)	0.18λ	20 dB	4.1 dB
15 meters	2.90 m (9′6″)	2.58 m (8′5-3/4″)	3.81 m (12′6″)	0.27λ	11 dB	5.0 dB
10 meters	2.13 m (7′)	2.10 m (6′10-5/8″)	2.74 m (9′)	0.36λ	6.4 dB	5.7 dB

beam to and from him as he gave me reports.

The distance from the driven element to the parasitic elements is 3.8 meters (12-1/2 feet) as recommended by W2FMI for his 20-meter vertical beam.[4] With this element spacing fixed, the front-to-back ratio falls off as the frequency is increased, but the gain over the single vertical increases.

Thus, the front-to-back ratio is 11 dB on 15 meters with 4.95 dB gain over the vertical. The front-to-back ratio on 10 meters is 6.35 dB with 5.68 dB gain over the center vertical by itself (see **table 1**). Even though this performance is a compromise between

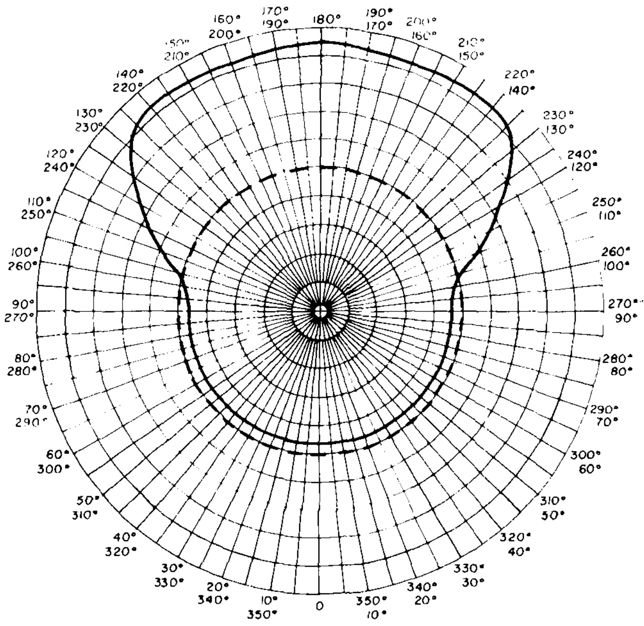


fig. 4. Relative field strength of the three-element vertical beam on 10 meters. Front-to-back ratio is 6.4 dB; gain is about 5.7 dB.

bands, it works out nicely since selectivity is more important for a crowded 20-meter band, and gain is more important for 10-meter DX work. (Comparative field-strength measurements for 10, 15, and 20 meters are shown in **figs. 2, 3,** and **4.**) The 10- and 15-meter elements were tuned in the same manner as the 20-meter elements using trial and error while taking field-strength measurements.

WB0SOT helped man the station while I took the field measurements and made adjustments to the elements. When Joe went back to college in the fall, I ran 75 meters (250 feet) of extension cord from my operating position out to the measurement site so I could key the transmitter and take field-strength readings. The readings were taken with a Heath HD-1426 relative field strength meter located about 50 meters (175 feet) from the antenna.

It is possible to build a combination antenna which works well as an 80- and 40-meter vertical and as a

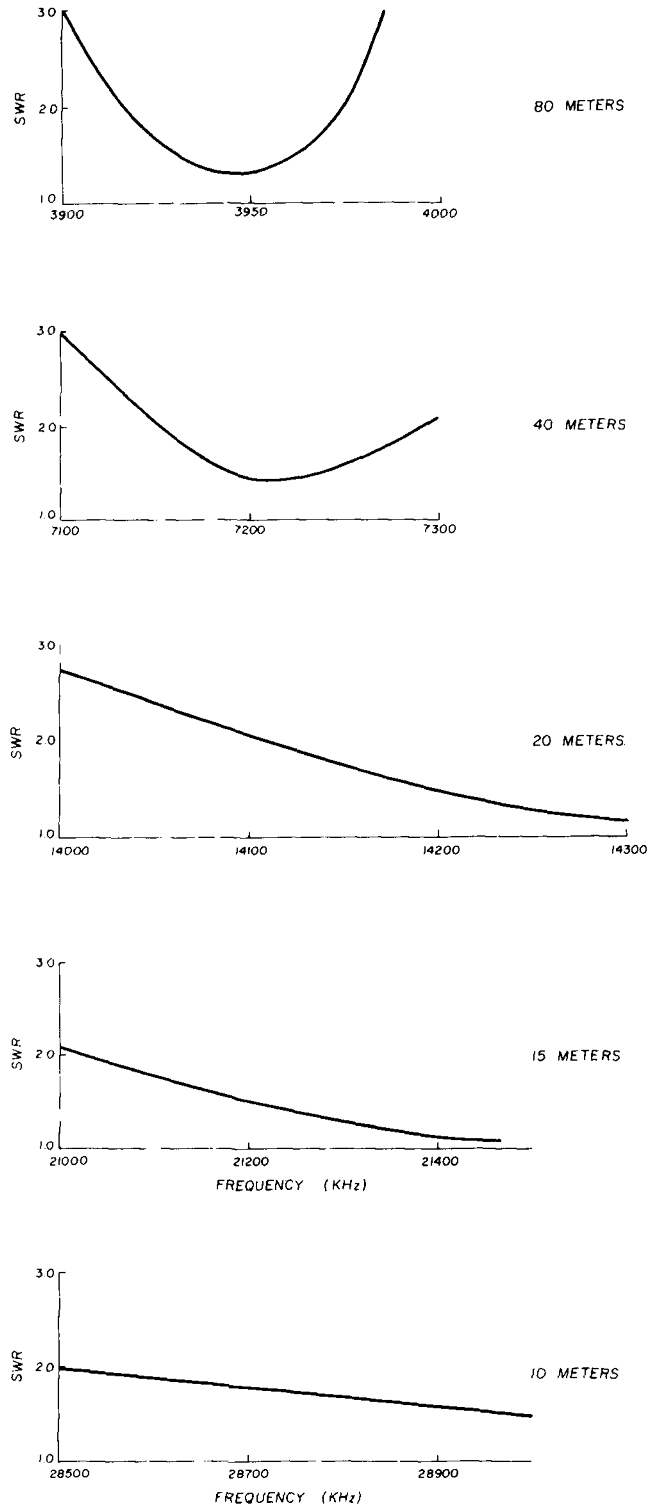
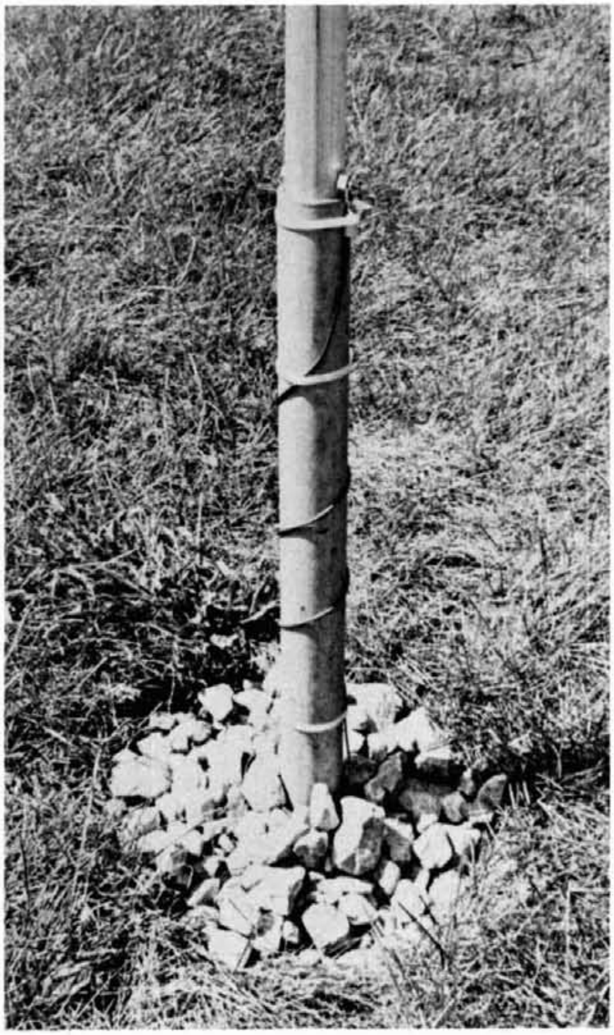


fig. 5. SWR curves for the multiband vertical antenna for each of the five high-frequency amateur bands. The antenna system could also be resonated for use on the CW ends of the bands if desired.

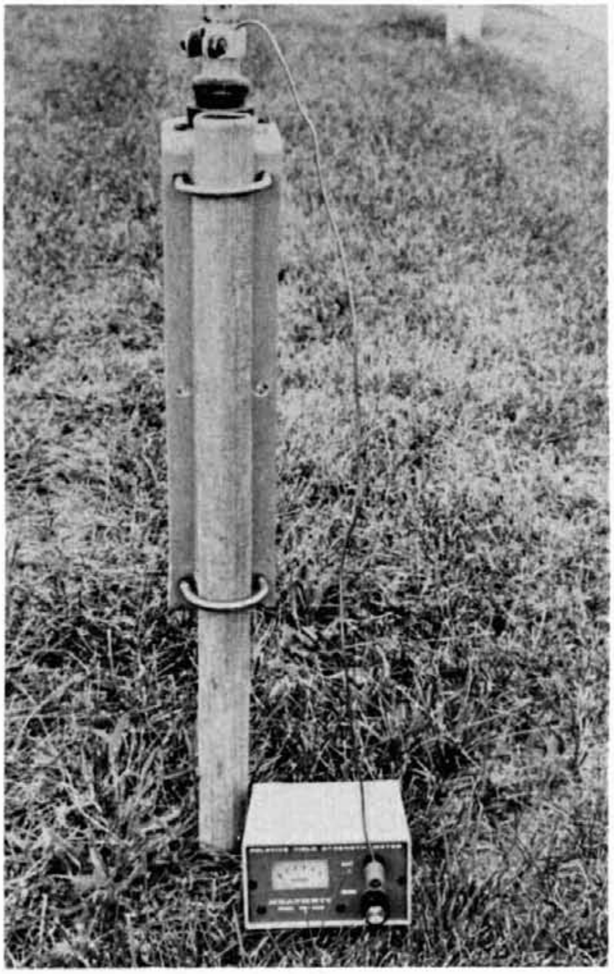
20-, 15-, and 10-meter rotatable vertical beam. It competes with the towers even though it is very close to the ground. In my opinion, the low-loss radial ground system is responsible for its good performance. Since there are no conspicuous towers, esthetically the antenna rates very high.

Many times I use the antenna as a single vertical, even on 20, 15, and 10 meters. If I hear weak DX stations, I install the director and reflector in the appropriate direction. It only takes three minutes to go from the shack to the garage, where I have a rack for the elements, out to the antenna, and back to the shack. The beam can be moved only in multiples of 45 degrees. However, this works nicely because the direction beamwidth is approximately 80 to 90 degrees.

In the near future I plan to remove the top hat section and install a 40-meter trap. I will then add another section (approximately 6 meters or 20 feet) for 80 meters. The completed antenna would be guyed at the 40-meter trap level with nylon ropes. This modification would provide greater bandwidth on 80 meters. I would also like to try traps in the director and reflector.



One of the ground mounts for the parasitic elements. The wire wound around the mount is soldered to three radials under the ground; it is attached to the base of the parasitic element with a wing nut.



Set up for making field-strength measurements using a Heath HO-1426 field-strength meter. The field-strength antenna was located 53 meters (175 feet) from the antenna site.

I was recently talking to a New Zealander on 10 meters who said, "It sounds like a tidy little system. You can take down the director and reflector and play croquet or soccer right there on the lawn." I assured him that my antenna was very versatile.

If you talk to someone who says, "Standby while I run outside to move my director and reflector," maybe you've found someone else crazy enough to try a multiband beam only 30 cm off the ground! The antenna has been fun to work with and I welcome any suggestions for improvement.

references

1. *The Radio Amateur's Handbook*, 54th edition, ARRL, Newington, Connecticut 06111, 1977.
2. John Stanley, K4ERO, "Optimum Ground System for Vertical Antennas," *QST*, December, 1976, page 13.
3. Richard Fenwick, K5RR, and R.R. Schell, "Broadband, Steerable Phased Array," *QST*, April, 1977, page 18.
4. Private correspondence from Hy-Gain Electronics Corporation, May, 1977.
5. Jerry Sevick, W2FMI, "The W2FMI 20-Meter Vertical Beam," *QST*, June, 1972, page 14.

ham radio

antenna bridge calculations

Using a hand-held programmable calculator to increase the utility of the RX noise bridge

A true picture of antenna impedance is important for maximum radio communications capability. The RX Noise Bridge is the most effective and simple measurement tool for the task.[1] Whether you use this or any other instrument, the final key to successful measurement is understanding and applying the complex impedance data.

The Smith chart is the traditional graphic tool for data reduction. It provides easy visualization of impedance and vswr as well as telling you what happens at the other end of the transmission line.[2] The modern programmable pocket calculator and basic transmission line formulas, however, will yield the same information with ease and better accuracy. Using all of these tools results in a better understanding of your antenna and transmission line.

Hewlett-Packard HP-25 calculator programs are provided for all formulas. Other types may be programmed by following the equations. A programmable calculator is recommended as a basic station tool; the HP-25 was considered the most cost effective and easiest to use.*

RX noise bridge review

The basic circuit of the RX noise bridge is shown in **fig. 1**. A noise source serves as a wideband, untuned generator, and the null detector is a frequency-accurate receiver. Balance occurs when parallel resistance and reactance of both arms are equal. Signal output is then at a minimum. A fixed capacitor at the unknown connection allows the variable capacitor to balance with both inductive or capacitive reactances.

The RX noise bridge actually measures admittance but is calibrated in terms of resistance and reactance. This is possible from duality in expressing series and parallel forms of impedance. Admittance is a parallel of conductance G and susceptance B while impedance is a series of resistance R and reactance X. Admittance is the complex inverse of impedance and vice-versa. In addition, the reciprocals of G and B can be used to express admittance as parallel resistance and reactance. The fundamental relations are

$$Z = R + jX \qquad Y = G + jB \qquad Z = \frac{1}{Y}$$

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*Nationwide discount house price for the HP-25 was $100 in early 1978.

Remember that the "j" signifies the imaginary part of the complex quantity pair. Impedance can be expressed in terms of conductance and susceptance by

$$Z = [G/(G^2 + B^2)] - j[B/(G^2 + B^2)] \qquad (1)$$

These are awkward to handle in actual values so we can use fundamental identities of $R_p = 1/G$ and $X_p = 1/B$. The "p" subscript signifies *parallel* resistance and reactance. The equation can now be expressed as

$$Z = \left[\frac{R_p X_p^2}{R_p^2 + X_p^2}\right] - j\left[\frac{R_p^2 X_p}{R_p^2 + X_p^2}\right] \qquad (2)$$

Note the sign of the imaginary terms. At first glance, this might appear that a capacitive susceptance has an inductive reactance dual and vice-versa. Not so. A capacitive susceptance and inductive reactance are positive quantities, inductive susceptance and capacitive reactance are negative.

It does not matter which way the parallel susceptance/reactance sign is used: as long as a parallel capacitor has a series capacitor dual and a parallel inductor has a series inductor dual, the noise bridge can be calibrated to your choice. The RX noise bridge by W6BXI and W6NKU uses a negative inductive susceptance calibration and this is carried through in the calculator programs which follow.

simplifying complex number operations

Here is where the scientific calcultor shines; the ability to convert from rectangular form to polar form and back again makes things easy. **Equations 1** and **2** are unwieldy in that the same values must be re-entered more than once for conversion of parallel to series. A better way is to use polar form in division and multiplication. If:

$$Z_a = Mag_a \angle Pha_a \quad and\ Z_b = Mag_b \angle Pha_b$$

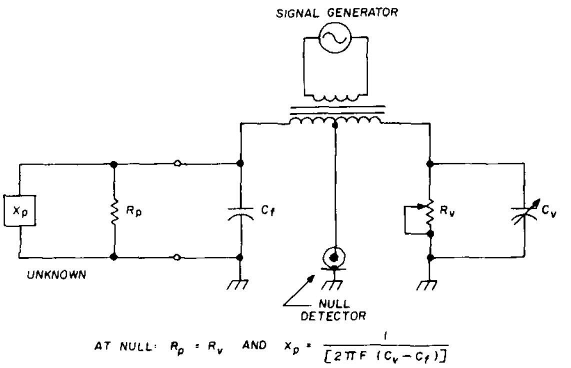

fig. 1. Basic diagram of the RX noise bridge. A complete description of this simple but useful instrument is given in reference 2.

where Mag is the magnitude and Pha is the phase angle, then

$$Z_a \cdot Z_b = (Mag_a \cdot Mag_b) \ \angle(Pha_a + Pha_b) \qquad (3)$$

$$\frac{Z_a}{Z_b} = \frac{Mag_a}{Mag_b} \ \angle(Pha_a - Pha_b) \qquad (4)$$

Another useful property to be used later is

$$\sqrt{Z_a} = \sqrt{Mag_a} \ \angle\frac{Pha_a}{2}$$

If your familiarity with rectangular and polar forms of complex numbers is a bit rusty, the rules are:

$$Z = R + jX \ or\ Mag \ \angle Pha$$

where

$Mag = \sqrt{R^2 + X^2}$	(magnitude)
$Pha = Arctangent\ (X/R)$	(phase angle)
$R = Mag \cdot Cosine(Pha)$	(real part)
$X = Mag \cdot Sine(Pha)$	(imaginary part)

A real and imaginary part is the rectangular form while a magnitude and phase angle is the polar form. They are just different ways to express the same thing. A calculator can convert forms in a single keystroke.

It may not be clear how all this will help you, so let's examine the problem of finding the parallel resistance and reactance after you have read the bridge dials. The steps are:

1. Calculate conductance and susceptance.
2. Enter G and B into the calculator in proper order.
3. Convert to polar form using the built-in function.
4. Invert the magnitude and change the angle sign.
5. Convert to rectangular form using the built-in function.
6. Read out real and imaginary part values R and X.

Measured values are entered just once. The calculator does the rest.

Still unconvinced? Suppose you get a bridge reading of 20 ohms R_p and -10 ohms for X_p at a particular frequency. Since these are parallel values, invert 20 ohms for 0.05 mho conductance and -10 ohms for -0.10 mho susceptance. Go through the calculator steps and you will get an R_s (series resistance) of 4 ohms and an X_s (series reactance) of 8 ohms at **step 6**. The s subscript is used for series values of impedance so you can keep them separated from parallel duals.

Eq. 2 will give equal results. R_s would come out as $2000/500 = 4$ and X_s would be $4000/500 = 8$. The dif-

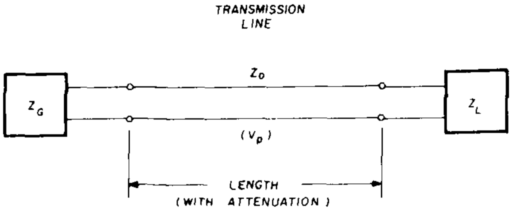

fig. 2. Basic transmission line configuration. Z_G is the impedance of the generator (transmitter); Z_L is the impedance of the load (antenna).

ference is that you have to enter R_p and X_p several times during the calculation. Try it both ways with a set of readings. Although polar/rectangular conversion is easier, you can make the programmable calculator do most of the work.

the first program

Program 1 is used to convert parallel bridge readings into series values of impedance. Steps 01 through 16 take care of this task. Steps 17 through 43 are optional and yield vswr at the point of measurement (more on vswr later).

Parallel-to-series value conversion uses the polar/rectangular conversion functions of the calculator as described before. Susceptance is calculated as $2\pi fC$. Since MHz and picofarads are assumed, the constant $2\pi \bullet 10^{-6}$ is stored in Register 6. Data entry steps are:

1. Press GTO, O, R/S (only for first set)
2. Key in frequency, press ENTER
3. Key in R_p, press ENTER
4. Key in C_p, press R/S

The first step is used only to tell the calculator where to start in the program. In this and subsequent programs, steps are arranged so that the last value will be displayed after a loop back to program start; you just enter the data for the next set and start again.

Frequency, resistance, and capacitance use the stack entry procedure at a single stop. Each time the ENTER key is pressed, the stack moves up once. The last item of a data set requires only the R/S key to begin the program; the last item is already in position after key-in.

The first stop (**step 14**) will display R_s. If you used an extender resistor, *subtract* its value at *this* stop. You would do this anyway and subtraction here will make the vswr values correct. Pressing R/S will display X_s at **step 16**. Pressing R/S once more will tell the program to proceed with vswr calculations which are displayed after looping back to the stop at **step 01**. The calculator is ready to take another stack entry for the next data set.

vswr equations

Voltage standing wave ratio or vswr is defined as[3]

$$VSWR = \frac{1 + |\rho_t|}{1 - |\rho_t|} \tag{6}$$

where

$$\rho_t = \frac{Z_t - Z_o}{Z_t + Z_o} \tag{7}$$

$\rho_t =$ complex reflection coefficient at an arbitrary point t on the line

$Z_o =$ transmission line characteristic impedance

$Z_t =$ *impedance of line at point t*

The difference between Z_o and Z_t is that Z_t has some sort of load on each end. If the load exactly equalled Z_o, then Z_t would equal Z_o and the vswr would be 1:1. Note that **eq. 6** requires the *magnitude* of ρ_t.

Transmission lines can usually be considered to be resistive-only at vhf and below. This makes $Z_o = R_o + j0$ so, with a bit of algebraic manipulation

$$VSWR = \frac{1 + \rho}{1 - \rho} \tag{8}$$

when

$$\rho = |\rho_t|$$

with

$$\rho = \sqrt{\frac{(R_t - R_o)^2 + X_t^2}{(R_t + R_o)^2 + X_t^2}} \tag{9}$$

Where R_t and X_t are the real and imaginary parts of Z_t in the usual series form.

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00	//////////	//////////					
01	74	R/S	C_p	R_p	F		STACK ENTRY
02	32	CHS	$-C_p$	R_p	F		
03	21	x↔y	R_p	$-C_p$	F		
04	15 22	g 1/x	G_p	$-C_p$	F		
05	23 03	STO 3	G_p	$-C_p$	F		
06	22	R↓	$-C_p$	F			
07	61	X	$-FC_p$				
08	24 06	RCL 6	$2\pi 10^{-6}$	$-FC_p$			
09	61	X	$-B_p$				
10	24 03	RCL 3	G_p	$-B_p$			ADMITTANCE
11	15 09	g →P	1/MAG	-PHA			POLAR CONVERT
12	15 22	g 1/x	MAG	-PHA			
13	14 09	f →R	R_s	X_s			SERIES VALUES
14	74	R/S	R_s	X_s			MANUAL SUBTRACT
15	21	x↔y	X_s	R_s			OF EXTENDER-R
16	74	R/S	X_s	R_s			
17	23 01	STO 1	X_s	R_s			BEGIN VSWR
18	23 61 01	STO X1	X_s	R_s			
19	21	x↔y	R_s	X_s			
20	23 00	STO 0	R_s	X_s			
21	14 05	RCL 5	R_o	R_s	X_s		
22	23 51 00	STO +0	R_o	R_s	X_s		
23	21	x↔y	R_s	R_o	X_s		
24	41	-	R_{SM}	X_s			
25	31	ENTER↑	R_{SM}	R_{SM}	X_s		
26	61	X	R_{SM}^2	X_s			
27	24 01	RCL 1	X_s^2	R_{SM}^2			
28	51	+	M_N^2				NUMERATOR^2
29	24 01	RCL 1	X_s^2	M_N^2			
30	24 00	RCL 0	R_{sp}	X_s^2	M_N^2		
31	31	ENTER↑	R_{sp}	R_{sp}	X_s^2	M_N^2	
32	61	X	R_{sp}^2	X_s^2	M_N^2		
33	51	+	M_D^2	M_N^2			DENOMINATOR^2
34	71	÷	ρ^2				
35	14 02	f $\sqrt{x}$	ρ				
36	23 02	STO 2	ρ				
37	01	1	1	ρ			
38	51	+	$1 + \rho$				
39	01	1	1	$1 + \rho$			
40	24 02	RCL 2	ρ	1	$1 + \rho$		
41	41	-	$1 - \rho$	$1 + \rho$			
42	71	÷	VSWR				
43	13 01	GTO 01	VSWR				DISPLAY VSWR
44							AT STOP OF
45							STEP 01
46							
47							
48							
49							

MEMORY REGISTERS
R0 R_s
$R_o + R_s = R_{sp}$
R1 X_s
X_s^2
R2 ρ
R3 G_p
R4
R5 R_o (CONSTANT)
R6 $2\pi\ 10^{-6}$ (CONSTANT)
R7

Program 1. Converting noise bridge readings to impedance and vswr.

Program 1 uses **eqs. 8** and **9** in **steps 17** through **42** and requires the value of R_o (characteristic impedance) stored in Register 5 as a constant. The program also uses R_s in place of R_t, X_s in place of X_t since the vswr point is no longer arbitrary.

Some of the program steps may appear confusing due to use of register arithmetic functions. This is used to accumulate X_s^2 in Register 1 and the value $R_s + R_o$ in Register 0. If needed, consult the *HP-25 User's Manual* for these functions.

A separate program for vswr could be made by using only **steps 16** through **43**. **Step 16** would become **step 01**, **step 43** would become **28**, and no step commands would be changed. Stack entry would require only R_s and X_s.

rotation equations

These equations are used to calculate an unknown impedance at the opposite end of a transmission line. The word "rotation" comes from Smith chart usage where the measurement point is rotated around the chart by a specified wavelength fraction. The wavelength is dependent on frequency, physical length, and velocity of propagation of the line.

The basic equations depend on which way you are looking along the line. **Fig. 2** shows the general case where Z_L is the load, Z_o is the line's characteristic impedance, and Z_G is the impedance seen at the generator end due to mismatch between Z_L and Z_o. Best power transfer occurs when the transmitter or generator has a source impedance equal to Z_G. From reference 3 the equations are

$$Z_G = \left[\frac{Z_L + Z_o Z_A}{Z_o + Z_L Z_A}\right] Z_o \qquad (10)$$

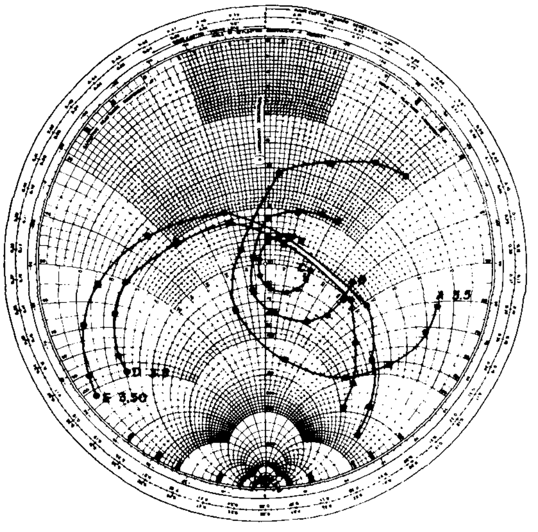

fig. 3. Impedance plots on a 50-ohm Smith chart. Each of the curves is discussed in the text.

$$Z_L = \left[\frac{Z_G - Z_o Z_A}{Z_o - Z_G Z_A}\right] Z_o \qquad (11)$$

Eq. 10 is the same as clockwise Smith chart rotation and marked there as "wavelengths toward generator." **Eq. 11** is counter-clockwise rotation and marked "wavelengths towards load" on the Smith chart. Notice that a new complex value of Z_A has been added. It has the value of $tanh(a+jb)$ where a = attenuation in nepers or decibels/8.686, and b is line wavelength expressed as an angle.

If you have followed this far, you are probably a bit confused. Either equation is enough to make a professional engineer seek computer help. Now you can understand why Phillip H. Smith invented his famous chart in the 1930s when computers were non-existent. Don't be afraid of the math because you now have calculator help.

The complex hyperbolic tangent of Z_A can be broken down to manageable terms by using some identities:

$$tanh(a+jb) = \frac{sinh(2a) + j sin(2b)}{cosh(2a) + cos(2b)}$$

Since

$$sinh(2a) = \frac{e^{4a} - 1}{2e^{2a}} \text{ and } cosh(2a) = \frac{e^{4a} + 1}{2e^{2a}}$$

Then

$$Z_A = \left[\frac{e^{4a} - 1}{D}\right] + j\left[\frac{2e^{2a} sin(2b)}{D}\right] \qquad (12)$$

where

$D = e^{4a} + 1 + 2e^{2a}cos(2b)$

$e = 2.718282\ldots$ *(base of natural logarithms)*

Z_A has now become manageable and has a real part R_A and an imaginary part X_A. Each part can be worked on separately and can use built-in calculator functions.

Many calculations can use rotations with lossless lines so let's find out what happens when attenuation, a, becomes zero. Recall that anything raised to a zero power becomes one. Under that condition, the R_A value becomes zero via its numerator and $X_A = tangent(b)$.

Before going further, **eqs. 10** and **11** have to be in a more manageable form. We can use the fact that Z_o is purely resistive at lower frequencies so

$$Z_G = \left[\frac{(R_L + R_A R_o) + j(X_L + R_o X_A)}{(R_o + R_L R_A - X_L X_A) + j(R_A X_L + R_L X_A)}\right] R_o \qquad (13)$$

$$Z_L = \left[\frac{(R_G - R_A R_o) + j(X_G - R_o X_A)}{(R_o - R_G R_A + X_G X_A) - j(R_A X_G + R_G X_A)}\right] R_o \qquad (14)$$

HP-25 Program

SWITCH TO PRGM MODE. PRESS [f] [PRGM]. THEN KEY IN THE PROGRAM

LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	74	R/S	X_G	R_G	F		STACK ENTRY
02	23 01	STO 1	X_G	R_G	F		
03	23 02	STO 2	X_G	R_G	F		
04	22	R↓	R_G	F			
05	23 00	STO 0	R_G	F			
06	23 03	STO 3	R_G	F			
07	22	R↓	F				
08	24 04	RCL 4	K	F			
09	61	X	KF				
10	14 06	f TAN	X_K				
11	23 61 02	STO X2	X_K				PARTIAL PRODUCT
12	23 61 03	STO X3	X_K				PARTIAL PRODUCT
13	24 05	RCL 5	R_o	X_K			
14	61	X	R_oX_K				
15	24 01	RCL 1	X_G	R_oX_K			
16	21	x↔y	R_oX_K	X_G			
17	41	–	IM_N				NUMBER IMAG PART
18	24 03	RCL 3	$-IM_D$	IM_N			BEGIN DENOM.
19	24 05	RCL 5	R_o	$-IM_D$	IM_N		
20	24 02	RCL 2	X_GX_K	R_o	$-IM_D$	IM_N	
21	51	+	RE_D	$-IM_D$	IM_N		
22	15 09	g →P	M_D	$-P_D$	IM_N		POLAR CONVERT
23	23 02	STO 2	M_D	$-P_D$	IM_N		
24	22	R↓	$-P_D$	IM_N			
25	23 03	STO 3	$-P_D$	IM_N			FINISH DENOM.
26	22	R↓	IM_N				
27	24 00	RCL 0	RE_N	IM_N			NUMERATOR
28	15 09	g →P	M_N	P_N			
29	21	x↔y	P_N	M_N			
30	24 03	RCL 3	$-P_D$	P_N	M_N		BEGIN DIVIDE
31	51	+	PHA	M_N			
32	21	x↔y	M_N	PHA			
33	24 02	RCL 5	R_o	M_N	PHA		RESTORE FROM
34	61	X	M_N'	PHA			*NORMALIZED FORM*
35	24 05	RCL 2	M_D	M_N'	PHA		
36	71	÷	MAG	PHA			FINISH DIVIDE
37	14 09	f →R	R_L	X_L			RECT. CONVERT
38	74	R/S	R_L	X_L			DISPLAY REAL
39	21	x↔y	X_L	R_L			
40	13 01	GTO 01	X_L	R_L			DISPLAY IMAG.
41							PART AT STOP
42							OF STEP 01
43							
44							
45							
46							
47							
48							
49							

MEMORY REGISTERS
R0 R_G
R1 X_G
R2 X_G, X_GX_K, M_D
R3 R_G, R_GX_K, $-P_D$
R4 K (CONSTANT) SEE TEXT
R5 R_o (CONSTANT)
R6
R7

Program 2. Lossless line impedance point rotation due to line length.

These may look worse, but they can now be mechanized on the calculator. The numerators and denominators are complex numbers but a program can use the simpler polar division operation for the answer. Term grouping and the signs suggest that a calculator program can be used for both expressions with only minor changes.

lossless line rotations

This should be limited to low frequencies or short line lengths since attenuation is considered zero. It has been established that only X_A is finite at zero loss and is simply the tangent of line wavelength expressed as an angle. Using degrees and one wavelength equal to 360 degrees, we can set up a constant

$$K = 0.366013 \frac{L_{ft}}{V_p} = 1.20083 \frac{L_M}{V_p} \qquad (15)$$

where

L_{ft} = line length in feet
L_M = line length in meters
V_p = velocity of propagation of the line

Velocity of propagation is usually the reciprocal of the square-root of the dielectric constant in coax. It is different for twin-lead or open-wire line. These can be found in handbooks or transmission line tables.[4]

Wavelength is frequency sensitive. Angular wavelength uses **eq. 15** multiplied by the frequency in MHz. A lossless line situation is a special case so we can use $X_A' = X_K = tangent(Kf)$ with f as the frequency. The lossless line rotation equations are simplified

$$Z_{GL} = \left[\frac{R_L + j(R_oX_K + X_L)}{(R_o - X_LX_K) + jR_LX_K} \right] R_o \qquad (16)$$

$$Z_{LL} = \left[\frac{R_G + j(X_G - R_oX_K)}{(R_o + X_GX_K) - jR_GX_K} \right] R_o \qquad (17)$$

the second program

This was written for **eq. 17**. Program start and stack entry procedures are the same as Program 1. Data set entry is frequency in MHz, R_G, and X_G, in that order. The ENTER key is not pressed after keying in X_G since it is the last item; simply press R/S to start.

Constant K from **eq. 15** must be preloaded into Register 4. R_o must also be preloaded into Register 5. The first stop at **step 38** will display R_L; X_L is displayed after loop-back to the stop at the first step. Another stack entry may be done after reading X_L.

Programming is straightforward but **step 31** might seem to violate polar division rules in that the angles are added instead of subtracted. **Step 12** accumulates the denominator imaginary part (R_GX_K) as a positive quantity in Register 3. After recall at **step 18** and conversion to polar form at **step 22**, the denominator phase angle sign is changed. Rather than using an extra CHS command, **step 31** is an ADD (subtraction of a negative number is the same as addition of a positive number).

Steps 33 and **34** multiply the numerator magnitude by R_o to achieve the same purpose as the right bracket multiply in **eq. 17**. Deleting these two steps would yield normalized impedance values for the result although the data input must be in conventional un-normalized form. Program 2 can be written to accept a normalized impedance input if the equation is re-arranged.

Program 2 can be easily modified to solve **eq. 16**; just change **step 17** to ADD, **steps 21** and **31** to SUB-

table 1. Calculated values obtained from original bridge readings.

f, MHz	R_p	C_p	Z_G	vswr	Z_L
3.50	149*	-163	15.93+j 61.92	8.15	30.57-j 95.96
3.55	164*	-163	20.99+j 72.14	7.63	26.95-j 85.04
3.60	202*	-141	42.75+j 91.97	5.81	24.29-j 64.62
3.65	240*	-98	85.91+j100.28	4.41	24.11-j 50.14
3.70	129	-66	124.14+j 24.57	2.60	28.59-j 30.76
3.75	78	152	72.35-j 20.21	1.64	32.49-j 10.05
3.80	144*	20	43.32-j 9.86	1.29	41.17+j 7.54
3.85	121*	-9	20.92+j 3.19	2.40	48.68+j 44.61
3.90	118*	-39	16.52+j 13.14	3.26	64.52+j 69.58
3.95	116*	-65	12.08+j 20.97	4.91	78.74+j106.83
4.00	117*	-84	10.27+j 27.24	6.36	98.80+j141.22

*indicates extender-resistor of 100 ohms used

TRACT, and use load impedance at stack entry. This can be seen by inspecting the signs of each equation.

practical example

The noise bridge dial readings of the W6BXI/W6NKU article are used to illustrate operations with Programs 1 and 2; the results are shown in **table 1**. R_s, X_s, and vswr at the generator are from Program 1 while antenna impedance R_L and X_L were derived via Program 2. The original 60-foot (18-meter) loss-line line of 50 ohms characteristic impedance and v_p of 0.66 were used for rotation. Program 2 data entry used two-decimal values, quite adequate for normal use.

Curves **A** and **D** of the Smith chart in **fig. 3** are plotted from **table 1**. Curve **A** is the measurement (or generator) end, and curve **D** is the antenna end. There is a slight difference which is due mainly to the resolution of the chart when rotating manually. Resolution and manual errors are minimized with the calculator programs; wavelength is automatically calculated instead of being a separate operation.

Note that **fig. 3** is a 50-ohm Smith chart and not the usual normalized version. This is slightly better *if* you are using a 50-ohm transmission line. Normalized impedances are found by simply dividing real and imaginary parts by R_o.

Impedance will change rapidly around resonance. For best accuracy, several more readings could be taken and calculated where the vswr appears low. Two to five times as many readings are possible with an accurate receiver frequency calibration. The calculator programs make short work of data reduction.

table 2. Calculated values from the first example with 0.8 dB line loss and the same impedance readings at the generator end of the line.

f, MHz	Z_G	generator vswr	R_A	X_A	Z_L	load vswr
3.50	15.93 + j 61.92	8.15	0.4474	− 1.9267	8.25 − j103.21	32.02
3.55	20.99 + j 72.14	7.63	0.4015	− 1.8021	8.59 − j 90.87	25.18
3.60	42.75 + j 91.97	5.81	0.3628	− 1.6889	12.03 − j 69.13	12.26
3.65	85.91 + j100.28	4.41	0.3299	− 1.5855	15.23 − j 53.94	7.27
3.70	124.14 + j 24.57	2.60	0.3018	− 1.4907	23.30 − j 33.85	3.29
3.75	72.35 − j 20.21	1.64	0.2774	− 1.4033	29.45 − j 11.27	1.83
3.80	43.32 − j 9.86	1.29	0.2563	− 1.3225	39.44 + j 8.74	1.36
3.85	20.92 + j 3.19	2.40	0.2379	− 1.2473	43.65 + j 52.91	2.96
3.90	16.52 + j 13.14	3.26	0.2217	− 1.1771	55.37 + j 86.95	4.52
3.95	12.08 + j 20.97	4.91	0.2075	− 1.1114	55.13 + j137.54	8.76
4.00	10.27 + j 27.24	6.36	0.1949	− 1.0497	54.91 + j189.95	15.08

operations with lossy lines

Fig. 4 shows the attenuation in dB per 100 feet (dB per 30.48 meters) of common coaxial lines. These are nominal since there is slight variation from one production run to another, and between manufacturers. You can see that attenuation becomes more pronounced at higher frequencies.

Complex attenuation/phase factor Z_A must be precalculated to use **eqs. 13** and **14** on the HP-25. This is a simple task when Program 3 is used and the output data can be tabulated with four-decimal accuracy. This data will be used with Program 4 following.

Program 3 takes advantage of the fact that attenuation variations are small over a narrow frequency band. Attenuation is treated as a constant and expressed as nepers (dB/8.686) stored in Register 7 at twice its value. Wavelength constant K is preloaded

table 3. Calculation of changes due to introducing attenuation pads between the noise bridge and the line in the first example.

	no pad		3 dB pad		6 dB pad	
f, MHz	Z_G	vswr	Z_G	vswr	Z_G	vswr
3.50	15.93 + j 61.92	8.15	44.15 + j39.56	2.29	51.09 + j20.20	1.49
3.55	20.99 + j 72.14	7.63	50.67 + j41.99	2.25	54.72 + j20.03	1.48
3.60	42.75 + j 91.97	5.81	65.48 + j40.45	2.10	60.37 + j16.88	1.43
3.65	85.91 + j100.28	4.41	78.26 + j30.64	1.92	63.97 + j11.59	1.38
3.70	124.14 + j 24.57	2.60	77.71 + j 6.51	1.57	62.27 + j 2.52	1.36
3.75	72.35 − j 20.21	1.64	60.82 − j 8.13	1.28	55.28 − j 3.66	1.13
3.80	43.32 − j 9.86	1.29	46.81 − j 5.27	1.14	48.45 − j 2.73	1.07
3.85	20.92 + j 3.19	2.40	32.99 + j 2.19	1.52	40.70 + j 1.31	1.23
3.90	16.52 + j 13.14	3.26	30.60 + j 9.43	1.72	39.54 + j 5.78	1.31
3.95	12.08 + j 20.97	4.91	28.59 + j15.72	1.99	38.85 + j 9.80	1.40
4.00	10.27 + j 27.24	6.36	28.66 + j20.64	2.15	39.51 + j12.79	1.49

into Register 6 as twice its value also. Doubling of the constants saves program steps.

The only stack entry is frequency in MHz at **step 01**. Real part R_A is displayed at **step 23** and imaginary part X_A displayed after loop-back to the first step. Just key in the next frequency and press R/S for the next Z_A set.

Program 3 follows **eq. 12** but may be confusing due to extensive use of register arithmetic functions. It is a fairly good example of program optimization and is worth study just for that reason.

Change the attenuation constant if tabulating Z_A for more than one band. Attenuation is proportional to physical length and can be found easily from **fig. 4**.

rotations with lossy lines

Program 4 is written to mechanize **eq. 14**. There are four items in stack entry: R_G, X_G, R_A, and X_A, in that order. Remember that only the R/S key is pressed after keying in X_A. The first display is R_L and another stack entry can be done after display of X_L. Constant R_o must be preloaded in Register 5.

We can use the previous example to show attenuation effects. Assume the same bridge readings with the same line length and v_p. Now add a total attenuation of 0.8 dB to the line, calculate Z_A with Program 3, and calculate the new antenna impedance with Program 4. The data is given in **table 2** and the new impedance is curve **E** of **fig. 2**. Note that **E** is more reactive than curve **D**.

If the antenna impedance was actually curve **D**, then the measurements would show another curve at the generator end that is closer to the center of the Smith chart. This would be a "masking" effect on impedance due to attenuation.

Program 4 can be altered to find the generator impedance from **eq. 13** by simply changing **steps 25**, **32**, and **35** to ADD, **step 41** to SUBTRACT. Stack entry would use R_L in place of R_G, and X_L in place of X_G: R_A and X_A would remain the same.

The Z_A tabulation used a constant of $(0.8/4.343) = 0.184204$ for a in Register 7 of Program 3; the $2K$ constant was $2(0.366013 \times 60/0.66) = 66.5478$ in Register 6. Six-place constants were used for accuracy. Output data for Z_A can be to four places.

impedance masking by attenuation

This has already been stated but another example is in order. Let's take the first example again and use only the measurement end data. Now assume that short pads of 3 dB and 6 dB were inserted between the noise bridge and line. To calculate this condition, Program 3 used a zero line length. X_A became zero and R_A was frequency-insensitive with a value of *0.332275* for 3 dB, and *0.598474* for 6 dB. Program 4 was modified to fit **eq. 13** and results are shown in **table 3** and **fig. 2** as curve **B** (3 dB) and curve **C** (6 dB).

The original 8.15:1 vswr at 3.5 MHz dropped to 2.29:1 at 3 dB and only 1.49:1 at 6 dB. This not only soaks up power but could fool a swr meter installed at the transmitter. Suppose you had a 10-meter rig with a total of 60 meters (200 feet) of RG-58/U coax feeding the antenna. Total line attenuation from **fig. 4** indicates a line loss of 4.5 dB. An antenna-end mismatch giving an 8:1 vswr would show up as less than

table 4. Solution of attenuation and test angle or unknown transmission line by open- and short-circuiting the load end of the line.

	shorted load end			open end load			Z_u solution	
f, MHz	R_p	C_p	Z_G	R_p	C_p	Z_G	dB	ϕ, degrees
21.0	190	125	17.56 + j 55.03	310	−75	29.78 − j91.34	1.3315	37.4396
21.1	220	110	19.48 + j 62.50	280	−85	25.56 − j80.64	1.3319	41.1636
21.2	250	100	20.68 + j 68.86	250	−100	20.68 − j68.86	1.2854	45.0000
21.3	280	85	25.12 + j 80.02	220	−110	19.15 − j62.02	1.3197	48.8323
21.4	310	75	28.78 + j 89.96	190	−120	18.27 − j56.02	1.3396	52.0243
21.5	340	65	34.30 + j102.40	160	−140	15.75 − j47.67	1.3150	56.2338

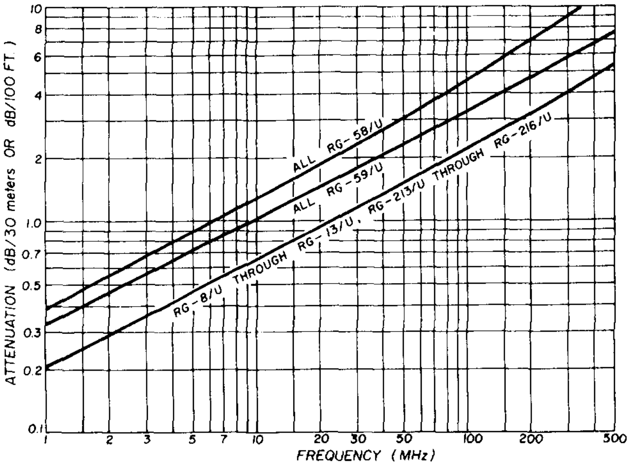

fig. 4. Attenuation versus frequency for common coaxial transmission lines.

2:1 at the transmitter end. Only a calibrated noise bridge and careful calculations tell the truth.

Larger coax for less loss is best at higher frequencies. It also improves antenna measurements since there is less "masking" of impedance; the impedance points move closer to the center of the Smith chart with increased line attenuation.

Impedance masking by attenuator pads is an old technique used in microwave bench measurements to improve the "flatness" of the source impedance; it's also a good idea at lower frequencies. Less expensive generators vary widely so a pad would make the device under test "see" a better impedance.

Comparison of the plot points of curves **A**, **B**, and **C** in **fig. 3** will show no change in rotation position. The same is true when comparing curves **D** and **E**. Attenuation changes only the chart position away from center even though the tabulated impedance seems to change rotation. Rotation is strictly a wavelength function and is due to frequency, line length, and velocity of propagation.

transmission line quality

It is common to run across some really juicy bargains in coaxial cable at surplus houses; it is just as common to find out there is no known identification and the cable probably has different characteristics than advertised. Is it good or bad? What is the characteristic impedance? Velocity of propagation? All are unknown. On the other hand, perhaps your four-year-old feedline is going bad due to weather.

Fortunately there is a way to find out, and fairly simply with an accurate RX noise bridge and another program. Just check out the line itself with a short and an open at the other end. This sounds too simple so let's investigate the theory.

Call the measured impedance with an open circuit Z_{oc} and the measured impedance with a short circuit Z_{sc}. From reference 3 we know that

$$Z_o = \sqrt{Z_{oc}\,Z_{sc}} \qquad (18)$$

This calculation is easily done manually in polar form using **eq. 5**. Do this over several frequencies and average the results. Averaging will dilute errors in calibration and dial readings.

The next problem is to find attenuation and the velocity of propagation. You can take a guess at v_p just from examining the dielectric. The most common type is polyethylene ($v_p = 0.6594$); other solid dielectrics are close to this value. Foamed dielectrics will vary a great deal. The v_p guess is best *since the line* wavelength should be an *odd* multiple of 1/8th wavelength.

Choice of length is better explained by calling attention to the bridge measurement limit contour. The shorted-end measurements will be directly opposite the chart from the open-end measurements. The RX noise bridge cannot measure impedances falling at the high-resistance end. The first step is to calculate

$$Z_u \text{ (polar form)} = \frac{1 + \sqrt{Z_{sc}/Z_{oc}}}{1 - \sqrt{Z_{sc}/Z_{oc}}} \qquad (19)$$

From this we can find

$$dB\ attenuation = 4.343\ Log_e\ (Z_u\ magnitude) \qquad (20)$$

$$\theta,\ the\ test\ angle = \frac{(Z_u\ phase\ angle)}{2} \qquad (21)$$

Eq. 20 uses natural logs, not the base 10 or common logarithm. We are stuck with a "test angle" since wavelength is not precisely known. Velocity of propagation is found from:

$$v_p = \frac{0.366013\,f\,L_{ft}}{\theta + 360n} = \frac{1.20083\,f\,L_M}{\theta + 360n} \qquad (22)$$

where

f = measurement frequency in MHz
L_{ft} = physical line length in feet
L_M = physical line length in meters
n = any integer number including zero

We can pin down the value of v_p by using **eq. 22** with several values of n at each frequency. **Eqs. 21** and **22** assume the test angle to be in degrees and only one n value will be correct; this can be found by inspection.

Program 5 was written to mechanize **eq. 19**. Stack entry requires four values; these are obtained from Program 1 — ignore the vswr data in this case. The first display stop of Program 5 is the attenuation in dB. The next display is test angle θ after looping back to the first program step.

example

You have a 20 meter (65 foot) length of coaxial cable that appears to be RG-58/U but the markings are unknown. The dielectric appears to polyethylene so you can estimate wavelength using **eq. 15**. K will be *36.0759* so it seems that multiplying this by 15-meter frequencies will be correct. At 21.0 MHz you get 757.670° (2.104 wavelengths) and 21.5 MHz gives 775.710° (2.155 wavelengths). Measurements are taken every 100 kHz.

table 5. Velocity of propagation test at different values of n.

	v_p at different n values				
f, MHz	**n = 0**	**n = 1**	**n = 2**	**n = 3**	**n = 4**
21.0	13.3439	1.2571	0.6596	0.4471	0.3382
21.1	12.1949	1.2513	0.6595	0.4477	0.3389
21.2	11.2081	1.2453	0.6593	0.4483	0.3396
21.3	10.2081	1.2395	0.6591	0.4489	0.3404
21.4	9.7863	1.2357	0.6595	0.4497	0.3412
21.5	9.0960	1.2289	0.6590	0.4502	0.3419

The results of bridge measurements and data from Programs 1 and 5 are given in **table 4**. Note: This particular noise bridge had a 500-ohm potentiometer instead of the original 250-ohm pot.

Characteristic impedance is found by manual calculation with **eq. 18** and is

$$74.49 \angle 0.1800^\circ \text{ at } 21.0 \text{ MHz}$$
$$74.42 \angle 0.1378^\circ \text{ at } 21.1 \text{ MHz}$$
$$71.90 \angle 0.0000^\circ \text{ at } 21.2 \text{ MHz}$$
$$73.78 \angle -0.1345^\circ \text{ at } 21.3 \text{ MHz}$$
$$74.60 \angle 0.1612^\circ \text{ at } 21.4 \text{ MHz}$$
$$73.63 \angle -0.1177^\circ \text{ at } 21.5 \text{ MHz}$$

The average Z_o is 73.80 ohms over six frequencies. The small angle residue comes from minor calibration and measurement errors can be ignored. If it is over 5° you have done something wrong or the bridge limits have been exceeded.

Average attenuation is found to be 1.3205 dB. Trials of n values for v_p are given in **table 5**. This table shows that low n values have a steadily decreasing v_p with increasing frequency while high n values have steadily increasing v_p. The correct n value has v_p bouncing around some average value. Data at $n \approx 2$ is steadiest and the average is *0.6593*. The guess at polyethylene dielectric was correct.

This cable example was actually RG-59/U and a manufacturer's reject sold as surplus. A check using an HP 8507A Automatic Network Analyzer over many more frequencies showed that the cable Z_o was actually 73.0 ohms and attenuation measured 1.30 dB. At only six frequencies the noise bridge was 1.1 per cent high for Z_o and 1.6 per cent high for attenuation, a very good score for equipment costing one-thousandth of laboratory instruments!

jumping over the measurement contour

The noise bridge is limited at high R_s or large X_s values. A novel suggestion by Dean Straw, N6BV, is to use a fixed resistor in *shunt* with the potentiometer.[5] This could be plugged into the TUNING arm of the bridge and still have the option of using the extender resistor on the UNKNOWN arm. A problem is that calibration must be done *very* carefully.

N6BV cites the case of a marine whip antenna having an impedance of 10-j400 ohms at 2 MHz. Even with the extender resistor, a null is not possible. The R_p value would have to be 1565 ohms. Placing a 220-ohm fixed resistor across the potentiometer allows a null at an R_p of 193 ohms. C_p would be −185 pF and at the capacitor limit.

The need for accurate calibration can be seen by comparing the impedance derived from measurements versus required values. A dial reading of 193 ohms R_p and −185 pF C_p will calculate as 21.24−j392.72 ohms. Correct dial readings *should* have been 192.88 ohms for R_p and −184.96 pF for C_p. A very small change in R_p gives a very large change in R_s.

Another method, more accurate with calculator help, is to use an *extender line*. This is used between the noise bridge and the unknown. Length should be between 1/8th and 3/8th wavelength to bring you back inside the limit contour. The extender line can be measured using the short/open method described before, then tagged for reference.

A small pad can also be used but should not exceed about 3 dB to retain accuracy. Program 4 is then used with the pad added to line attenuation. Eighth-watt carbon composition resistors are suitable for the pad and the finished unit must be checked carefully.

using an extender line

The procedure for using an extender line is simple. Take bridge readings with the extra line between bridge and unknown. Check for good nulls; a poor null will tell you that the extra line is either too short or too long. Try an extra extender if necessary. Find the total line length.

The first step is to use Program 1 to obtain "extended" R_G and X_G values. The second step is to

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00	//////	//////					
01	74	R/S	F				ENTER FREQUENCY
02	24 06	RCL 6	2K	F			
03	61	X	2KF				
04	23 03	STO 3	2KF				
05	14 05	f COS	COS 2KF				
06	24 07	RCL 7	α	COS 2KF			
07	15 07	g e^x	e^a	COS 2KF			
08	23 00	STO 0	e^a	COS 2KF			FORM $2e^a$
09	23 51 00	STO +0	e^a	COS 2KF			IN R0
10	23 01	STO 1	e^a	COS 2KF			FORM e^{2a}
11	23 61 01	STO X1	e^a	COS 2KF			IN R1
12	31	ENTER↑	e^a	e^a	COS 2KF		
13	51	+	$2e^a$	COS 2KF			
14	61	X	$2e^a$COS				
15	24 01	RCL 1	e^{2a}	$2e^a$COS			
16	51	+	DEN.−1				
17	01	1	1	DEN.−1			
18	23 41 01	STO −1	1	DEN.−1			
19	51	+	DENOM.				
20	23 71 00	STO ÷0	DENOM.				REGISTER
21	23 71 01	STO ÷1	DENOM.				ARITHMETIC
22	24 01	RCL 1	RA				
23	74	R/S	RA				DISPLAY REAL PART
24	24 03	RCL 3	2KF				
25	14 04	f SIN	SIN 2KF				
26	24 00	RCL 0	IM/SIN	SIN 2KF			
27	61	X	X_A				
28	13 01	GTO 01	X_A				DISPLAY IMAG.
29							PART AT STOP
30							OF STEP 01
31							
32							
33							
34							
35							
36							
37							
38							
39							
40							
41							
42							
43							
44							
45							
46							
47							
48							
49							

MEMORY REGISTERS
R0 e^a
$2e^a$
IMAG/SIN 2KF
R1 $e^a \rightarrow e^{2a}$
$e^{2a}-1$
REAL
R2
R3 2KF
R4
R5
R6 2K
(CONSTANT,
SEE TEXT)
R7 α=dB/4.343
(CONSTANT)

Program 3. Complex attenuation-wavelength factor.

"look backward" on the total line to find impedance at the desired point. This can be done with either Program 2 or 4, each modified to solve for Z_L. Length for the second step is that of the extender alone, not the total line length. The data output is the impedance at the desired point.

Puzzled? The first step brought you within the measurement limit contour but the length was longer than actual; the impedance was rotated too far in a counter-clockwise direction. The second step found the correct impedance by clockwise rotation of a wavelength determined by the extender line.

Characteristic impedance of the extender line must be equal to the measured line for the two-step procedure. A solution with different impedances is possible but the math is lengthy. A direct measurement without transmission line would use the Z_o of the extender line.

An extender line can be of different dimensions as long as the equal Z_o rule is met. A word of caution with RG-58/U: Common RG-58/U and RG-58B/U have a Z_o of 53.5 ohms, while RG-58A/U is 52.0 ohms; RG-58C/U is 50.0 ohms. RG-58C/U is recommended as a general-purpose 50-ohm line since it is extra flexible and more durable for test work.

keeping tabs on that new antenna

Weather takes its toll on all antenna installations.

HP-25 Program

SWITCH TO PRGM MODE, PRESS [f] [PRGM], THEN KEY IN THE PROGRAM

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00	//////////	//////////					
01	74	R/S	X_A	R_A	X_G	R_G	STACK ENTRY
02	23 03	STO 3	X_A	R_A	X_G	R_G	
03	23 07	STO 7	X_A	R_A	X_G	R_G	
04	22	R↓	R_A	X_G	R_G	X_A	
05	23 02	STO 2	R_A	X_G	R_G	X_A	
06	23 06	STO 6	R_A	X_G	R_G	X_A	
07	22	R↓	X_G	R_G	X_A	R_A	
08	23 01	STO 1	X_G	R_G	X_A	R_A	
09	21	x↔y	R_G	X_G	X_A	R_A	
10	23 00	STO 0	R_G	X_G	X_A	R_A	
11	23 61 06	STO X 6	R_G	X_G	X_A	R_A	PARTIAL PRODUCT
12	23 61 07	STO X 7	R_G	X_G	X_A	R_A	ACCUMULATE
13	22	R↓	X_G	X_A	R_A		
14	61	X	X_GX_A	R_A			
15	23 41 06	STO − 6	X_GX_A	R_A			
16	21	x↔y	R_A	X_GX_A			BEGIN DENOM.
17	24 01	RCL 1	X_G	R_A	X_GX_A		
18	61	X	X_GR_A				
19	24 07	RCL 7	R_GX_A	X_GR_A			
20	51	+	$-IM_D$				IMAGINARY PART
21	24 05	RCL 5	R_o	$-IM_D$			
22	23 61 02	STO X 2	R_o	$-IM_D$			PARTIAL PRODUCT
23	23 61 03	STO X 3	R_o	$-IM_D$			ACCUMULATE
24	24 06	RCL 6	R6	R_o	$-IM_D$		
25	41	−	RE_D	$-IM_D$			REAL PART
26	15 09	g →P	M_D	$-P_D$			POLAR CONVERT
27	23 06	STO 6	M_D	$-P_D$			
28	21	x↔y	$-P_D$	M_D			
29	23 07	STO 7	$-P_D$	M_D			FINISH DENOM.
30	24 01	RCL 1	X_G				BEGIN NUMERATOR
31	24 03	RCL 3	R_oX_A	X_G			
32	41	−	IM_N				IMAGINARY PART
33	24 00	RCL 0	R_G	IM_N			
34	24 02	RCL 2	R_oR_A	R_G	IM_N		
35	41	−	RE_N	IM_N			REAL PART
36	15 09	g →P	M_N	P_N			POLAR CONVERT
37	24 05	RCL 5	R_o	M_N	P_N		RESTORE FROM
38	61	X	M_N'	P_N			NORMALIZED FORM
39	21	x↔y	P_N	M_N'			
40	24 07	RCL 7	$-P_D$	P_N	M_N'		
41	51	+	PHA	M_N'			
42	21	x↔y	M_N'	PHA			
43	24 06	RCL 6	M_D	M_N'	PHA		
44	71	÷	MAG	PHA			
45	14 09	f →R	R_L	X_L			RECT. CONVERT
46	74	R/S	R_L	X_L			DISPLAY REAL PART
47	21	x↔y	X_L	R_L			
48	13 01	GTO 01	X_L	R_L			DISPLAY IMAG PART
49							AT STOP OF STEP 01

MEMORY REGISTERS	
R0	R_G
R1	X_G
R2	R_A
	R_oR_A
R3	X_A
	R_oX_A
R4	
R5	R_o
	(CONSTANT)
R6	R_A →R_GR_A
	$R_GR_A - X_GX_A$
	M_D
R7	X_A→R_GR_A
	$-P_D$

Program 4. Impedance point rotation on lossy lines due to line length.

A good way to locate potential problems is to periodically measure the line and antenna. Take readings in both good and bad weather; snow, ice, or rain can cause impedance changes on good antennas.

You might be able to catch a line break before it happens. A poorly sealed coax can cause erosion of the center conductor, changing both the impedance and attenuation at one end. Direct cable feeds to rotating antennas can bend and stretch the coax out of shape.

The near-field or Fresnel Zone (about five wavelengths from center) can include trees and poles. This may cause impedance changes with rotary antennas. Try measurements at different azimuths. Impedance changes will warn you to expect different transmitter loading at certain antenna directions.

Fixed antennas are not immune to change. Trees grow slowly and old ones may be cut down. House remodelling can cause changes, too. Replacement of rain gutters can change the local ground of a rooftop mount. In one observed change, a 40-meter vertical mounted on a back wall was changed by adding an aluminum patio roof. The pattern was also changed, but the noise bridge can't measure that.

the forgotten receiver

Receiver agc can fool the best of us even if the input impedance is grossly mismatched (many receivers are faulty in this respect). A good communications link needs best performance in *both* directions so it's a good idea to check your receiver, too.

If you have one accurate-frequency receiver and it was used as the bridge detector, you have to substitute. A general-purpose communications receiver is good enough and you can calibrate it using received signals from local stations.

Most receiver designs use preselector tuning of the front end. Don't forget to peak the preselector; noise from the noise bridge generator is good for this. An off-peak preselector will reflect a different impedance. A known input impedance and thorough schematic study will be enough to tell you what to do for a best match.

cautions with high vswr

High vswr may be unavoidable, but peak voltages and currents must be considered in relation to the line and matching network. The following relations help

$$E_{peak} = E_{nominal} \sqrt{2\ VSWR} \qquad (23)$$

$$I_{peak} = I_{nominal} \sqrt{2\ VSWR} \qquad (24)$$

Nominal values are rms at a perfect match.

HP-25 Program

SWITCH TO PRGM MODE. PRESS [f] [PRGM]. THEN KEY IN THE PROGRAM.

LINE	CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00							
01	74	R/S	X_{oc}	R_{oc}	X_{sc}	R_{sc}	STACK ENTRY
02	21	x↔y	R_{oc}	X_{oc}	X_{sc}	R_{sc}	
03	15 09	g →P	M_{oc}	P_{oc}	X_{sc}	R_{sc}	POLAR DENOM.
04	15 22	g 1/x	$1/M_{oc}$	P_{oc}	X_{sc}	R_{sc}	
05	23 00	STO 0	$1/M_{oc}$	P_{oc}	X_{sc}	R_{sc}	
06	22	R↓	P_{oc}	X_{sc}	R_{sc}		
07	32	CHS	$-P_{oc}$	X_{sc}	R_{sc}		
08	23 01	STO 1	$-P_{oc}$	X_{sc}	R_{sc}		
09	22	R↓	X_{sc}	R_{sc}			
10	21	x↔y	R_{sc}	X_{sc}			
11	15 09	g →P	M_{sc}	P_{sc}			POLAR NUMERATOR
12	23 61 00	STO X 0	M_{sc}	P_{sc}			POLAR
13	21	x↔y	P_{sc}	M_{sc}			MULTIPLY IN
14	23 51 01	STO + 1	P_{sc}	M_{sc}			R_0 & R_1
15	24 01	RCL 1	2P				
16	02	2	2	2P			
17	71	÷	P				
18	24 00	RCL 0	M^2	P			
19	14 02	f $\sqrt{x}$	M	P			$(Z_{sc}/Z_{oc})^{1/2}$ TERM
20	14 09	f →R	RE	IM			RECT. CONVERT
21	23 02	STO 2	RE	IM			
22	01	1	1	RE	IM		
23	51	+	RE + 1	IM			
24	21	x↔y	IM	RE + 1			
25	23 03	STO 3	IM	RE + 1			
26	21	x↔y	RE + 1	IM			
27	15 09	g →P	M_N	P_N			POLAR NUMER.
28	23 00	STO 0	M_N	P_N			
29	21	x↔y	P_N	M_N			
30	23 01	STO 1					
31	24 03	RCL 3	IM				
32	01	1	1	IM			
33	24 02	RCL 2	RE	1	IM		
34	41	−	1-RE	IM			
35	15 09	g →P	M_D	$-P_D$			POLAR DENOM.
36	23 71 00	STO ÷ 0	M_D	$-P_D$			POLAR
37	21	x↔y	$-P_D$	M_D			DIVIDE IN
38	23 51 01	STO + 1	$-P_D$	M_D			R_0 & R_1
39	24 00	RCL 0	M_U				Z_U MAGNITUDE
40	14 07	f LN	LN(M_U)				
41	24 04	RCL 4	4.343	LN(M_U)			
42	61	X	dB				
43	74	R/S	dB				ATTENUATION
44	24 01	RCL 1	P_U				ZU PHASE ANGLE
45	02	2	2	P_U			
46	71	÷	ø				TEST ANGLE
47	13 01	GTO 01	ø				DISPLAY ø
48							AT STOP OF
49							STEP 01

MEMORY REGISTERS	
R0	$1/M_{oc}$ / M^2 / M_N M_U
R1	$-P_{oc}$ / 2P / P_N P_U
R2	RE
R3	IM
R4	4.343 (CONSTANT)
R5	
R6	
R7	

Program 5. Calculating properties of unknown transmission line from open- and short-circuited measurements.

Assume perfect conditions with a 400-watt transmitter. Line voltage with a 50-ohm load is 141.42 rms or 200 volts peak. Line current is 2.8284 amps rms or 4 amps peak. Now change the load so that the vswr is 4:1.

At a certain line wavelength, the transmitter may see an impedance of 200 + j0 ohms. At maximum real part and minimum imaginary part of impedance, **eq. 23** will apply. Peak voltage is 400 or twice the perfect condition. Changing the line length by exactly a quarter wavelength will change the load to 12.5 + j0 ohms. **Eq. 24** indicates a peak current of 8 amperes. Peak values increase by the square-root of vswr at both lengths. *Both* the transmission line and matching network must handle these peaks or breakdown occurs.

Other line wavelengths will present different loads to the transmitter. This can be seen by following the vswr circle on a Smith chart. A long transmission line can have both peaks; voltage breakdown rating is the prime consideration. Power handling capability is next. It is a good idea to check both from cable tables.[4,6]

If at all possible, try to match *at the antenna*. The reason is peak line current. Center conductor heating in coax will increase attenuation in peak current regions; heating increases conductor resistance for more loss. It is worst in warm weather. Balanced lines have a similar condition but it's less pronounced — impedance is generally higher and the current is divided equally in each wire.

High power impedance *may* be different at high vswr than that measured with low power bridge readings.

baluns and balanced lines

A balun transformer allows you to check twinlead and open-wire lead with the RX noise bridge. However, the balun itself must be carefully checked since it becomes part of the bridge. Checks and calibration are done in the same manner as original bridge calibration.

Impedance-change baluns will affect bridge readings. Both real and imaginary series values must be multiplied by the impedance-change value. All other programs use the balanced line characteristic impedance.

Remember to keep away from the field around balanced lines. Measurement errors can be made if you are too close to the balun attachment point. You can also use Program 5 with balanced lines. A suggestion is to suspend the line to be measured by string from a rope. A 1-meter (three foot) distance above ground should be sufficient.

other uses

The RX noise bridge lends itself to any frequency in the high-frequency band. It is well within the FCC rules for incidental radiation. N6BV cited its use with a marine whip; it can also be used to check an SWL's long-wire.

Your CB neighbor might even be won over to amateur radio. Offer to check out the CB antenna and give suggestions if the readings aren't good.* The double handful of bridge and calculator, plus scratchpad and this issue of *ham radio* (for the programs, of course) can be impressive. Just be prepared for a lot of questions if friendly neighbor starts reading the magazine!

*Tape the transmit switch to receive-only or you can damage the bridge and don't forget you need a commercial ticket to adjust the CB transmitter.

references

1. Robert A. Hubbs, W6BXI, and A. Frank Doting, W6NKU, "Improvements to the RX Noise Bridge," *ham radio*, February, 1977, page 10.
2. James Fisk, W1DTY, "How to Use the Smith Chart," *ham radio*, November, 1970, page 16 (reprinted March, 1978, page 92).
3. Robert A. Chipman, *Theory and Problems of Transmission Lines*, Schaum's Outline Series, McGraw-Hill, New York, 1968 (still sold at large paperback stores and a bargain at $5.95. Math heavy but gives all the answers. Recommended for serious study).
4. Doug DeMaw, W1CER, *ARRL Electronics Data Book*, The American Radio Relay League, Newington, Connecticut, 1976.
5. R. Dean Straw, N6BV, letter to Charles Carroll, K1XX, June 1, 1977.
6. *RF Transmission Line Catalog and Handbook*, Catalog TL-3, 1970, Times Wire and Cable, Wallingford, Connecticut, 1970.

ham radio

noise bridge calculations

with Texas Instruments Programmable 58/59 calculators

The description of the RX noise bridge which recently appeared in *ham radio*[1] came at a propitious time — I wanted to build a three-element vertical phased array for 20 meters and faced the prospect of tuning the system with a dip meter! A capable noise bridge seemed an ideal answer and the version described by W6BXI and W6NKU was successfully built and tested.

Only one problem remained: I had no Smith charts to convert the series impedance values at the end of the feedline to series impedance values at the antenna feedpoint. A little study also revealed that it would be difficult to read or enter values with any precision in certain regions of the Smith chart. Finally, since I normally have five or six frequency-impedance combinations to process for each trial antenna length (times three for my phased array), the graphical construction techniques using a Smith chart would take a very long time.

the solution — a calculator program

My solution to this dilemma appeared in the form of a newly-purchased Texas Instruments *Programmable 58* calculator. This remarkable calculator packs up to 480 program steps or up to 60 memories in a small hand-held unit. (The user may partition the available memory between program and data storage to suit his needs.) More importantly, the 25-program "Master Library Module," a small plug-in element containing the equivalent of 5000 program steps, provides extensive vector algebra capability; — this is precisely what I needed to perform Smith chart calculations. (If you can afford it, the *Programmable 59* possesses even more memory and provides the capability for storing and entering programs using small magnetic cards; this is especially useful for long programs.)

A little digging in a local library provided the necessary formulas;[2] some rearrangement to suit the problem led to **eq. 1**, below. This calculates the equivalent series impedance Z_t of the termination of a transmission line (characteristic impedance, Z_o) of a given electrical length λ', when the equivalent series input impedance Z_i is known.

$$Z_t = \frac{Z_o\,(Z_i \cos\lambda' - jZ_o \sin\lambda')}{(Z_o \cos\lambda' - jZ_i \sin\lambda')} \qquad (1)$$

Note that all impedances must be treated as *vector* quantities. Once the impedance at the termination (the antenna) is known, an additional calculation quickly provides the vswr at the antenna

$$vswr = \frac{1 + |\rho'|}{1 - |\rho'|} \qquad (2)$$

$$\rho' = \frac{Z_t - Z_o}{Z_t + Z_o}$$

Again, the impedances must be treated as vector quantities.

The program presented in **fig. 1** handles all these complex calculations in a matter of seconds, including the series-parallel impedance conversions and the *range extender* correction described in the original noise bridge article. This makes it easy to perform many trials on an antenna system in a very short time. While the program is long (184 steps), the time spent entering it pays large dividends in time saved during antenna adjustments. (For the *Programmable 59* owners, magnetic card storage makes this a one-time concern.)

program details

The flow chart (**fig. 2**) depicts the various modules of the program. You should consult the noise bridge article for the equations for calculation of X_p, conversion of parallel to series impedance, and the range extender correction.

The correction factor for the range extender can vary with frequency, so register 19 is set aside to contain that value. When you have determined the exact value, simply store it in register 19 when you initialize the calculator.

Since you will normally be working with many bridge readings taken with the same piece of transmission line, each case after the first does not require the input of the line length or its characteristic impedance.

You will notice that the electrical length of the line is expressed in radians instead of degrees. When the calculator enters the "Master Library Module" programs 04 and 05 for the vector calculations, it is set in the radian mode and remains there upon its return. If the equation for the electrical length was expressed in degrees, each case after the first would be in error.

By T. J. Anderson, WD4GRI, 1907 Lodgepole Avenue, North Augusta, South Carolina 29841

TITLE NOISE BRIDGE CALCULATIONS

PROGRAMMER WD4GRI

Partitioning (Op 17) 2 3 9 2 9 Library Module Master Library Printer ____ Cards ____

TI Programmable Program Record

PROGRAM DESCRIPTION

Calculates antenna impedance and VSWR given noise bridge readings.

USER INSTRUCTIONS

STEP	PROCEDURE	ENTER	PRESS		DISPLAY
1	INITIALIZE:				
	2nd Rad				
	-159000 STO 15				
	0.006386/v STO 16				
	extender value STO 19		CLR	RST	0
2	Enter line impedance.	Z_0	R/S		Z_0
3	Enter line length in feet.	l	R/S		l
4	Enter frequency in MHz.	f	R/S		f
5	Enter bridge resistance.	R_p	R/S		R_p
6	Enter bridge capacitance.	C_p	R/S		C_p
7	Enter "1" if range extender used, "0" if not.	"1" or "0"	R/S		R_s (ant)
8	Read R_s (ant)		R/S		X_s (ant)
9	Read X_s (ant)		R/S		VSWR
10	Read VSWR; If another case with same line, go to STEP 4. If not, press CLR and RST & go to STEP 2.				

USER DEFINED KEYS	DATA REGISTERS	
A	0	10 Z_0
B	1	11 λ'
C	2	12 f
D	3	13 R_p
E	4	14 $C_p \rightarrow X_p$
A'	5	15 -159000
B'	6	16 0.006386/v
C'	7	17 R_s
D'	8	18 X_s
E'	09 l	19 Extender value

LABELS (Op 08)

FLAGS 0 1 2 3 4 5 6 7 8 9

TITLE NOISE BRIDGE CALCULATIONS

PROGRAMMER WD4GRI

TI Programmable Coding Form

LOC	CODE	KEY	COMMENTS
0	42	STO	
1	10	10	
2	91	R/S	
3	42	STO	
4	09	9	
5	91	R/S	
6	42	STO	
7	12	12	
8	91	R/S	
9	42	STO	
10	13	13	
11	91	R/S	
12	42	STO	
13	14	14	
14	91	R/S	
15	94	+/-	
16	42	STO	
17	17	17	
18	43	RCL	
19	19	19	
20	49	2nd Prd	
21	17	17	
22	43	RCL	
23	09	9	
24	65	x	
25	43	RCL	
26	12	12	
27	65	x	
28	43	RCL	
29	16	16	
30	95	=	
31	42	STO	
32	11	11	
33	43	RCL	
34	15	15	
35	55	÷	
36	43	RCL	
37	12	12	
38	55	÷	
39	43	RCL	
40	14	14	
41	95	=	
42	42	STO	
43	14	14	
44	65	x	
45	43	RCL	
46	13	13	
47	71	SBR	
48	02	2	
49	00	00	
50	42	STO	
51	18	18	
52	43	RCL	
53	13	13	
54	65	x	
55	43	RCL	
56	14	14	
57	71	SBR	
58	02	2	
59	00	00	
60	44	SUM	
61	17	17	
62	43	RCL	
63	17	17	
64	65	x	
65	43	RCL	
66	11	11	
67	39	2nd cos	
68	95	=	
69	42	STO	
70	01	1	
71	43	RCL	
72	11	11	
73	39	2nd cos	
74	65	x	
75	43	RCL	
76	18	18	
77	75	-	
78	43	RCL	
79	11	11	
80	38	2nd sin	
81	65	x	
82	43	RCL	
83	10	10	
84	95	=	
85	42	STO	
86	02	2	
87	43	RCL	
88	11	11	
89	39	2nd cos	
90	65	x	
91	43	RCL	
92	10	10	
93	85	+	
94	43	RCL	
95	11	11	
96	38	2nd sin	
97	65	x	
98	43	RCL	
99	18	18	
100	95	=	
101	42	STO	
102	03	3	
103	43	RCL	
104	11	11	
105	38	2nd sin	
106	65	x	
107	43	RCL	
108	17	17	
109	95	=	
110	94	+/-	
111	42	STO	
112	04	4	
113	36	2nd Pgm	
114	04	04	
115	18	2nd C'	
116	65	x	
117	43	RCL	
118	10	10	
119	95	=	
120	91	R/S	
121	32	x⇄t	
122	65	x	
123	43	RCL	
124	10	10	
125	95	=	
126	91	R/S	
127	43	RCL	
128	10	10	
129	49	2nd Prd	
130	01	1	
131	49	2nd Prd	
132	02	2	
133	43	RCL	
134	01	1	
135	42	STO	
136	03	3	
137	43	RCL	
138	02	2	
139	42	STO	
140	04	4	
141	43	RCL	
142	10	10	
143	22	INV	
144	44	SUM	
145	01	1	
146	44	SUM	
147	03	3	
148	36	2nd Pgm	
149	04	04	
150	18	2nd C'	
151	36	2nd Pgm	
152	05	05	
153	12	B	
154	42	STO	
155	01	1	
156	85	+	
157	01	1	
158	95	=	
159	55	÷	
160	53	(	
161	01	1	
162	75	-	
163	43	RCL	
164	01	1	
165	54	)	
166	95	=	
167	91	R/S	
168	61	GTO	
169	00	0	
170	06	6	
200	33	x^2	
201	55	÷	
202	53	(	
203	43	RCL	
204	13	13	
205	33	x^2	
206	85	+	
207	43	RCL	
208	14	14	
209	33	x^2	
210	54	)	
211	95	=	
212	92	INV SBR	

MERGED CODES: 62, 63, 64, 72 STO, 73 RCL, 74 SUM, 83 GTO, 84, 92 INV SBR

TEXAS INSTRUMENTS

fig. 1. TI-58/59 program for calculating antenna impedance and vswr from RX noise bridge measurements.

Eq. 3 gives the electrical length, λ', in radians.

$$\lambda' = \frac{0.02095lf}{v} (meters) = \frac{0.006386lf}{v} (feet) \qquad (3)$$

Where

l = physical line length in meters or feet
f = frequency in MHz
v = line velocity factor

The factor ($0.02095/v$ or $0.006386/v$) is entered in register 16. For RG-58/U ($v = 0.66$), this factor is 0.03174 for metric lengths (0.009675 for feet). A subroutine (steps 200-212) is used in the parallel-series conversions to reduce the length of the program.

using the program

You should become fairly familiar with your calculator's programming manual before you attempt to enter this program. The *TI-58* is a very complex logic system. When you are ready, do the following:

1. Turn the calculator on and enter the program. The machine's extensive editing capabilities and the program key codes will assist you if difficulties arise.

2. Convert the calculator to the radian mode (2nd Rad).

3. Enter the appropriate constants (see program) in registers 15, 16, and 19.

4. Press CLR and RST.

5. In order: enter characteristic line impedance in ohms, line length in meters (feet), frequency in MHz, bridge resistance, bridge capacitance, and range extender status, pressing R/S after each entry. (Range extender status = 1 if the extender was used, 0 if not).

6. After a few seconds the calculator will display the series resistance at the antenna; record the value and press R/S.

7. Next the calculator will display the series reactance at the antenna; record the value and press R/S.

8. Finally, the calculator will display the vswr.

9. If you are using the same transmission line at a new frequency, enter the new frequency in MHz, the bridge resistance, the bridge capacitance, and the range extender status pressing R/S after each entry. The results are displayed as before. If you are using a different transmission line go to **step 4.** (Remember to change register 16 if the velocity factor of the line is different.)

It is very important to verify correct entry of the program before use. You can do this by runnning a couple of cases from **table 4** of the noise bridge article.[1] Choose one case using the adaptor and one

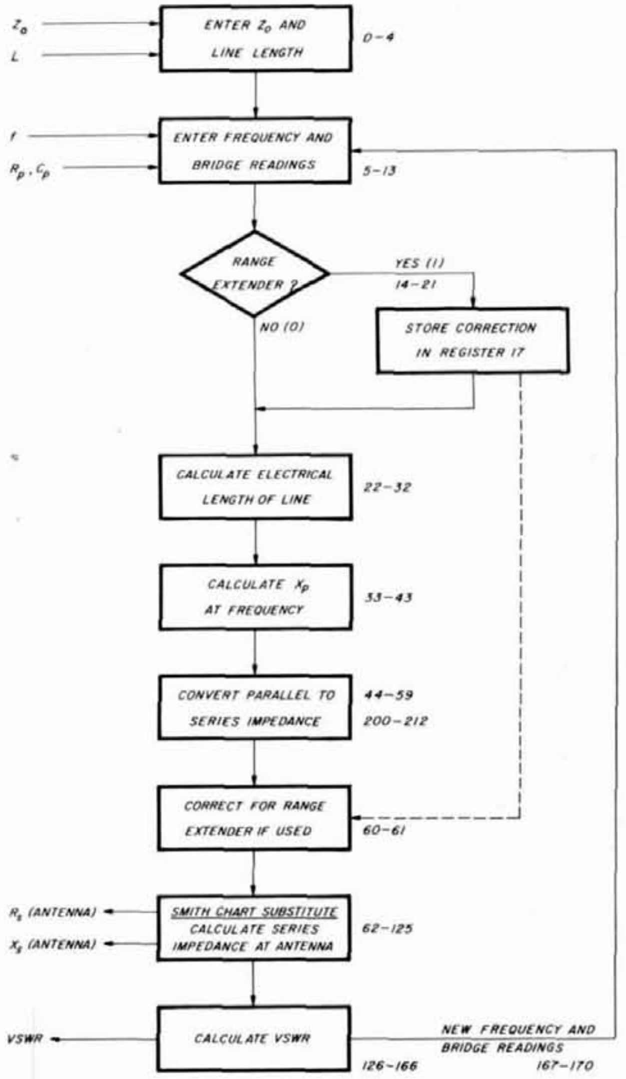

fig. 2. Flow chart for the TI-58/59 program for calculating antenna impedance and vswr.

without. If you run into trouble, check the key codes versus those appearing in the program.

summary

I have used this program with great success to tune my 20-meter array. What would have taken weeks to accomplish with Smith charts required only a couple of evenings with the Programmable *TI-58* calculator. As you become familiar with the calculator and the program, you can adapt them to your own special needs. I'm sure you'll be very impressed with the calculating power this system provides.

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ham radio

tree-mounted ground-plane antenna for 80 meters

Using elevated ground-plane antennas you eliminate the need for an extensive ground radial system

Even though winter has passed, and solar activity is on the rise, 80 meters always provides excitement for the nocturnal amateur who craves DX. Many amateurs have a desire to erect an antenna that will punch through the pile-ups better than the standard dipole or inverted V. And, people's thoughts ultimately turn to the vertical antenna with its inherent low angle of radiation. However, most amateurs discard this idea because of the necessity for an extensive ground radial system. Having faced the same perplexing problem, a viable solution was found in Orr's *Radio Handbook*[1] — his description of a low-band ground-plane antenna. Its simplicity and four elevated ground radials readily pointed to tree mounting the antenna, and with several tall trees on the property, my solution was near at hand.

My initial fear of rf absorption by the foliage and oblique branches was dispelled by the realization that at the 80-meter frequencies the wavelength is much too long. There would be some inductive coupling to the tree trunk, which is conductive to a degree, especially when the sap is flowing, but the effect is negligible.

The first hurdle was mounting the 18.9 meter (62 foot) vertical section 4.6 meters (15 feet) above the ground in a maple tree only 18.3 meters (60 feet) tall. I came to the conclusion that since Marconi's original grounded antenna was in the shape of an inverted L, my tree-mounted version could have a bent top section as well. An added benefit of the bent top is the additional capacitance which makes the antenna appear electrically longer.

construction

The first ground plane was constructed in my basement using no. 12 AWG (2.1 mm) insulated wire for the vertical section and ground-plane radials (**fig. 1**). From past experiences, I'd learned that insulated wire considerably reduces precipitation static and has no noticeable effect on the radiation or reception of signals.

I designed the antenna for 3.8 MHz by using the standard quarter-wavelength vertical formula of:

vertical element length =

$$\frac{71.3}{f(MHz)}\ meters\ or\ \frac{234}{f(MHz)}\ feet$$

Since the horizontal portion tended to top load the antenna, the antenna resonated at 3785 kHz, slightly lower than the design frequency. This was of little consequence because of the broadband characteristics of the antenna.

The four radials were cut 2-1/2 per cent longer than the vertical radiator. Each was soldered to a mounting hole of an SO-239, with the quarter-wave vertical section soldered to the center pin of the connector. For mounting and tying down, small ceramic insulators were attached to the ends of the ground radials and quarter-wavelength vertical section.

mounting and installation

Mounting the antenna in the maple tree was much easier than I had anticipated. After deciding where the antenna would fit with no obstructions to the spreading ground radials, I used the bow and arrow technique to thread a small fishing line over the appropriate branch. Adding weight to the tip of the

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arrow will ensure a vertical descent and prevent hang-up in the branches.

With 36.5 meters (120 feet) of plastic clothes line attached to the monofilament line and also the radiator of the antenna, it was then easy to pull the rope through and subsequently raise the antenna base 4.6 meters (15 feet) above the ground. As a result, the top portion of the quarter-wave section extends about 4.6 meters (15 feet) from the top branches, slanted towards the point of tie down. I attached the clothes line as high as possible in a nearby tree to minimize the slant of the bent top.

Since most of the antenna current flows in the lower half of the vertical quarter-wavelength section, the entire upper half can be bent over without serious degradation. It should be noted that the longer the flat top or bent portion, the greater will be the top loading, and consequently, the lower the resonant frequency.

The radials were slanted to tie points about 1.5 meters (5 feet) above the ground, and spaced approximately 90 degrees apart. They also serve as guys to stabilize the base of the antenna, but should not be brought too close to the ground because the electric field must be confined between the vertical radiator and the groundplane, rather than the lossy ground. The return current will then flow in the highly conductive groundplane radials with little power

The first ground plane antenna, but with six radials instead of the original four. The spiral at the base of the antenna is plastic line taped to the base portion for reinforcement.

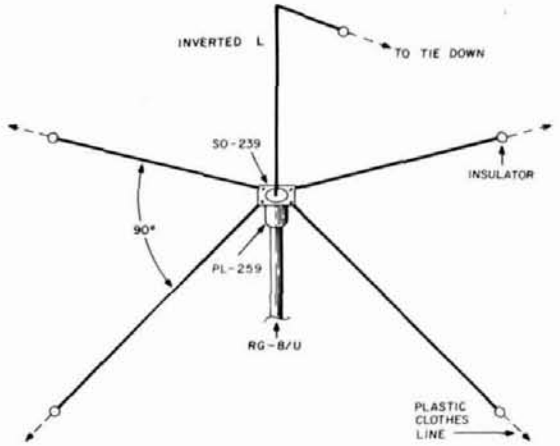

fig. 1. Diagram of one of the tree-mounted ground-plane antennas. The radials are spaced approximately 90 degrees apart and slope at an angle of about 20 degrees from the horizontal.

loss. This is an important consideration close to the base of the antenna, since the current is at a maximum. If the base of the groundplane must, from necessity, be located a short distance above the ground, additional radials (at least four more) must be installed for efficient operation of the antenna. Even spacing of the radials provides a more effective groundplane.

A few words of caution are in order here. The end of each radial is at a high rf voltage point. The position where a radial is tied down should not be in a location where it can be accidentally touched.

base impedance

The impedance at the base of a ground-plane antenna is about 35 ohms with horizontal groundplane radials, but will increase as the radials are sloped downward. A slope of 45 degrees from the horizontal will raise the base impedance to about 50 ohms. However, a 45-degree slope is quite difficult to achieve in an 80-meter tree-mounted version. My radials slope at an angle of 20 degrees, producing a base impedance of 40 ohms. Attempts to lower the radials or increase the droop did not significantly change the impedance. By using a 50-ohm, low loss, coaxial cable, the 1.25:1 mismatch is of little consequence.

initial testing

The tree-mounted, ground-plane antenna performed beyond my expectations. It loaded up beautifully and the swr at the end of a one-half wavelength of transmission line was close to the expected 1.25:1. At the band edges, the swr did not exceed 2:1, which verified the broadband characteristic of the antenna.

On the first night of extensive on-the-air tests,

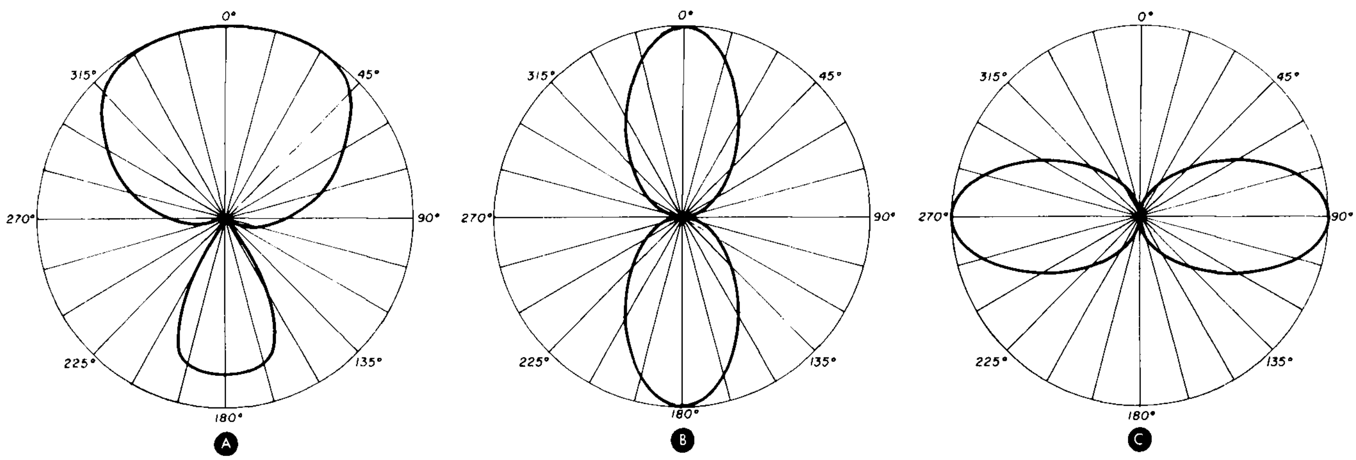

fig. 2. Horizontal radiation patterns for the pair of phased ground planes, spaced one-half wavelength apart. Pattern A is with a phase difference of 90 degrees, B with 180 degree phase difference, and C, no phase difference.

contacts into Europe showed the ground-plane to be one S-unit stronger than my sloping quad antenna.[2] Subsequent European contacts, and a later contact with ZL2BT, confirmed the superior performance of the ground-plane antenna.

During several weeks of operating with a single ground-plane antenna, I found that the horizontal portion of the vertical element provided an advantage I had not anticipated. It contributed sufficient high angle radiation to permit good short skip contacts. Although the sloping quad was the superior antenna for short and intermediate skip, the ground-plane held its own.

phasing distance

I followed the reasoning that if one ground-plane antenna worked well on 80 meters, then two antennas should perform better, especially a pair properly phased. Using this as the basis, I began pacing distances between the tall trees on the property, and found another maple 39.6 meters (130 feet) from the tree supporting the ground-plane antenna. At 3.8 MHz, 39.6 meters (130 feet) is approximately one-half wavelength since:

$$\lambda/2 = \frac{150}{f(MHz)} \text{ meters or } \frac{492}{f(MHz)} \text{ feet}$$

The second maple tree was over 16.5 meters (55 feet) tall, and before long, another ground-plane antenna was constructed and installed in the same manner as the first.

phasing the antennas

Since simplicity was the byword in constructing and mounting the antennas, I planned a simple system of phasing. By design, I cut the length of the two transmission lines to make them 90 degrees out of phase. In this case, the length of coaxial line to the first antenna was 1/2 wavelength at 3.8 MHz, and the length of line to the second was 3/4 wavelength at the same frequency. After initially cutting the coaxial lines a little long, I grid dipped and pruned the lines to their exact electrical lengths.[3] With both coaxial lines fed from a common feedline, the antennas were then 90 degrees out of phase. This relationship produced a pattern with a broad forward lobe to the Northeast, and a minor lobe to the Southwest as shown in **fig. 2**.[4] Now it was just a simple matter of cutting a half-wave phasing line to make the antenna system perform the same in the opposite direction.

One method of phasing antennas is to have equal electrical lengths of coaxial line from each antenna to the operating position. Then it is a simple procedure to use a quarter wavelength coaxial line to switch directivity. With the high cost of coaxial cable, it may be desirable to space the antennas one-quarter wavelength apart, and use a quarter wavelength of coaxial line from each antenna to a coaxial relay. This method permits the use of a single coaxial line from the relay to the operating position. Also, when quarter-wave spacing is used, the horizontal plane radiation is in the form of a cardioid (broad heart-shaped) pattern, providing a better front-to-back ratio as shown in **fig. 3**.

matching

As mentioned earlier, the impedance measured at the base of each antenna was 40 ohms. Feeding the antennas with 50-ohm coaxial lines presented no major problem when each antenna was used separately. However, when the two coaxial lines were joined at the coaxial-tee connector, the resultant impedance was about 20 ohms. The 2.5:1 mismatch presented no problem in antenna operation or amplifier loading. In addition, the loss caused by an swr of 2.5:1 in 30 meters (100 feet) of RG-8/U at 4 MHz, is a fraction of a dB and can be ignored.[5]

For those who do have loading problems, it might be desirable to use a 70-ohm coaxial line from each antenna to the coaxial tee connector. This line must be one-quarter wavelength long or odd multiples of a quarter wavelength. Each line acts as a quarter-wave matching transformer (Q section). Assuming a 50-ohm base impedance, the matching sections transform this up to 100 ohms, giving you a 50 ohm impedance when the lines are joined together.

performance

After three years of constant use, the phased ground-plane system consistently outperformed my sloping quad and new delta loop.[6] Despite the minor lobe to the rear, I've experienced a 10 to 15 dB front-to-back ratio. The very broad major lobe provides approximately 120 degrees of effective coverage. The minor lobe to the rear, although contributing to some QRM and noise from the undesired direction, can be used to an advantage when rapid switching of direction is not possible.

Receiving with the phased ground-planes was a new experience. Unreadable signals on the loop antennas became Q-5 on the phased ground-planes. Precipitation static was never bothersome during the heaviest rain or snowfall. The capability of the antennas to accept high-angle as well as low-angle signals provided additional versatility when simultaneously operating with stations close by and far off. The very broad major lobe contributes little front-to-side rejection. This characteristic provides advantages as well as disadvantages. I used a quarter-wavelength phasing line to change the radiation pattern to an in-phase broadside array, or a 180 degree out-of-phase, end-fire array as shown in **fig. 2.**

A loop antenna normally accepts less atmospheric noise than a dipole or similar antenna, due to the noise cancelling effect of the closed loop. However, over a period of three noisy summers, I experienced a greater signal-to-noise ratio on the phased ground-planes than with the loop antennas. The phased ground-planes are certainly more susceptible to man-made noise, due to its predominant vertical polarization.

conclusion

As can be expected with any antenna system, improvements are attempted. Some succeed, while others do not. I found that increasing the number of radials from four to twelve slightly improved antenna performance when transmitting. On receiving, the improvement was quite noticeable from a standpoint of signal-to-noise ratio. The base impedance of the antennas decreased slightly due to the increased current in the additional radials.

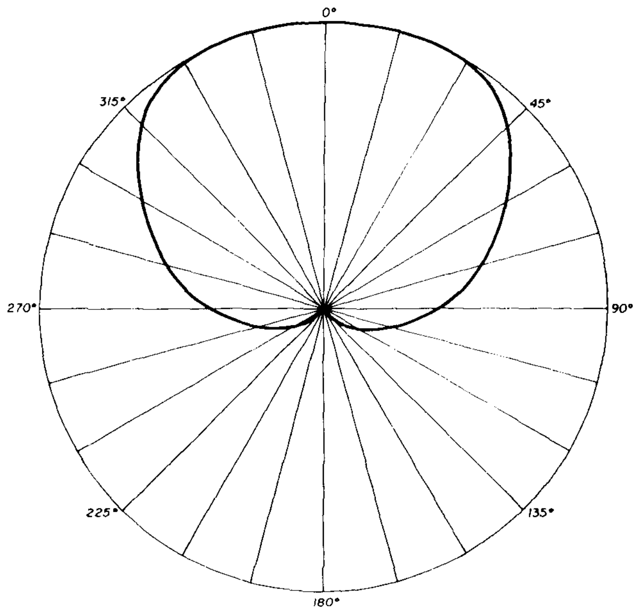

fig. 3. The unidirectional cardioid pattern is produced by spacing the antennas one-quarter of a wavelength apart and using a 90 degree phasing line.

I also tried connecting the ends of the sloping radials where they almost overlap between antennas. The rationale was to provide a common ground-plane. This experiment was a failure, since each antenna must operate independently, coupled only by their fields.

It would appear that the antennas are a bit unsightly with the radials fanning out in all directions. On the contrary, I had to point out the antennas to several visiting amateurs who had difficulty spotting them from the operating position.

And finally, I want to express my gratitude to all the US and foreign amateurs who patiently contributed their time during the months of checking and testing the antenna system.

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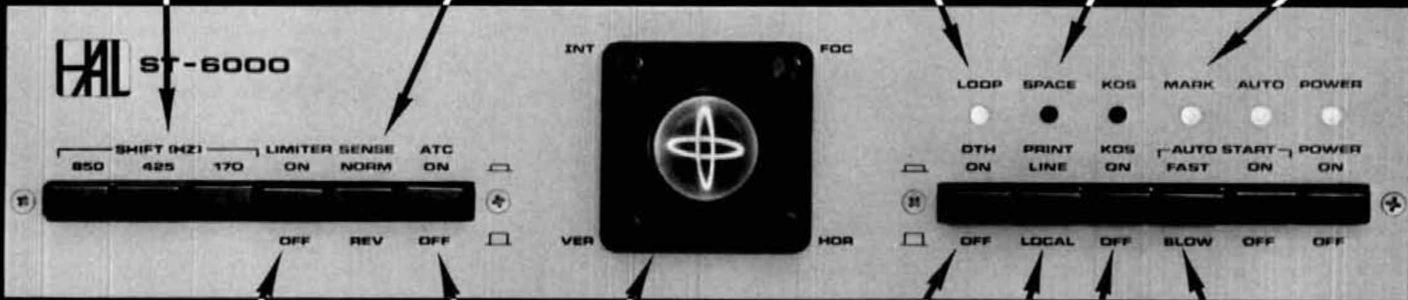

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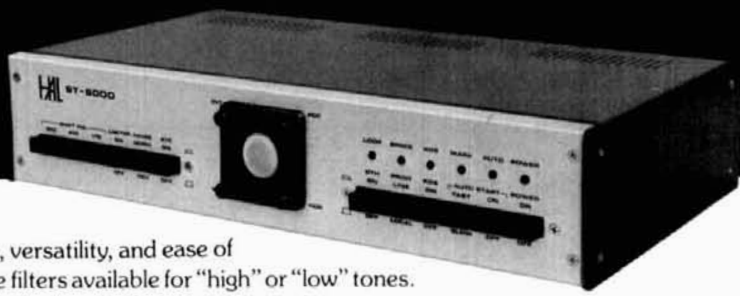

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design of
omega-matching networks

A straightforward procedure for designing antenna networks

The omega network has been used for years, mostly by amateur experimenters, to match a coaxial line or open-wire feedline to a thin, linear antenna element as shown in **fig. 1**. Although this matching network looks somewhat like the more common gamma network, two capacitors are used. In this illustration L is the length of one-half of the driven dipole element (or total length of a monopole element). The dimension l_Ω is the length of the omega-matching rod and is the same as a gamma matching rod; S is the center-to-center spacing between the omega rod and a *parallel* driven element.

The reactance X_Ω is the shunt omega capacitor at the omega-rod feedpoint. X_C is the series capacitor used to tune out any reactance in the antenna system's input impedance. R_o is the characteristic resistance of the coaxial transmission line, which is assumed, for the purposes of this discussion to be lossless.

The practical omega match shown in **fig. 1** (less capacitor X_C) can be represented by the simple block diagram of **fig. 2**. In this diagram H_z is the impedance step-up ratio, Z_a is the antenna impedance, X_S is the reactance of the omega rod, and X_Ω is the reactance of the omega capacitor. The antenna impedance step-up ratio, H_z, is a function of the diameters of the omega rod and driven element, as well as the spacing between them and is given by the following formula:[1]

$$H_z = \left[1 + \frac{\left(\cosh^{-1} \dfrac{4S^2 - D^2 + d^2}{4Sd} \right)}{\left(\cosh^{-1} \dfrac{4S^2 + D^2 - d^2}{4SD} \right)} \right]^2$$

where D is the diameter of the driven element, d is the diameter of the omega rod, and S is the spacing between the two. This factor is plotted in normalized form in **fig. 3** for omega match designs. In plotting this graph it was assumed that H_z is at least 4:1, a realistic assumption in most amateur applications.

The quantity Z_a is one-half the total input impedance of a balanced antenna and is equal to

$$Z_a = \frac{Z_a'}{2} = R_a \pm jX_a \; ohms$$

where the quantity Z_a' is the antenna driving point impedance.

The reactance of the omega rod, X_s, is equal to $jZ_o \tan kl_\Omega$ ohms when the quantity kl_Ω is less than 90 electrical degrees as will be assumed here and $k = 2\pi/\lambda$ radians per meter. The impedance of the omega rod in air, Z_o is given by

$$Z_o = 60 \cosh^{-1} \frac{4S^2 - D^2 - d^2}{2Dd} \; ohms$$

where

D = diameter of antenna driven element

d = diameter of omega rod

S = center-to-center spacing between the omega rod and the *parallel* antenna driven element

Fig. 4 shows a graph of Z_o plotted as a function of normalized values of S/D and D/d.

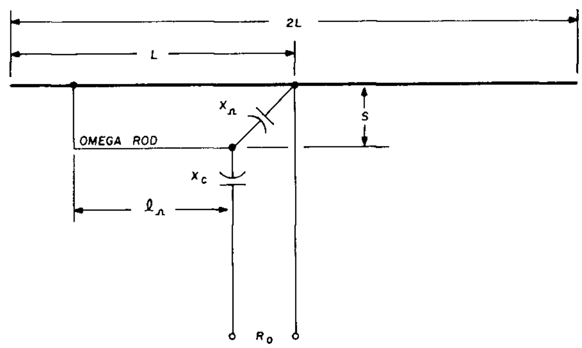

fig. 1. Layout of the basic omega match. L **is one-half the length of the driven dipole.** X_Ω **is the omega capacitor, and** X_C **compensates for the inductive reactance component appearing at the omega rod input point.**

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The length of the omega rod is approximately

$$l_\Omega = \frac{(kl_\Omega)^o}{0.01233 f_{MHz}}\ cm = \frac{(kl_\Omega)^o}{0.03188 f_{MHz}}\ inches \quad (1)$$

From **fig. 2** the complex antenna system input impedance, Z_i, may be written as

$$Z_i = M \angle \psi\ ohms \quad (2)$$

Where

$$M = \frac{(X_\Omega X_s H_z \sqrt{R_a^2 + X_a^2})}{\sqrt{[H_z R_a (X_\Omega + X_s)]^2 + [H_z X_a (X_\Omega + X_s) + X_\Omega X_s]^2}}\ ohms$$

$$\psi = tan^{-1}\left[\frac{(X_\Omega + X_s) H_z (R_a + X_a)}{X_\Omega X_s R_a} + \frac{X_a}{R_a}\right]\ degrees$$

For **eq. 2** to match a high-frequency lossless coaxial transmission line with a characteristic resistance, R_o,

$$M\ cos\ \psi = R_o\ ohms \quad (3)$$

By using the above substitutions and manipulating, the solution may also be written as a quadratic in Q whose positive discriminant root is

$$Q = A + \sqrt{A^2 + B}\ ohms \quad (4)$$

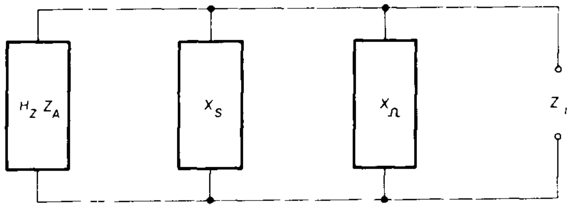

fig. 2. Equivalent circuit of the omega-matching system shown in fig. 1. Z_a is one-half the dipole input impedance, H_z is the impedance step-up ratio, X_s is the reactance of the omega rod, and X_Ω is the reactance of the omega capacitor.

Where

$$Q = \frac{X_s X_\Omega}{H_z (X_s + X_\Omega)}$$

$$A = \frac{R_o X_a}{H_z R_a - R_o}$$

$$B = \frac{R_o (R_a^2 + X_a^2)}{H_z R_a - R_o}$$

Equation 4 has the restriction that the negative sign of X_Ω must be larger than the positive sign of X_s (in this assumed case) and that

$$H_z > \frac{R_o}{R_a}\ numeric$$

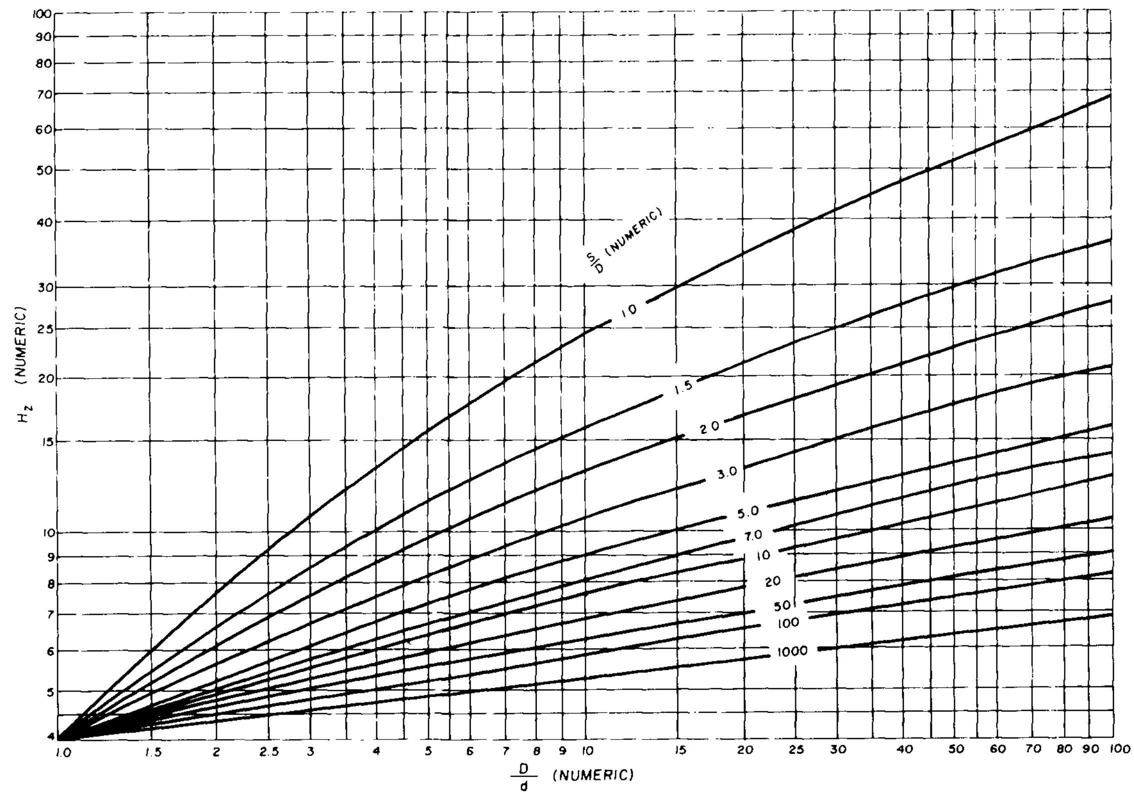

fig. 3. Matching network's impedance step-up ratio, H_z, as a function of element diameters and spacing. D is the diameter of the dipole element, d is the diameter of the omega rod, and S is the center-to-center spacing between them.

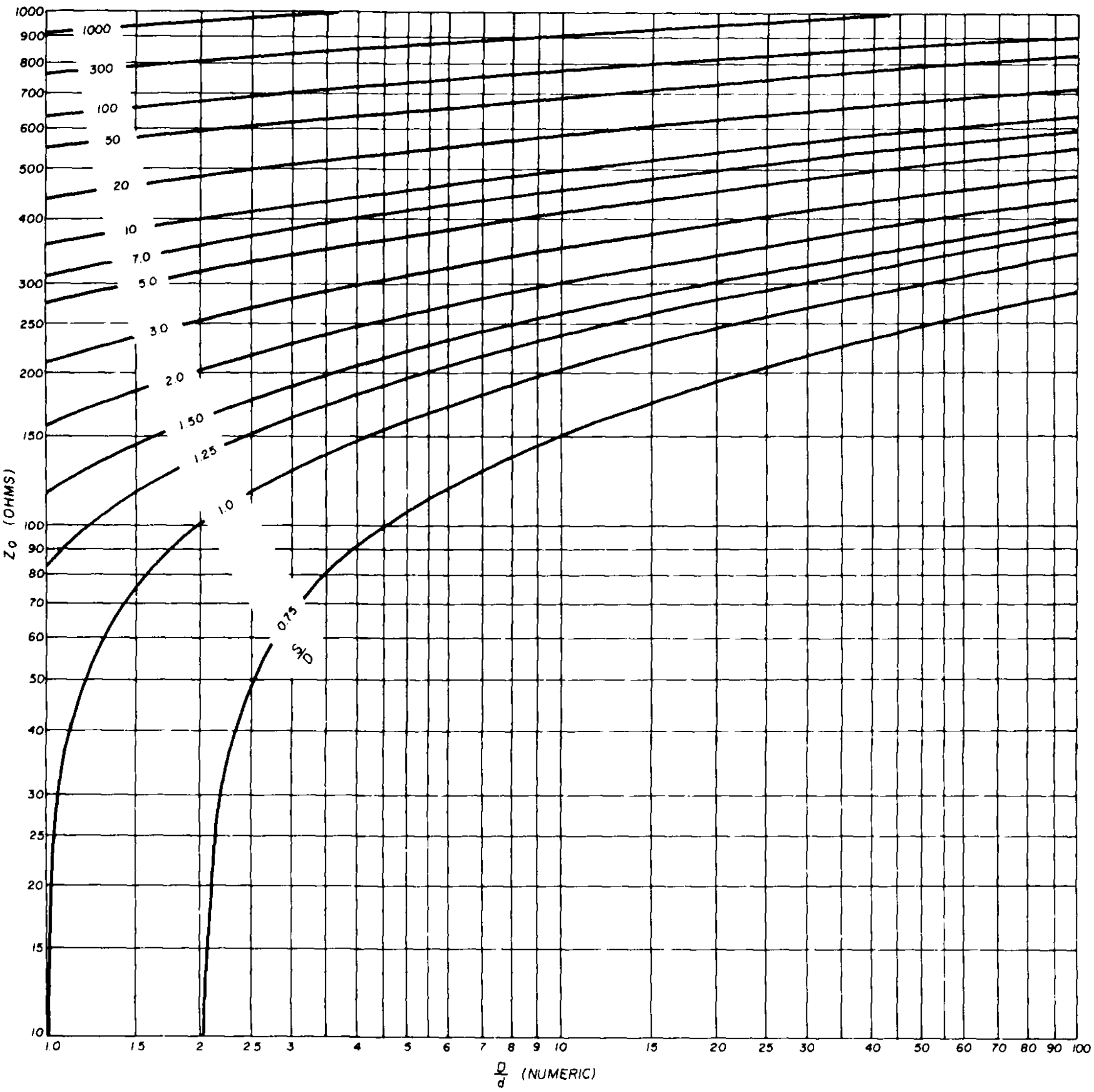

fig. 4. Characteristic impedance of the omega rod, Z_o, as a function of element diameters and spacing.

In the above definition of Q, when X_Ω is infinite (open circuit), the solution of Q is exactly the same as that for a gamma-matching network[1] because it can be seen from inspection that when $X_\Omega = \infty$, the gamma and omega matching networks are identical.

Once Q in **eq. 4** has been determined as a function of A and B (i.e., R_o, H_z, R_a, and X_a) the solution for X_Ω and X_s appears to be an arbitrary one of two circuit elements X_Ω and X_s in parallel or

$$H_z Q = k' = \frac{X_s X_\Omega}{X_s + X_\Omega} \; ohms \; |X_\Omega| > |X_s| \qquad \textbf{(5)}$$

A graph of this equation, for various values of $H_z Q$, is plotted in **fig. 5**. In **fig. 6** is plotted a graph of the omega rod reactance X_s as a function of the impedance of the omega rod, Z_o, the omega rod length kl_Ω.

For **eq. 2** to provide a reactive component equal in magnitude but opposite in sign to X_C in **fig. 1**

$$M \sin \psi = -X_C \; ohms \qquad \textbf{(6)}$$

Combining this with **eq. 3** yields

$$X_C = -R_o \tan\psi$$

$$= -\frac{R_o}{R_a}\left[\frac{(X_\Omega + X_s)H_z(R_a^2 + X_a^2)}{X_\Omega X_s} + X_a\right] ohms$$

And, combining with Q,

$$X_C = -\frac{1}{2\pi f C_C}$$

$$= -\frac{R_o}{R_a}\left[\frac{R_a^2 + X_a^2}{Q} + X_a\right] ohms$$

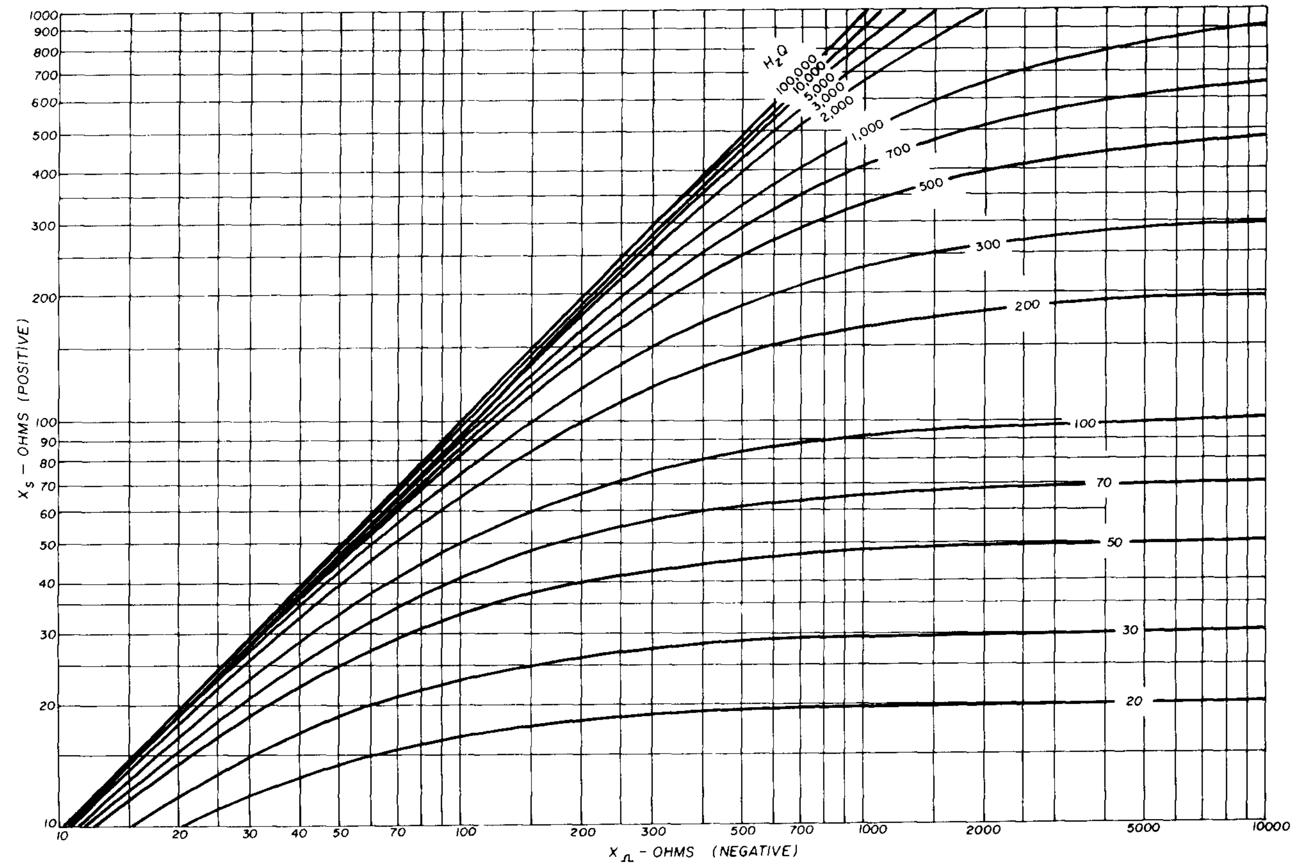

fig. 5. Relationship of the reactance of the omega rod, X_s, and reactance of the omega capacitor, X_Ω, as function of the H_zQ product (see text).

From which

$$C_C = \frac{10^6}{2\pi(E+F)f_{MHz}}\ pF \qquad \textbf{(8)}$$

Where

$$E = \frac{R_o(R_a^2 + X_a^2)}{R_aQ}\ ohms$$

$$F = \frac{R_oX_a}{R_a}\ ohms$$

design procedure

When designing an omega-matching network it might seem that the following procedure could be used with good success:

1. Use the graph of **fig. 3** to select a transformation ratio H_z *vs* the desired S/D and D/d where $H_z > R_o/R_a$.

2. Use **fig. 4** to determine the impedance of the omega rod, Z_o, from the S/D and D/d ratios selected in **step 1**.

3. Use **fig. 6** to determine a practical, *realistic* value of omega rod reactance, X_s, *vs* the rod's impedance, Z_o, and length, kl_Ω.

4. When kl_Ω is known in degrees from **fig. 6**, the physical length of the omega rod, l_Ω, may be calculated from **eq. 1**.

5. Calculate Q from **eq. 4**. When this value of Q is multiplied by H_z, you can select the correct H_zQ line on **fig. 5**. This provides the reactance of the omega capacitor, X_Ω, at the bottom of the graph.

6. The value of the omega capacitor in pF is calculated at the frequency of interest using **eq. 8**.

The problem with this design procedure is in the selection of a "realistic value" of the omega rod reactance in **step 3**. That is, should X_s be selected large so X_Ω is large, or should X_s be selected small so X_Ω is small? In gamma-matching networks (the same omega-matching networks with $X_\Omega = \infty$, as noted above), it's desirable to make X_s large to minimize the IR losses. In the omega network the current I is

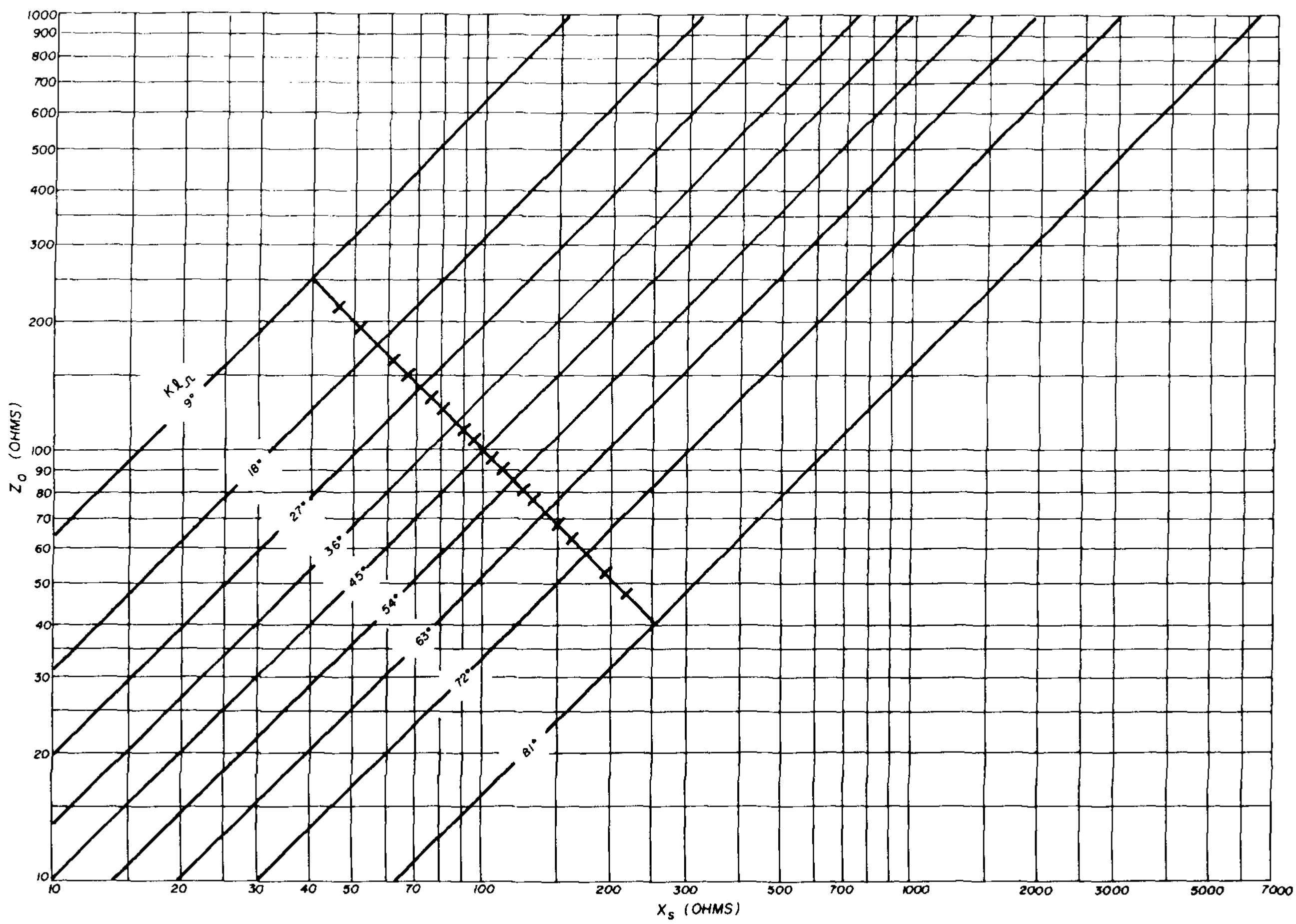

fig. 6. Reactance of the omega rod, X_s, as a function of the rod's characteristic impedance, Z_o, and length, kl_Ω.

divided between X_s and X_Ω so the answer may not be so readily apparent.

The answer will become apparent, however, if two vector graphs are plotted so the currents can be examined for the two cases. The two vector graphs, **figs. 7** and **8**, assume the same rf power, antenna driving point impedance, Z_a' as well as H_z and R_o. **Fig. 7** uses a gamma-matching network as an example of $X_\Omega = \infty$, and **fig. 8** shows an omega-matching network with $|X_\Omega| = |2X_s|$.

These two graphs give complete insight into the operation of both the gamma- and omega-matching networks. For any particular antenna impedance H_zZ_a to resistively match a transmission line through a series capacitive reactance, it's necessary that the antenna impedance, together with the impedances of its shunt correcting elements, provide a complex input impedance with a positive angle; also, the magnitude of the real part must be equal to the characteristic resistance of the transmission line, and the imaginary part must be equal but opposite in sign to the series capacitive reactance. With this in mind it shouldn't be surprising to see that the magnitudes of V_{in}, I_{in}, V_{Xs}, V_{HzRa}, V_{HzXa}, and I_a of the gamma match in **fig. 7** are the same as those for the omega match in **fig. 8**.

The difference between the two matching networks is that I_{X_s} in the gamma network is the vector sum of I_{X_s} and $I_{X\Omega}$ in the omega network. From this it is obvious that I_{X_s} in the gamma network is less than I_{X_s} in the omega network, and that the magnitude of X_s in the gamma network is greater than the magnitude of X_s in the omega network.

The omega network is easier to design than the gamma network because the only restriction upon X_s in the omega network is that it must be *less* in magnitude than it is in the equivalent gamma network. In the omega network the difference between the two values of X_s can be compensated by adjusting the value of X_Ω. Because of this, the rod length kl_Ω in the omega network *can* be made shorter than kl_Γ of the equivalent gamma network.

At the lower operating frequencies where the length of the rod is apt to be very long, the omega network has merit. However, it is likely that the efficiency and bandwidth of the omega network will be less than the equivalent gamma network because of the higher IR losses in omega network's reactance

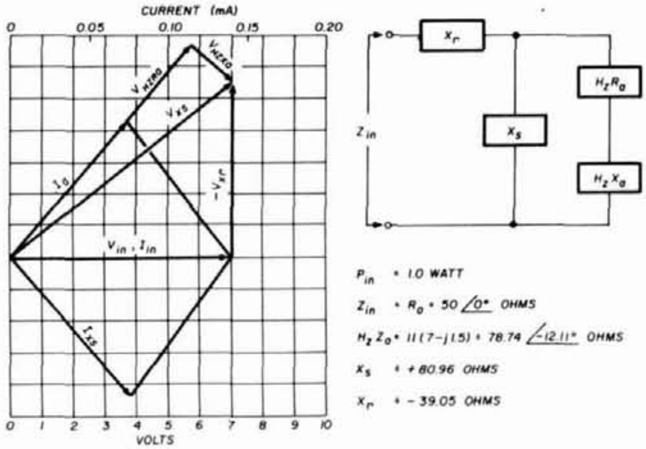

fig. 7. Voltage and current distribution of a gamma-matching network designed to match a 50-ohm coaxial line to a 10-element beam ($Z_a = 7 - j1.5$ ohms). Voltage and current values are based on a power input of 1 watt.

X_s. Any predication on the fact that these reactive circuit elements are lossless is rather naive!

The double omega (or what I call the theta network) can be used to match a balanced two-wire transmission line to the driven element of a balanced dipole. In this case the balanced line characteristic resistance R_o' is halved to R_o, and the solutions to *each* arm are X_s, X_Ω, and X_C in terms of R_o, H_z, R_a, and X_a. The results are merely imaged or flipped over to the other arm of the theta network.

reference

1. Harold Tolles, W7ITB, "Gamma Match Design," *ham radio*, May, 1973, page 46.

ham radio

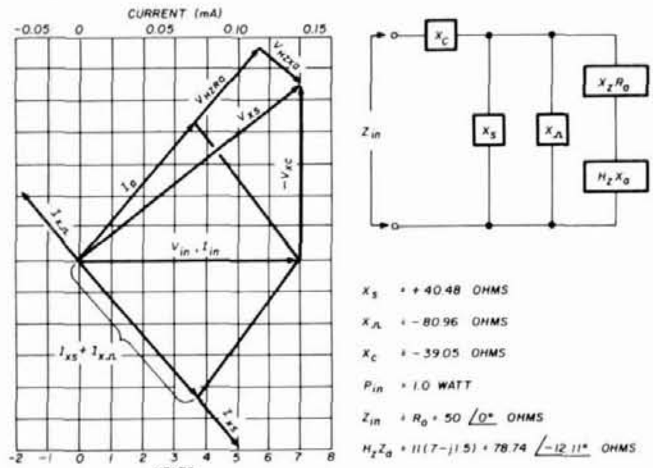

fig. 8. Voltage and current distribution of an omega-matching network designed to match a 50-ohm coaxial line *to the same load as the gamma match in fig. 7* ($Z_a = 7 - j1.5$ ohms). Voltage and current values are based on a power input of 1 watt. Note that I_{X_s} in the omega network is greater than I_{X_s} in the gamma network when the rod is too short to use a gamma network. So, current (and *IR* loss) is greater in the omega match.

improved indicator system for the Hy-Gain 400 antenna rotator

With this circuit you can read antenna pointing direction during antenna rotation

The designers of the Hy-Gain 400 rotator apparently didn't realize that most amateurs want to know which way their antenna is pointing at any moment, and an indicator was provided that didn't provide this information during antenna rotation. The *ARRL Antenna Book* (13th edition) contains information on a replacement directional indicator that certainly will work. However, it requires an 8-wire control cable whereas the original indicator required a 5-wire cable and Hy-Gain supplied such a cable suitable for the 110 volts that the rotator motor requires. So it was rather frustrating to read about an approach that's available and convenient for those who had not yet installed the rotator on a tower, but less so for those who had already installed the rotator and wished to change the indicator.

This article describes a replacement direction-indicating control unit that involves no modifications of the rotator and is compatible with the original 5-wire cable. It has an additional feature that permits control from multiple locations – a convenience for those who operate from several locations as I do. The ideas in this article are adaptable to other types of rotators where multiple control is wanted. Further, it's one way to get a direction indication from a rotator if one end of the indicator pot has become disconnected or if a failure has occurred in the indicator circuit. The latter failure is probably rare but is mentioned as an emergency or permanent solution that could still make use of the original indicator meter.

The crucial matter is that the indicator resistor in the Hy-Gain 400 rotator was wired as a rheostat (even though it was physically a three-terminal pot). If the familiar ohmmeter circuit is used to read the rheostat setting, and hence antenna heading, the scale will be grossly nonlinear and not very satisfactory. The cure is, therefore, to supply the rheostat with a *constant-current* source and then read the voltage across it, which will then be directly proportional to the angle of rotation.

constant-current supply

A recent advertisement gave the clue to a conveniently available constant-current source, the popular voltage-regulator ICs, which are used to provide a regulated 5, 12, or 15 volts. A single resistor connected between what is normally the output terminal and the normal ground – now using the ground as the constant-current output – does the trick. The current is constant *only* if the supply voltage is high enough. In this case, 24 Vdc was required to supply 10 mA constant current through a resistor varying between 0-1000 ohms. (This is the value of the rheostat in my Hy-Gain 400; but watch it, as the *ARRL Antenna Book* cites 5k and there may have been a design change.) A value of 1.2k between terminals 2 and 3 of a 7805 IC establishes the 10-mA current level. This turned out to be a very convenient value, as it causes the voltage across the rheostat to vary between 0-10 volts.

Next, to preserve linearity, a high-impedance 10 Vdc meter is needed; I borrowed a simple vtvm circuit using a single fet from Radio Shack's book *Transistor Projects Vol. 2*. Using a 1-mA meter and a 9-volt battery, this unit reads full scale at 10 volts.

indicator circuit

The complete indicator circuit is given in **fig. 1**. The spdt center-off switch for rotation could be replaced by push-button switches. All parts for this indicator are available from Radio Shack stores. Note the arrangement for zero and full-scale adjust of the vtvm. This can easily be dispensed with, as the calibration of the vtvm is quite stable. The vtvm isn't difficult to recalibrate with a short circuit for zero and a 1000-ohm resistor for full scale. A semi-adjustable pot is used with the regulator IC to set current at precisely 10 mA.

Fig. 1 also includes circuitry for a control and indicator for a second position. This could be multiplied for a third or more positions. For the indicator at the

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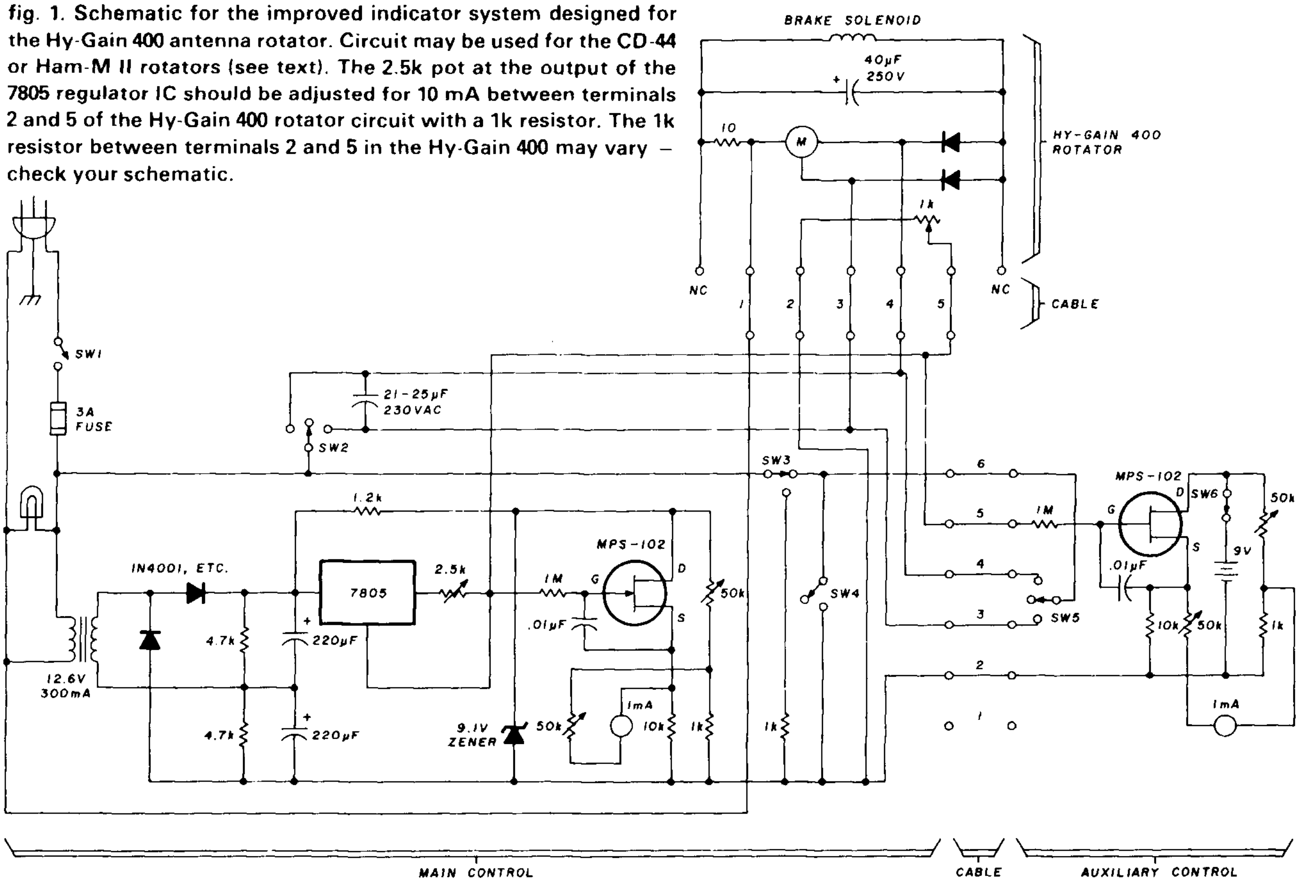
fig. 1. Schematic for the improved indicator system designed for the Hy-Gain 400 antenna rotator. Circuit may be used for the CD-44 or Ham-M II rotators (see text). The 2.5k pot at the output of the 7805 regulator IC should be adjusted for 10 mA between terminals 2 and 5 of the Hy-Gain 400 rotator circuit with a 1k resistor. The 1k resistor between terminals 2 and 5 in the Hy-Gain 400 may vary — check your schematic.



quantity	part
1	Hy-Gain 400 rotator
1	transformer 115- 12.6 Vac at 8.3 amp RS 273 1385 (a 24-Vac transformer could be substituted with a bridge rectifier and single filter capacitor)
2	diodes 1N4001 to 1N4007 RS 276 1101
1	voltage regulator IC such as 7805, 7812, 78L12AC (the latter is cheapest as RS 276-801)
1	zener diode, 9.1 V 1 watt 1N7439 RS 276 562
1 (2)	fet transistor MPS 102 or RS 276 2028
1	semiadjustable potentiometer 2.5k RS 271 228
2 (4)	semiadjustable potentiometers 50k RS 271 219
2	resistors 2.7k
1	resistor 1.2k
1 (2)	resistors 1 meg
2 (3)	resistors 1k
1 (2)	resistors 10k
2	capacitors 220 μF 35 Vdc
1	capacitor 21-25 μF 230 Vac (can be taken from original Hy-Gain 400 control)
1 (2)	.01 mfd μF ceramic
1 (2)	milliameter 1 mA RS 22052
2	switch spst main power (SW1 & SW4)
3	switch spdt center neutral (SW2 & SW5) RS 275 325
1	switch spdt RS 275 326

Notes:

1. Quantities in parentheses are needed only if an auxiliary indicator is desired.
2. Part numbers preceded by RS are available from Radio Shack stores.

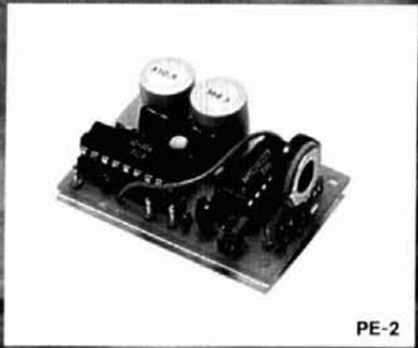
auxiliary positions, a 9-volt battery is practicable, although an ac supply may be preferred, especially by those who want a continuous reading.

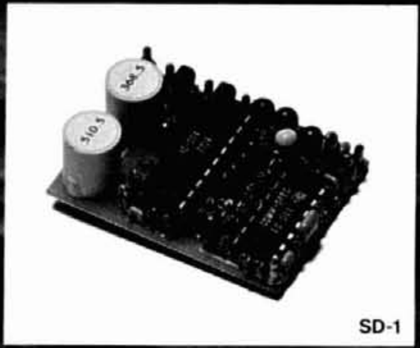
Note that the ac line that goes to the auxiliary control is *not* the ac line that goes to the rotator itself. Allocating the rotator ac to terminal 1 on the terminal switch and the auxiliary-unit ac to terminal 6 will help avoid confusion and fireworks.

other rotators

Some change in values might be needed if the circuit were to be adapted to the CD-44 or Ham-M II rotators, which have 500-ohm pots connected to terminals 3 and 7 of their rotator units with the sliding contact grounded and connected to terminal 1. I haven't tried the circuit on these units. Although it appears feasible, it would seem to be worthwhile only if multiple-position control were wanted, in which case the vtvm circuit shown here might be used to read the voltage between terminals 1 and 3, or 1 and 7 of the CD rotators. It also seems feasible to use the meter and case of the original control and put the few extra parts inside.

ham radio





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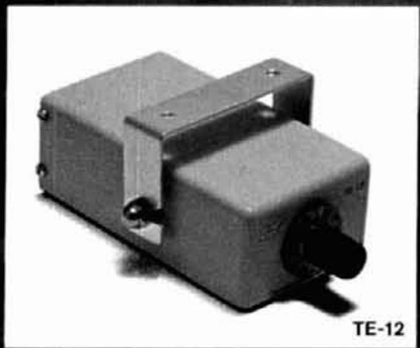
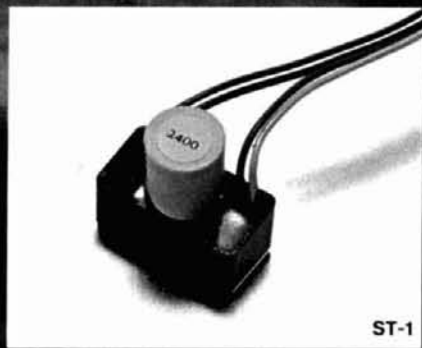

TE-12


ST-1

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The design philosophy behind the new Drake "7 system" has created a most sophisticated system concept, extending from engineering to the visual appearance of the system and each of its parts.

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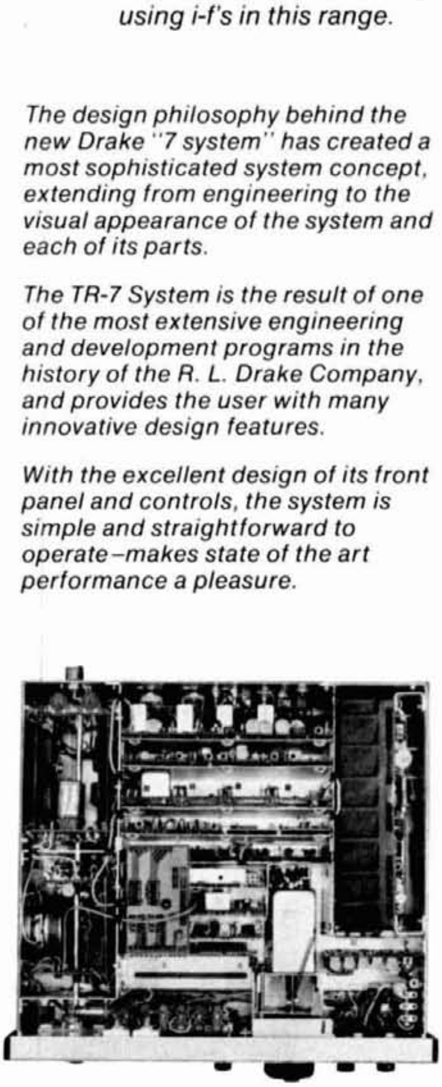

*Note: Out-of-band transmitter coverage for MARS, Government, etc. is available only in ranges authorized by the FCC, Military, or other government agency for a specific service. Proof of license for that service must be submitted to the R. L. Drake Company, including the 500 kHz range to be covered. Upon approval, and at the discretion of the R. L. Drake Company, a special range IC will be supplied for use with the Aux-7 Range Program Board. Prices quoted from the factory. See operator's manual for details.

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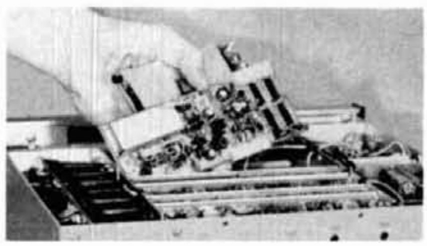
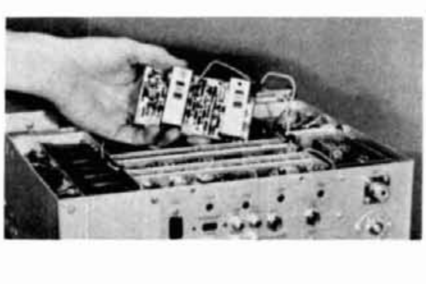
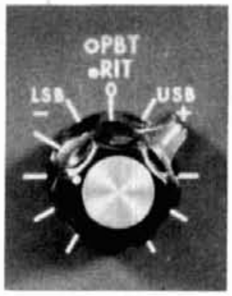


Continuous, Wide Range Frequency Coverage—The TR-7/DR-7 provides reception from 1.5 thru 30 MHz—continuously, and **zero** thru **30 MHz** continuously with the optional Aux-7 Range Program Board. The highly advanced Drake Synthesizer makes this possible, and is an industry first. The TR-7/DR-7 provides transmit coverage for all Amateur Bands 160 thru 10 meters. With the optional Aux-7 Range Program Board, diode-programmable out-of-band transmit coverage is available for MARS, Embassy, Government, and future band expansions in the range 1.5 thru 30 MHz.* The Aux-7 Board provides 0 thru 1.5 MHz receive coverage and crystal-controlled fixed channel operation for Government, Amateur, or semi-commercial applications anywhere in the hf range. The TR-7 w/o DR-7 and Aux-7 provides coverage of the Amateur Bands 160 thru 15 meters and the 28.5-29.0 MHz range of 10 meters. The Aux-7 Range Program Board is also useable in the standard TR-7 for extra range coverage as noted.



State of the Art Receiver Design—The Drake TR-7 introduces another industry first for amateur transceivers: "Up-Conversion," in combination with a special uhf high level double balanced mixer for superior strong signal handling, spurious and image response performance. The first i-f of 48.05 MHz places images well outside the receiver passband, and provides for true general coverage operation without i-f gaps.

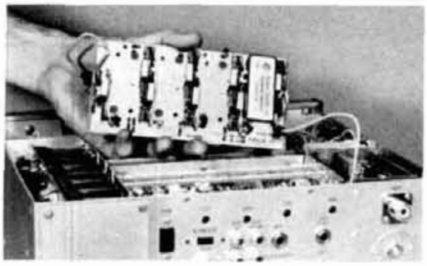
True Passband Tuning—The TR-7 employs the famous Drake Full Passband Tuning instead of the limited range "i-f shift" found in some other units. The Drake System tunes from the top edge of one sideband, through center, to the bottom edge of the other sideband. In fact, the range is even wider to accommodate RTTY. Full passband tuning greatly improves receiving performance in heavy QRM.



(TR-7 features continued on next page)

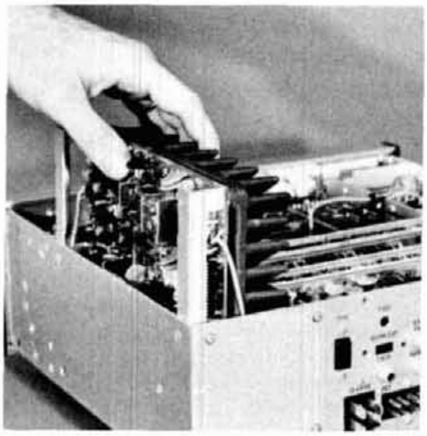
DRAKE TR-7 solid state continuous coverage synthesized hf system

Unique Independent Receive Selectivity—Optional receiving selectivity filters can be installed **internally** and pushbutton-selected from the front panel. These may be selected independently of transmit mode and provide optimum response for various conditions of ssb, cw, RTTY, and a-m. You may also transmit cw while receiving ssb, or vice versa, or even transmit one sideband while receiving the other. The standard filter is 2.3 kHz for ssb. You may choose from optional 300 Hz, 500 Hz, a special 1.8 kHz for crowded ssb, or 6 kHz filter for a-m.

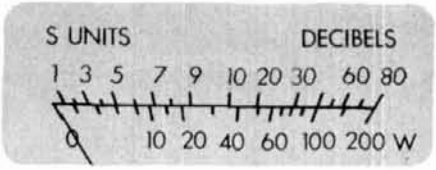


Effective Noise Blanker—This accessory is custom engineered to provide true blanking performance.

Special High Power Solid State PA—A Drake custom-designed diagonal heat sink provides for an internally mounted power amplifier with nothing mounted outboard subject to physical damage. The unique air ducting effect of this amplifier allows an optional rear-mounted fan to provide continuous duty on SSTV/RTTY. Continuous ssb/cw operation is available without the fan, due to the excellent heat sink design. The optional Drake PS-7 Ac Supply is rugged, rated for continuous duty, and will easily handle power requirements. The System is rated 250 watts input—in any of its modes. Fully VSWR protected.



TR-7 Internal Test Facilities—As well as the standard "S" meter function, the TR-7 metering includes a built-in rf Wattmeter/VSWR Bridge. Also, the DR-7 digital counter reads frequencies to **150 MHz** for test purposes. Access to the counter is from the rear panel.



Receiver Incremental Tuning (RIT)—Complete RIT flexibility is provided for **both** the TR-7 and RV-7 remote VFO for maximum convenience. The RV-7 also includes a special "spot" function for easy zero beating.

Model 1337	Drake TR-7 Transceiver	$ 886.00
Model 1530	Drake DR-7 General Coverage/ Digital Readout Board	186.00
Model 1336	Drake TR-7/DR-7 General Coverage Digital R/O Transceiver	1072.00
Model 1338	Drake RV-7 Remote VFO	156.00

Model 1502	Drake PS-7 120/240V Ac Supply includes special wide range voltage and frequency capability. Operates from any nominal line voltage (90-132 V/ 180-264 V; 50-60 Hz) ideal for overseas	$166.00
Model 1536	Drake Aux-7 Range Program Board	38.00
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Model 1335	Drake MMK-7 Mobile Mounting Kit	to be announced
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More Details? CHECK—OFF Page 150

calculating antenna bearings

for geostationary satellites

Using a pocket calculator to accurately determine antenna bearings and range to geosynchronous satellites

Many members of the amateur community have recently "discovered" the geosynchronous communications satellite. This surge of interest is in response to a program of weather satellites now affording an outstanding view of the earth from a 22,000 mile (35,000 km) vantage point.[1] It is further spurred by the recent availability of low cost, high quality microwave receiving equipment which enables the individual to recover not only this weather data,[2] but promises reception of direct-broadcast closed-circuit TV programs.

Although a great deal of material has been published on generating tracking information for polar orbiting spacecraft such as Oscar 7,[3,4] as well as the highly elliptical orbit planned for AMSAT Phase III,[5] little has appeared in the amateur magazines regarding the geostationary, or earth-synchronous, orbit. Peter Thompson has published a set of generalized equations which could be applied to tracking satellites in a variety of orbits,[6] but the calculations are unnecessarily cumbersome when considering the simple geostationary case. Ralph Taggart has outlined an appealing method for estimating azimuth and elevation bearings with a globe and string,[7] but like most graphical plotting techniques, this one offers resolution limited by the finite size of the globe. The equations presented here can be readily solved on a pocket calculator and afford bearing accuracy which is limited only by your ability to point the antenna.

*If the computed value for L is greater than 180°, correct it by subtracting 360°; if the computed value is less than −180°, correct it by adding 360°.

If an imaginary line is drawn which connects the center of a satellite with the center of the earth, that line intersects the earth's surface at a location known as the Sub-Satellite Point or SSP (**fig. 1**). Generating *azimuth* angles to a satellite is a function only of the location of the observer and the sub-satellite point, and is completely independent of the satellite's altitude (altitude does, of course, determine the elevation bearing, which is computed later). Hence, the azimuth data is based solely upon terrestrial coordinates; the law of cosines for determining distance and bearing coordinates in a great circle will apply.

Jerry Hall reported the relationships in his article on terrestrial great-circle computations;[8] those same equations may be used to calculate satellite azimuth angles:

$$\cos D = \sin A \sin B + \cos A \cos B \cos L \qquad (1)$$

$$\cos C = \frac{\sin B - \sin A \cos D}{\cos A \sin D} \qquad (2)$$

Where A = latitude of Point 1 (the observer)

B = latitude of Point 2 (the sub-satellite point)

L = longitude of Point 1 *minus* the longitude of Point 2*

D = distance between Point 1 and 2, degrees of arc

C = true bearing if L is positive

or $360 - C$ = true bearing if L is negative

Interestingly enough, the geostationary satellite presents a special case because the latitude of the sub-satellite point is always approximately zero degrees; under these conditions B in the above formulas is 0.

Since $\sin B = 0$ and $\cos B = 1$, these equations simplify to:

$$\cos D = \cos A \cos L \qquad (3)$$

$$\cos C = -\frac{\sin A \cos D}{\cos A \sin D} \qquad (4)$$

By trigonometric identities, **eq. 4** further simplifies to

$$\cos C = -\frac{\tan A}{\tan D} \qquad (5)$$

Thus, by solving **eqs. 3** and **5** for D and C, respectively, antenna azimuth information can be found.

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calculating elevation

If a straight line is drawn between a satellite and the center of the earth, and another straight line is used to connect the observer with the earth's center, the angle formed at the intersection of these two lines represents D, the distance between the observer and the sub-point, in degrees of arc, just computed (**fig. 2**). If the observer happened to be situated very near the center of the earth, angle D would relate directly to elevation angle; that is, D would be the angular *displacement from vertical* for aiming the antenna. Since elevation angles are generally specified with respect to the horizontal, *uncorrected* elevation data is computed from

$$EL\ (assumed) = 90 - D \qquad (6)$$

This assumes that the observer is located at or near the center of the earth. Since this is obviously impossible, let's consider corrections to the assumed elevation which would apply to an observer on the earth's surface.

If **eq. 6** is used, the error is negligible when R_s, the radius of the satellite's orbit, is at least two orders of magnitude greater than r_e, the radius of the observer's orbit (the radius of the earth). Thus, **eq. 6** is used in determining elevation information for radio astronomy, where the object being tracked is very far from earth. Under these conditions the observer *is* near the earth's center, *relative* to the object being tracked. **Eq. 6** has also been used by EME operators for tracking the moon; although the radius of the lunar orbit is only about 50 times the radius of the earth, elevation data calculated from **eq. 6** is correct to within a fraction of a degree.

It is interesting to note that in both of the above applications, the location of the object being tracked is generally specified in Greenwich Hour Angle (GHA) and declination. These coordinates correspond exactly to the longitude and latitude, respectively, of the sub-satellite point shown in **fig. 1**.

What do you do to correct **eq. 6** for geostationary satellites, whose orbital radius is *not* significantly

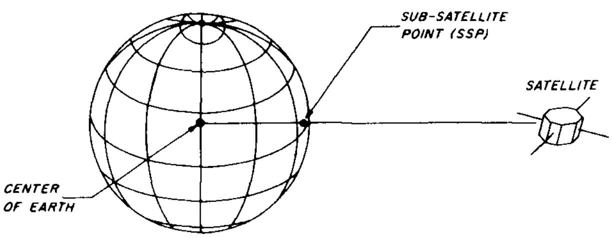

fig. 1. When a straight line is drawn from the center of the earth to a satellite, the intersection of the line with the earth's surface is known as the Sub-Satellite Point or SSP. The SSP is used to calculate the azimuth bearing to the satellite.

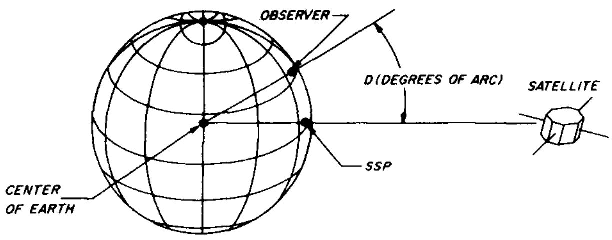

fig. 2. The observer's position on the earth's surface, in relation to the imaginary straight line shown in fig. 1, yields the angle *D*, which is used to calculate antenna elevation angle as discussed in the text.

greater than the radius of the earth? A correction formula published by L. R. Larson[9] applies:

$$tan\ EL = tan\ (90 - D) - \frac{1}{K\ cos\ (90 - D)} \qquad (7)$$

where K is the ratio of satellite orbital radius to earth radius

$$K = R_s / r_e$$

Since the mean radius of the earth is approximately 3444 nautical miles, and the mean radius of geostationary orbits is approximately 22,766 nautical miles, I use the value $K = 6.61$ when working with geostationary orbits. In the interest of minimizing computational steps, **eq. 7** may be restated

$$tan\ EL = \frac{sin\ (90 - D) - (1/K)}{cos\ (90 - D)} \qquad (8)$$

where $\frac{1}{K} = r_e / R_s = 0.1513 \qquad (9)$

Eq. 8 may be further simplified

$$tan\ EL = \frac{cos\ D - (1/K)}{\sqrt{1 - (cos\ D)^2}} \qquad (10)$$

calculating slant range

System performance predictions, which include link calculations of signal margin, require a knowlege of the exact distance from the ground station to the orbiting satellite. One convenient equation for determining slant range is stated as follows[10]

$$Range = R_s\ x\ \sqrt{1 - 2(1/K)\ cos\ D + (1/K)^2} \qquad (11)$$

If you want to calculate the range in nautical miles, use 22,766 nautical miles for R_s. If you would rather compute the range in kilometers, use 42,166 km for R_s. The ratio K, of course, is the same regardless of the units used to express R_s and r_e.

Example. My station is located approximately 37.3° N, 121.9° W;* I wish to receive signals from SMS-2,

*If you know your latitude and longitude to within 1/10 degree, you have pinpointed your location to within about 6 miles!

in geostationary orbit above 135° W. The longitude difference is thus $121.9-135=-13.1°$. Note that L is *negative*; this information will be used later.

The distance D is found from **eq. 3** to be 39.22°

From **eq. 5**, C is found to equal 158.99°. Since L is negative, true azimuth bearing is $360-C$, or 201.01°.

From **eq. 10**, the corrected elevation bearing equals 44.61°.

Eq. 11 yields a slant range of 20,215 nautical miles.

These computations agree well with my actual experience in tracking SMS-2.

calculator programs

The above relationships, with their conditional branching requirement depending on the sign of L, are ideal candidates for solution on a programmable calculator. I use a Hewlett-Packard 25 calculator which, with its 49 available program steps, demands a degree of programming ingenuity. After several false starts, I came up with the program listed in **table 1**, which appears to be valid for any point on the Earth*. Note that southern latitudes and eastern longitudes must be entered as *negative* numbers.

The following steps show how to use the program.

1. Key in the program, switch to RUN, depress f PGM.
2. Store the constant 180 in Register 3 (180 STO3)
3. Key in 3444 (for km, key in 6368)
4. ENTER
5. Store 22766 in Register 4 (for km store 42166)
6. Depress divide (÷) key
7. Store answer in Register 2
8. Store latitude of observer in Register 0
9. Depress *f* COS keys
10. Store answer in Register 1
11. Key in longitude of observer.
12. ENTER
13. Key in longitude of SSP.
14. Depress R/S key
15. Calculator displays range in nautical miles (or kilometers)
16. Depress rolldown key (R↓)
17. Calculator displays elevation angle in degrees
18. Depress RCL 6

*In reviewing this manuscript WB8DQT observed that the program will display ERROR if the observer is located directly under the satellite (on the equator at the SSP). Under these conditions the required azimuth bearing is undefined (the bearing is 90 degrees and the slant range is equal to $R_s - r_e = 19{,}322$ *nautical miles* [35,798 *km*]).

HP-25 Program

SWITCH TO PRGM MODE, PRESS f PRGM, THEN KEY IN THE PROGRAM

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	REMARKS
00	//////////	//////////					
01	41	–					
02	23 05	STO 5					
03	14 06	TAN					
04	32	CHS					
05	24 00	RCL 0					
06	14 04	SIN					
07	71	÷					
08	15 06	TAN⁻¹					
09	24 03	RCL 3					
10	51	+					
11	23 06	STO 6					
12	24 00	RCL 0					
13	15 51	x ≥ 0					
14	13 17	GTO 17					
15	24 03	RCL 3					
16	23 51 06	STO+6					
17	24 06	RCL 6					
18	24 03	RCL 3					
19	02	2					
20	61	X					
21	14 51	x ≥ y					
22	13 24	GTO 24					
23	23 41 06	STO–6					
24	24 05	RCL 5					
25	14 05	COS					
26	23 61 01	STO X1					
27	24 01	RCL 1					
28	24 02	RCL 2					
29	41	–					
30	01	1					
31	24 01	RCL 1					
32	15 02	x^2					
33	41	–					
34	14 02	$\sqrt{x}$					
35	71	÷					
36	15 06	TAN⁻¹					
37	01	1					
38	24 02	RCL 2					
39	24 01	RCL 1					
40	61	X					
41	02	2					
42	61	X					
43	41	–					
44	24 02	RCL 2					
45	15 02	x^2					
46	51	+					
47	14 02	$\sqrt{x}$					
48	24 04	RCL 4					
49	61	X					

MEMORY REGISTERS
R0 Latitude
of
observer
R1
D
R2 0.1513
(1/K)
(re/Rs)
R3
180
R4 Rs
(22766 naut
miles)
R5 Longitude of
observer minus
Long. of sat.
R6
Azimuth
R7

table 1. HP-25 program for calculating antenna azimuth and elevation angles and slant range to a geostationary satellite from any point on earth. Note that southern latitudes and eastern longitudes must be entered as negative numbers.

19. Calculator displays azimuth bearing in degrees
20. To perform further calculations, return to **Step 8**.

I would like to thank Glenn Thomas, WB6YZI, for his assistance in evaluating these computations. Glenn works at a satellite tracking facility and says, "Look-angles for geosynchronous satellites are my stock in trade." Believe me, he really makes it all look ridiculously easy!

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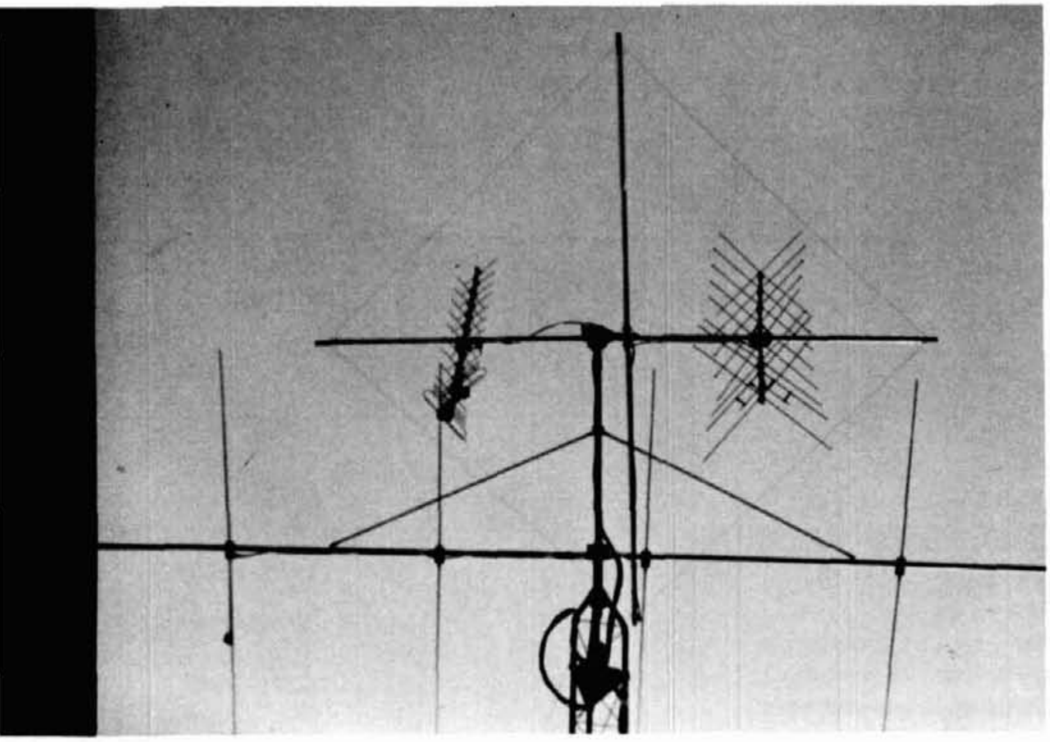
ham radio



Oscar az-el antenna system

Construction details for a complete antenna system for tracking Oscar including antennas for 28, 144 and 432 MHz

After checking all your charts and graphs, and double-checking all the calculations, you're left with only one problem, how to point your antenna at the satellite, and how to keep it accurately pointed as the bird moves through the sky. Obviously, some type of antenna tracking system is very desirable. Unfortunately, however, a check of commercial manufacturers did not yield any system that provided full frequency capabilities for working Oscar (the lack of 10-meter receiving antennas was the biggest stumbling block). After talking with many other amateurs, I decided to build my own tracking antenna system.

One of the methods used by others used some form of counterbalance to offset the antenna weight, permitting the rotator to elevate the antennas with a minimum of torque. This type of system tends to be impractical, however, because you should try to eliminate as much weight and wind resistance as possible. Also, the counterbalance system can become a mechanical nightmare if you try to design it to track from horizon to horizon in elevation (180-degree coverage). With these thoughts in mind, I decided that my tracking system should have the following features:

1. Capabilities for 28, 144, and 432 MHz
2. Full horizon-to-horizon elevation tracking
3. Cross polarization on both up and down links to prevent signal nulls
4. Elimination of the counterweight

By Al Greenwood, WA1NXP, RR1, Greenville Road, Mason, New Hampshire 03048

construction details

Cross frame. *The cross frame is the heart of the entire* antenna system, supporting both the 144- and 432-MHz beams and also acting as spreaders for the 10-meter quad loop. The basic layout of this assembly is shown in **fig. 1**. The horizontal and vertical members are made from two lengths of aluminum tubing, each with an outside diameter of 35 mm (1-3/8 inches) and 3.7 meters (12 feet) long. The plate which holds the two spreaders together can either be purchased ready-made* or built from 6.5 mm (1/4 inch) thick aluminum plate.

Phenolic insulators are inserted into the ends of the tubing to insulate the quad loop from the spreaders (see **fig. 2**). Each insulator is held in place with a *1/4-20 (M6) bolt through the tubing and insulator.* Another 6.5 mm (1/4 inch) hole is drilled through the insulator to hold the quad-loop wire.

After the two spreaders are fastened *together and* the insulators inserted in the tubing, the horizontal spreader is inserted into the U-100 rotator. Before tightening the clamps, you should run the rotator

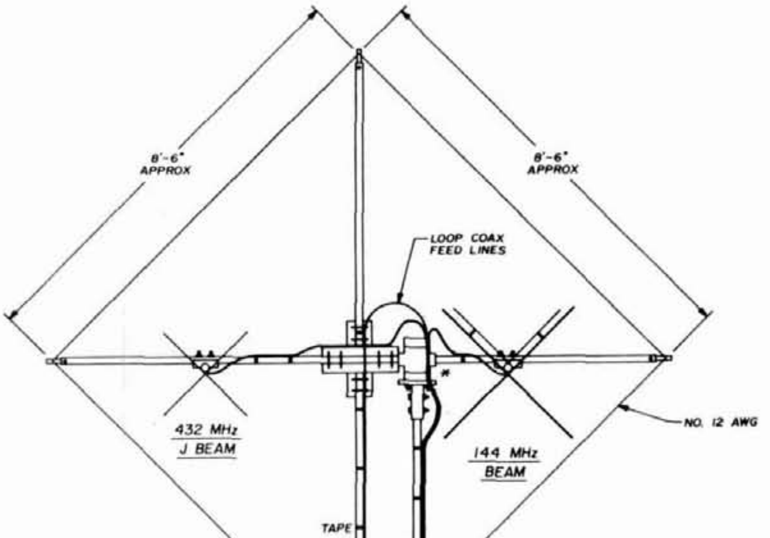
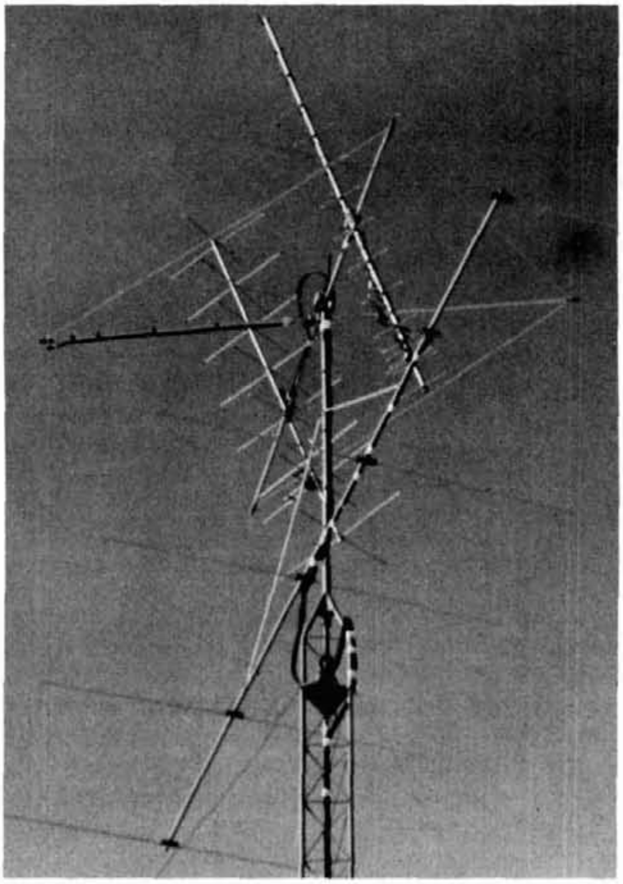


fig. 1. *Head-on view of the az-el antenna system.* The 144- and 432-MHz beams should be positioned along the horizontal spreader to balance the array at the base of the mast. The quad loop is made from number 12 AWG (2.1-mm) wire.



Satellite tracking antenna at WA1NXP combines antennas for 10 meters, 144 MHz, and 435 MHz on one tower. The six-element beam located below the OSCAR antenna is for 50 MHz.

from stop to stop to make sure that you have it in the desired position for assembly.

The final part of the cross frame is the adapter between the elevation rotator and the mast from the azimuth rotator. This was fashioned by drilling a piece of 6.5 mm (1/4 inch) *thick steel plate with four* holes that matched the bolt pattern of the U-100 rotator. To the bottom of the steel plate I welded a 15 cm (6 inch) piece of steel tubing. The inside diameter of the tubing must be large enough to accommodate the mast from the azimuth rotator.

Two-Meter Antenna. The 144-MHz cross-polarized Yagi consists of two 8-element beams mounted on a common boom (elements 90 degrees apart). The dimensions are standard and can be found in any antenna manual. The elements are mounted through a 32 mm (1-1/4 inch) aluminum boom, with the two beams displaced 12.5 mm (1/2 inch) along the boom.

*A parts kit including X bracket, spreaders, phenolic insulators, U-100 welded mounting bracket, coax bracket, and wire is available from Alden Engineering Company, P.O. Box 493, Greenville, New Hampshire 03048.

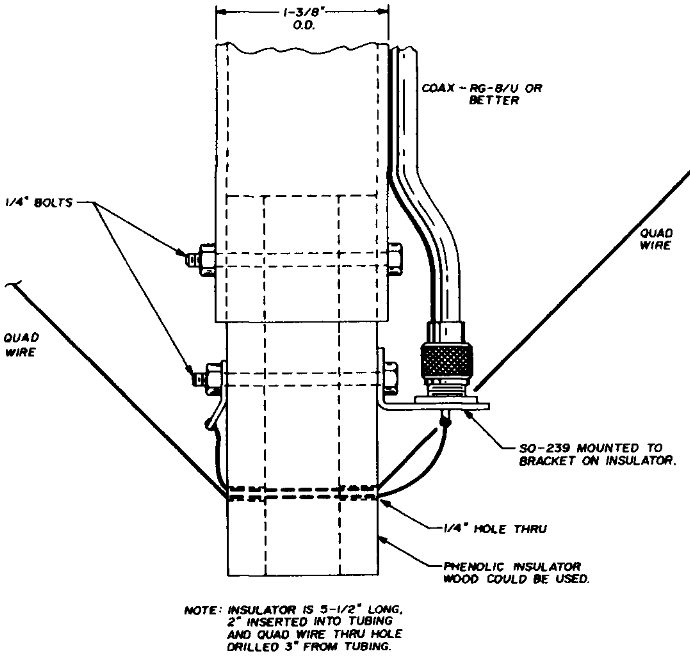


fig. 2. Detail of the four phenolic insulators which are used to support the 10-meter quad loop. The insulator is 14 cm (5-1/2 inches) long; the wire passes through the insulator approximately 7.5 cm (3 inches) from the end of the tubing.

The feed system for the beams is shown in **fig. 3**. The capacitor used for the gamma match is a split stator unit with approximately 50 pF per section. Be sure to provide an insulated mount for this capacitor. Prior to mounting the beam on the horizontal spreader, the longitudinal balance point of the beam must be determined. By attaching at this point, the elevation rotator will not be excessively strained.

Seventy-Centimeter Antenna. For 432 MHz, I

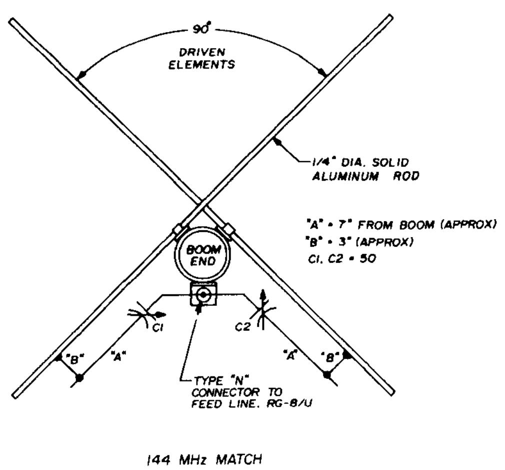


fig. 3. Gamma match for the two 144-MHz beams. The gamma capacitor is a split-stator capacitor with approximately 50 pF per section. After final adjustment, the entire unit was fiberglassed for weather proofing.

used a commercial 26-element *J Beam*. As with the two-meter beam, this antenna is also mounted to provide cross polarization. **Fig. 4** shows the baluns and the matching necessary to simultaneously feed both antennas from one 50-ohm feedline. You should also determine the balance point for this beam before attaching it to the horizontal spreader.

Ten-Meter Antenna. As a final step, the wire for the quad loop is strung through the holes in the phenolic insulators. With the dimensions shown, the loop should resonate at approximately 29.6 MHz. When using this system, the pattern exhibited by the loop tends to be sharper than would be expected, or desired. As a remedy, I opened the loop at both ends of the horizontal spreader, effectively producing a dipole which has a wider beam-width than the loop.

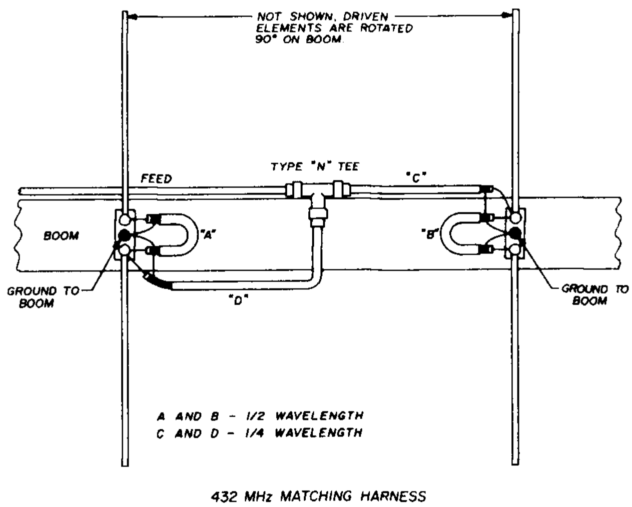


fig. 4. The two 432-MHz beams require a balanced 300-ohm feed. With this arrangement, the 1/2-wavelengths of coax act as 4:1 baluns, while the 1/4-wave sections step the impedance up to approximately 100 ohms. The two beams then closely match a 50-ohm feedline.

Either version can be used; it depends only on your personal preference and the resolution of your azimuth rotator.

summary

In the described configuration, the U-100 rotator easily moves the array through a full 180 degrees elevation. As an added feature, the quad wire acts as a mechanical stop when it hits the azimuth rotator mast, stalling the elevation rotator. One drawback to this system is the resolution of the rotators. For extremely accurate pointing, infinite control is a necessity. Unfortunately, this is not provided by either rotator.

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high-gain 1296-MHz antenna

For those interested in an easy-to-build antenna for the 1215-1300 MHz amateur band, the design described here might be the answer. It features high gain, good capture area, and shielding from high-power local signals. The design also includes both vertical and horizontal polarization and, with appropriate phasing equipment, right and left circular polarization. The simplicity of the feed system and overall antenna design should make it a good candidate for just about any requirement in the 1215-1300-MHz band.

I have received signals as far as 80 km (280 miles) away with this antenna. With a motorized tilt control, the antenna has possibilities for use in amateur satellite work.

The design features a waveguide-type can, which bypasses strong local signals and shields direct radiation interference from the 1/4-wavelength dipoles. The addition of 10-degree tubes doubles the gain, concentrates the signal, and increases signal capture area.

construction

The antenna design is shown in **fig. 1**. To simplify matters, I had a local tinsmith make a 178 mm (7 inch) ID x 406 mm (16 inch) long round can of 24-gauge galvanized tin, with a solid back, crimped and soldered. (Brass or copper could be used.)

Four aluminum tubes are required, measuring 12.7 mm (1/2 inch) ID x 16 mm (5/8 inch) OD x 140 cm (55 inches) long. Don't bend the tubes and don't substitute any other diameter or length.

Next, you'll need four lengths of galvanized-steel water pipe. The dimensions are 6.4 mm (1/4 inch) diameter x 457 mm (18 inches) long. Bend at the 305-mm (12-inch) point (sharp bend) at a 10-degree angle (**fig. 1**). Aluminum tubing measuring 13 mm (1/2 inch) ID x 16 mm (5/8 inch) slips snugly over the extending 152-mm (6 inch) portion of the water pipe. The aluminum tubes are secured by a self-tapping screw, one screw per tube.

Drill a 9.5-mm (3/8 inch) hole in the top, and drill another hole 90-degrees away for the TV F-type coaxial fittings. The holes should be drilled 54 mm (2 1/8 inch) from the back of the can. Two 54 mm (2 1/8 inch) lengths of 1-mm (18-AWG) wire should be soldered to the TV F-type fittings for the dipole radiators.

I obtained a piece of steel measuring 1.6 x 102 x 229 mm (3/16 x 4 x 9 inches) from a local metal shop, which I used to mount the antenna to a 25-mm (1-inch) diameter mast made of galvanized-steel water pipe. A steel plate was drilled for three TV mast clamps. These clamps were placed 51 mm (2 inches) apart. **Fig. 1** gives the

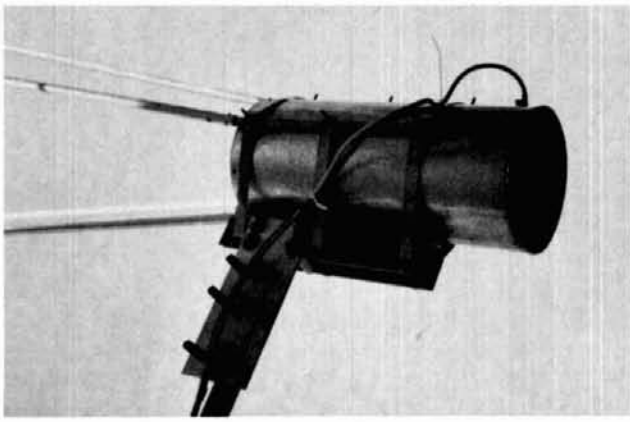
The can has a piece of clear plastic measuring 178 x 6.4 mm (7 x 1/4 inch) mounted snugly in the open end. See **fig. 1**. Drain holes should be drilled into the bottom of the can.

The TV-type chimney mast straps are used for reinforcement, as shown in **fig. 1**. These reinforcement straps are absolutely necessary.

When, completed, the distance between the ends of each set of tubing (horizontal or vertical) should be between 686 and 711 mm (27 and 28 inches).

weather protection

I used polyurethane liquid plastic coating for weather protection. However, almost any good spray enamel could be used. Extra coats of paint on the steel plate are desirable to inhibit corrosion. (Don't paint the plastic cover for the cans.) Liberal use of PVC tape over the coaxial-cable fittings where they fasten to the cans will help to make waterproof joints.



Complete antenna showing the can-type resonator and feed-system details. Two coaxial cables are used.

ham radio

By Paul F. Magee, W3AED, RR2, Box 432, Berlin, Maryland 21811

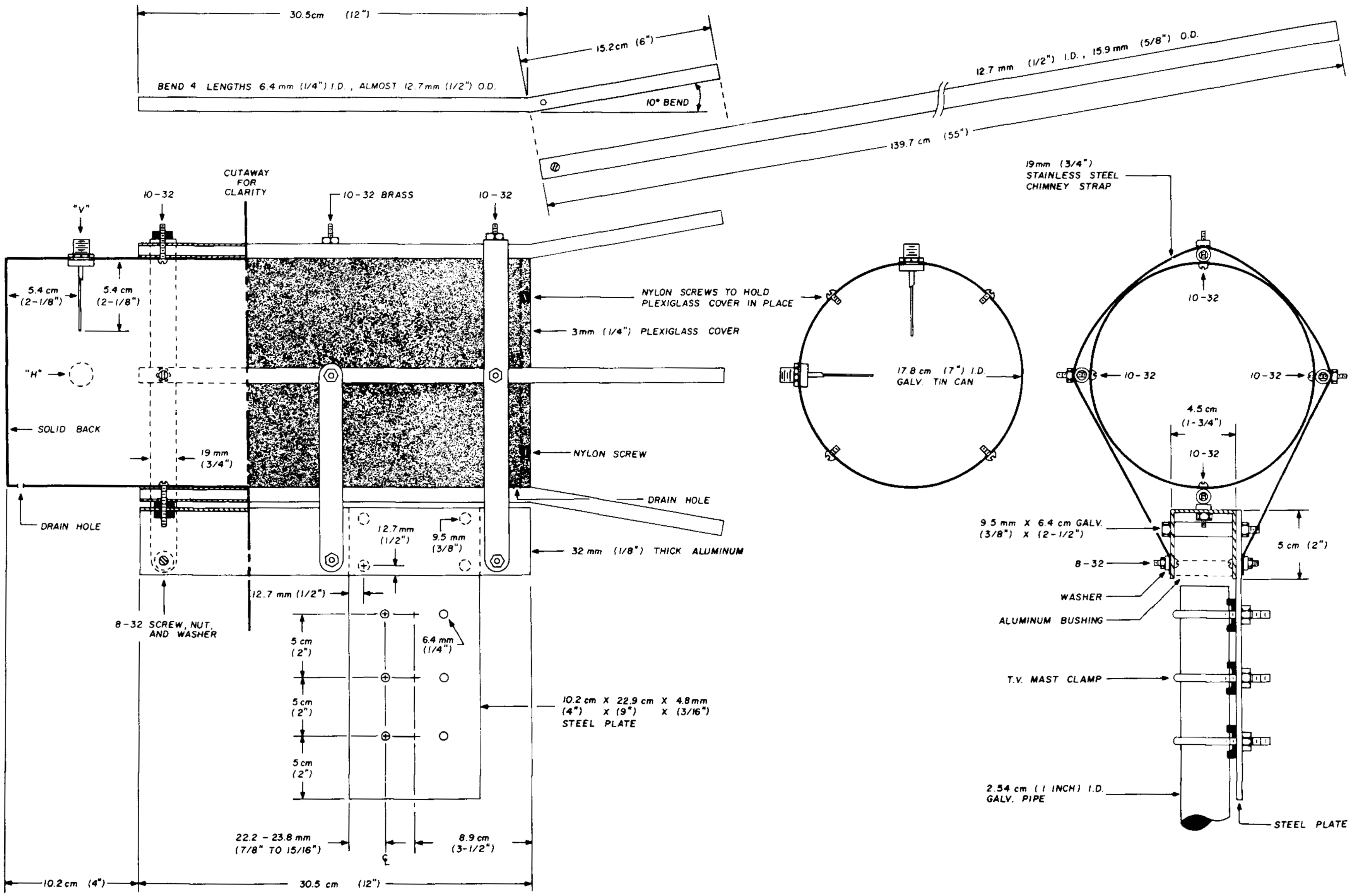


fig. 1. Construction details of the high-gain 1215-1300-MHz antenna. Design features both vertical and horizontal polarization and an enclosed, weather-proof feed system.

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Kenwood offers this totally unique unit as a perfect compliment to your TS-820S or TS-520S station.** The SM-220 permits you to monitor your transmitted signals, thus assuring optimum linearity and maximum performance. With the addition of the BS-5 or BS-8 Pan Display option you will be able to determine visually the location and strength of adjacent signals without tuning your receiver off frequency. The choice of options allows you to adapt the SM-220 to either the TS-820S or TS-520S, depending on which rig you now have or may acquire.

The SM-220 has a built-in two-tone audio generator with full provisions for tuning your exciter and linear amplifier (160 m through 2 m).

All this costs little more than a general-purpose oscilloscope. And, of course, it's pure Kenwood quality.

*With BS-5 or BS-8 option

**For other models check with appropriate manufacturer for compatibility.

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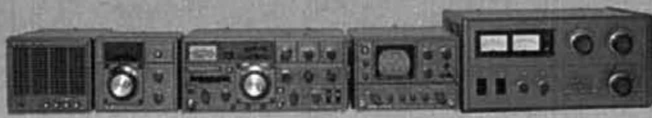
SM-220



Function: Selects operation mode: OSC/RTTY: General testing of station equipment, experimental design of new equipment, or troubleshooting; display of receiver IF output allows you to give "signal quality reports".

Power ON indicator.

Power switch.

Intensity: Controls brightness of 'scope display.

Band Scope: (Pan Display) With BS-5 or BS-8 option, allows you to "see" the signals on both sides of your operating frequency without tuning your receiver off frequency. Useful for determining "band conditions", band crowding, source of interference from adjacent stations . . . a visual display of what you would hear if you tuned across the band, without having to touch your receiver's dial.

Focus: Controls sharpness of 'scope display.

Vertical Attenuator: Precision step attenuator (gain control) switch adjusts vertical input level.

Vertical Input: Accepts IF input, RTTY input or oscilloscope input.

Vertical Gain: Potentiometer to fine-adjust vertical input level.

▲▼: Adjusts display along vertical axis.

◄► Adjusts display along horizontal axis.

Sweep Range: Step switch controls sweep band width or switches horizontal input/external sync terminal "ON".

RF Attenuator: Level control used in MONI/TRAP mode.

Tone: Step switch selects Wien bridge tone generators, 1000 Hz, 1575 Hz or both tones simultaneously.

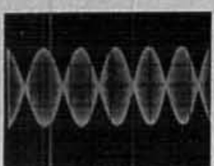
Out: Output of the audio generator can be connected to the transceiver's microphone input for "two-tone test". Also for trapezoidal test of transceiver linear amplifier.

Synchronization Marker: Selects internal or external sync (similar to horizontal hold on TV. Turns On or Off the built-in marker which shows operator where his receiver is actually tuned.

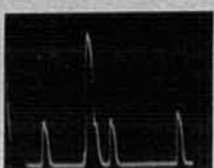
Scan Width: Selects width of "window" or receive band display when using the Pan Display option. (100 kHz or 20 kHz)

Variable sweep control/External gain: Controls (1) sweep speed of display in any sweep range. (2) optional Pan Display (Band Scope) speed of display, (3) level of horizontal input/external syncronization input when sweep range is in RTTY/Ext or Trap.

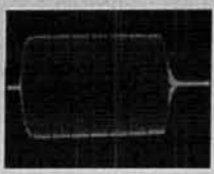
Horizontal Input/External Sync: Accepts either (1) RTTY input for tuning, (2) external sync input for test (oscilloscope functions), (3) external oscillator for Lissajous display.



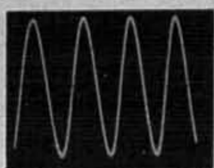
Two-Tone Wave Envelope For "performance" tune-ups or checking proper tranceiver operation.



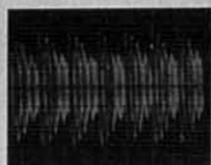
Pan Display Use to check source of interference during "QSO" without moving off-frequency. Also determines location and strength of adjacent frequencies. (Requires BS-5 or BS-8 option)



Keyed Waveform Shows detail of CW keying. Use to monitor the quality of your CW note. (Photo shows ideal waveform produced by TS-820S.)



Oscilloscope Operation (1 kHz) Oscillator function allows Sine, square wave, Lissajous patterns for testing or design work.



Trapezoid (TS-820S w/ TL-922) Shows linearity of power amplifier. Used primarily for testing.

Wave Envelope shows full SSB voice modulation, with processor on (full compression), and "clean signal" at full power.

arma

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simplified antenna gain calculations

A straightforward approach for comparing the gain of a number of popular amateur antennas

It is well recognized by amateurs that improvements in an antenna installation can usually do more to enhance a station's capabilities than any other improvement. The purpose of this article is to help you choose or optimize an antenna arrangement from the standpoint of gain. Of course, many other factors have a large influence on antenna performance including ground conditions, ohmic losses, height above ground, polarization, bandwidth, multiband operation, effects of nearby objects, and so on. This article will not deal with these topics, but will attempt to consolidate some results of gain calculations made on some of the more popular antennas in use today. Along the way, a relatively straightforward approach for making such calculations will be described so that you can make similar calculations for configurations in which you are interested. It should be observed, however, that the methods described are applicable to only some antenna types. For example, multi-element arrays with parasitic elements are not suited to the simple technique described here. However, this technique does allow you to gain a significant amount of insight into a number of different antennas.

basic assumptions

As it turns out, very few assumptions must be made to compare the gains of the antenna types which are considered here. These assumptions are listed below:

1. The feedpoint is at a point of maximum current.

2. The feedpoint resistance (at resonance) is known.

3. All power fed into the antenna is radiated.

4. The current distribution for wire resonant antennas is sinusoidal and for antennas containing inductive loading is approximately trapezoidal.

5. The current in each part of the antenna is either in phase or 180° out of phase with the current in all other parts of the antenna, with phase reversals occurring where the current goes through a

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minimum. Although this assumption is not 100 per cent valid,[1] the results should be close enough to reality to provide insight for those of us who enjoy experimenting with different antennas.

All gain calculations will be made relative to a half-wavelength dipole. With the exception of the monopole example given later, the antennas will be assumed to be in free space.*

Although the calculations are made for transmitting antennas, the results apply equally to receiving because of the principle of reciprocity.

theory

The field strength due to a part of an antenna is proportional to the magnitude of the current flowing in that part. This dependence is shown in **fig. 1**. It is assumed, of course, that the observer is many wavelengths from the antenna; all parts of each of the antennas considered here will be essentially the same distance from the observer. By "essentially the same" I mean that any differences in distance from the observer are very small compared to a wavelength. A surprisingly large number of practical antennas exhibit maximum gain in a direction where this distance constraint is valid.

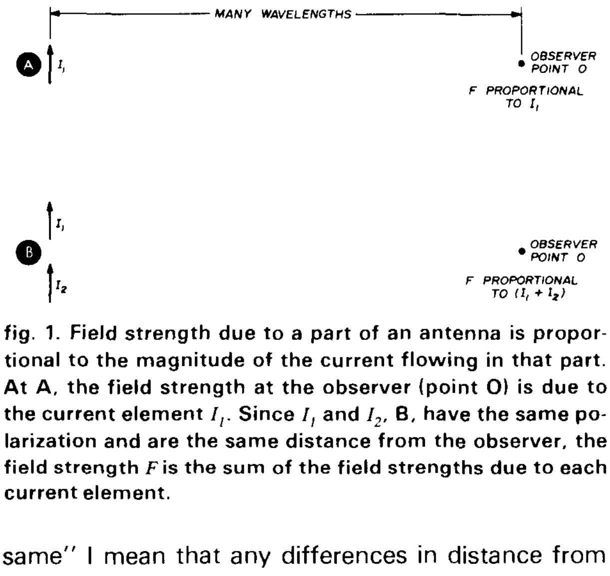


fig. 1. Field strength due to a part of an antenna is proportional to the magnitude of the current flowing in that part. At A, the field strength at the observer (point O) is due to the current element I_1. Since I_1 and I_2, B, have the same polarization and are the same distance from the observer, the field strength F is the sum of the field strengths due to each current element.

Another simplifying assumption is that in the direction for which the gain is to be calculated some radiation cancellation takes place, as seen in **fig. 2**.

*An assumption of free-space conditions simplifies the calculations, but many would argue that it's both unrealistic and misleading to discuss the radiation characteristics of either straight or bent dipoles for use in a ground-oriented communications link if the effects of ground reflections are ignored. Nevertheless, the gain technique discussed by the author provides a basis of antenna comparison. It should be kept in mind, however, that the gain and radiation patterns presented here are not necessarily representative of those found in a real life, near ground situation. **Editor**.

feedpoint resistance

To determine the relative magnitude of the current flowing in an antenna, the feedpoint resistance must be either assumed, measured, or calculated. Through the use of an impedance bridge, using either an oscillator or noise generator as a source,

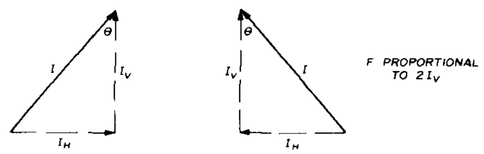


fig. 2. The observer in this case is looking into the plane of the page. Note that the vertical components, I_V, add while the horizontal components, I_H, cancel. Field strength F is proportional to $2I_V$.

quite accurate measurements of feedpoint resistance may be made (see references 2 through 4). Anyone who is seriously interested in optimizing an antenna installation should have a means for measuring feedpoint resistance.

Direct calculation of antenna feedpoint resistance is a complex mathematical process except for the simplest of structures. The engineering literature yields some useful results, but much more work needs to be done in this area.

Fig. 3 shows how the currents are related for two antennas with different feedpoint resistances when you assume that the same amount of power is fed into each antenna. Thus G_R, that part of gain due to feedpoint resistance, is given by

$$G_R(dB) = 20 \log \frac{I_2}{I_1} = 20 \log \sqrt{\frac{73}{R}} = 10 \log \frac{73}{R}$$

where R is the feedpoint resistance of the antenna under consideration (73 ohms is the free-space feedpoint resistance of a half-wavelength dipole).

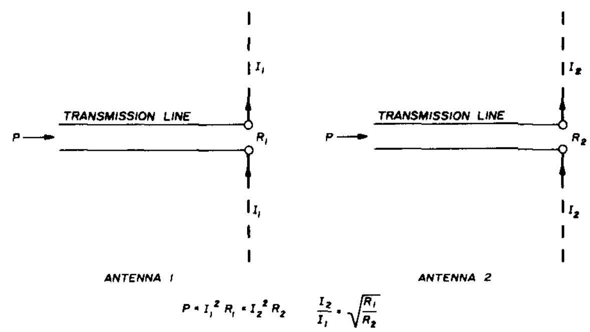


fig. 3. Effect of feedpoint resistance on current when the same power P is fed to two different antennas.

current distribution

The currents which affect the field strength vary along the length of the antenna. At the open ends of a wire antenna (e.g., the ends of a dipole), the current is practically zero. It is customary to assume that the current varies along a resonant antenna sinusoidally and this will be done here; **fig. 4** shows the assumed current distribution along a half-wavelength dipole.

Since it will arise frequently, it's useful to compute the field strength due to a sinusoidal current distribution along a straight wire. **Fig. 5** shows the general result for the field strength due to a section of antenna wire of length x_l with a sinusoidal current distribution as shown. For a half-wavelength dipole, $x_l = \lambda/4$, and since the fields radiated from the two

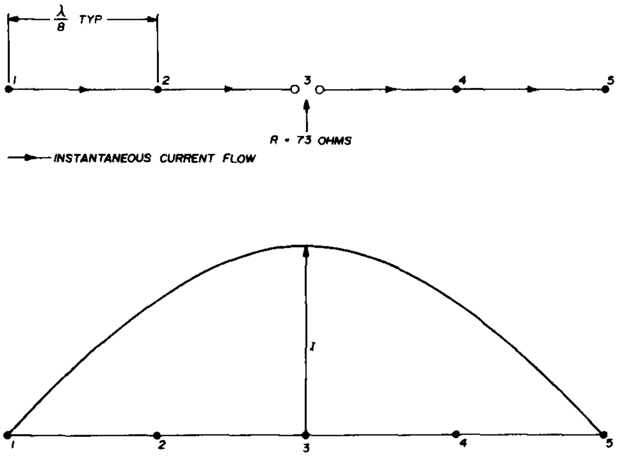


fig. 4. Current distribution along a half-wavelength dipole.

halves add in phase (in the plane normal to the radiator, at its center)*

$$F_R = 2\frac{k\lambda I}{2\pi}\ sin\ \frac{2\pi}{\lambda} \bullet \frac{\lambda}{4} = \frac{k\lambda I}{\pi}$$

This is the reference field strength due to current distribution for gain calculations. G_C is defined as the part of the gain due to current distribution and is given by

$$G_C\,(dB) = 20\ log\ \frac{F}{F_R} = 20\ log\ \frac{\pi F}{k\lambda I}$$

where F is the field strength of the antenna under consideration. Thus the gain relative to a dipole in dB is given by

$$G = G_R + G_C = 10\ log\ \frac{73}{R} + 20\ log\ \frac{\pi F}{k\lambda I}$$

*Off the plane normal to the radiator the fields add out of phase, to a degree related to the angle off the plane, resulting in the dipole radiation pattern.
Editor.

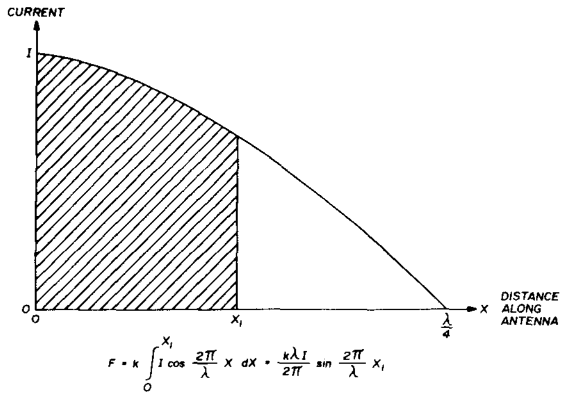


fig. 5. Field strength, *F*, due to a sinusoidal current distribution; *k* is a constant of proportionality.

Since k, λ, and I will cancel in the subsequent calculations, it is clear that if R and F can be determined, the gain relative to a dipole may be found.

Another current distribution of interest is the trapezoidal distribution, which is a good approximation for inductively-loaded, short antennas.[5] The only antenna having this current distribution that will be considered here is the quarter-wavelength dipole with loading coils located as shown in **fig. 6.**

other factors

As seen in **fig. 2**, in some cases the current in the direction of the wire is not the appropriate current to use. What is required is the current in the direction of the polarization being considered. Hence, for **fig. 2**, the current component of interest is

$$I_V = I\ cos\ \theta$$

Another consideration is the number of radiating elements. In this case, you must account for the fact

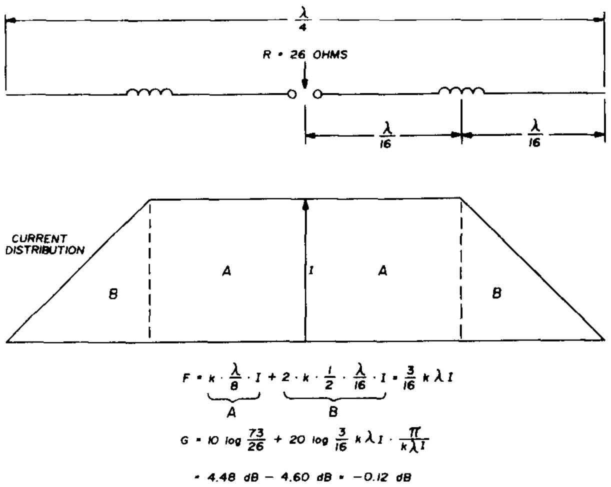


fig. 6. Gain of a quarter-wavelength inductively-loaded dipole.

that the power is divided among these elements. For the two-element broadside and collinear arrays considered here, the power is equally divided between the two radiators. Hence, the multi-element factor G_M, which must be included in the gain calculations, for the two-element case is

$$G_M = 10 \log \frac{1}{2} = -3.01\ dB$$

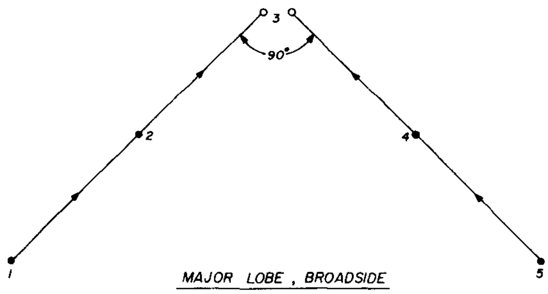
With these preliminaries disposed of, we can now consider several specific antenna types.

inductively-loaded dipole

Consider the inductively loaded dipole shown in **fig. 6**. The feedpoint resistance is assumed to be 26 ohms.[5,6] Further verification of this value is required, however. The fact that the gain is essentially that of a full-size dipole is consistent with the popularity of this antenna for many sloping-dipole installations on 80 meters. However, it is hampered somewhat by its relatively narrow bandwidth.

inverted vee

Fig. 7 shows the instantaneous currents in a dipole having legs formed as an inverted-V. When the legs form an angle less than 180 degrees, the radiation resistance is reduced, resulting in a corresponding reduction in feedpoint resistance.[7,8] However, the current distribution is practically the same as that for the straight dipole of **fig. 4**. In the broadside direction the fields resulting from the vertical current components cancel; however, the horizontal field components add, but they are diminished by the folding. On the other hand, this reduction is almost completely compensated for by the overall increase in magnitude of *all* the field components, resulting from the higher antenna current I, due to the lower feedpoint resistance. When the included angle is 90°, the feedpoint resistance in free space is typically around 40 ohms, and the gain relative to a straight, 180-degree dipole is approximately −0.4 dB.



$F = F_R \cdot \cos 45° = \frac{F_R}{\sqrt{2}}$

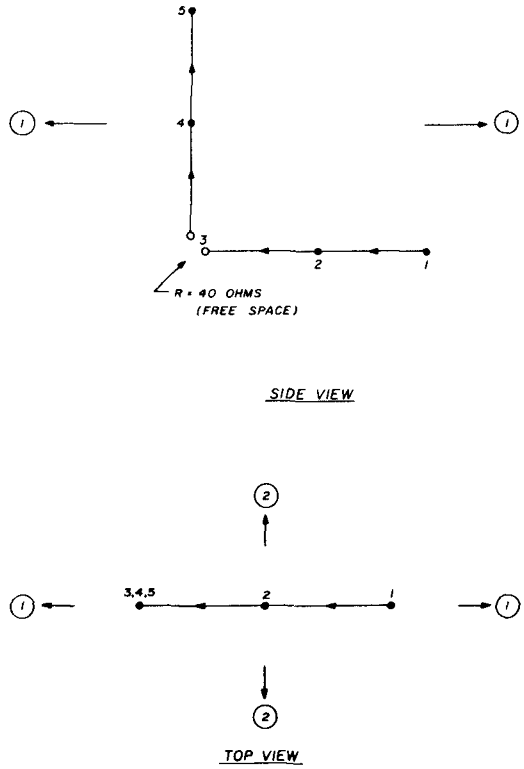
$G = 10 \log \frac{73}{40} + 20 \log \frac{1}{\sqrt{2}}$

$= 2.61\ dB - 3.01\ dB = -0.40\ dB$

fig. 7. Gain of the inverted-V dipole. Observer is facing the page, and polarization is horizontal. An observer in the plane of the paper, looking into the antenna from left to right would see a small component of vertical polarization due to uncancelled vertical components radiated from each leg. The vertical components cancel in the broadside direction.

L antenna

The L antenna shown in **fig. 8** is of interest because it has been used where height is limited. In direction **1** the polarization is vertical, but in direction **2**, where the gain is maximum, the polarization is half horizontal and half vertical. In direction **1** only half the antenna is seen by the observer and the gain is



$F_1 = \frac{F_R}{2}$

$G_1 = 10 \log \frac{73}{40} + 20 \log 1/2$

$= 2.61\ dB - 6.02\ dB = -3.41\ dB$

G_2 = SAME AS INVERTED V = −0.40 dB

fig. 8. Gain of the L antenna. All points on the equatorial plane are broadside to the antenna. In the polarization plane normal to the equatorial plane the gain is uniformly maximum everywhere on the equator. Maximum gain is in direction 2 and is the same as an inverted V with a 90 degree included angle.

−3.41 dB (free space). In direction **2**, the gain is the same as that for the inverted V, which is obvious when you remember that the L is a rotated V. In direction **3** the gain in only slightly less than in direction **2**, but will not be calculated here.

multi-element driven arrays

You can use the methods described here to determine the relationship between gain and feedpoint resistance for certain types of driven arrays. Such data are available in the ARRL *Antenna Book* for ideal situations, but the method given here will enable you to infer gain directly from feedpoint resistance measurements.

The cases considered are the collinear and broadside arrays, where two in-line or parallel half-wavelength dipoles are fed in phase as shown in **fig. 9**. The current distribution is the same as for the reference dipole but since there are two radiators, $G_C = 6.02$ dB. The power is split between the two radiators, however, so the field strength is modified by $G_M = -3.01$ dB. Thus, as seen in **fig. 9**, the gain

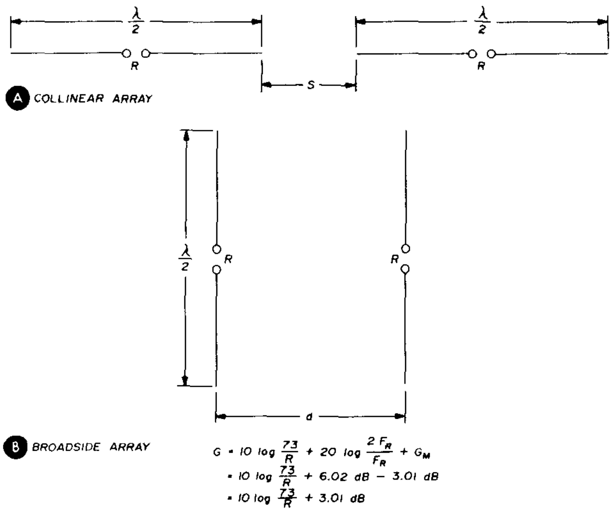


fig. 9. Gain calculation for two in-phase parallel half-wavelength dipoles in a collinear arrangement, A, or broadside array, B.

table 1. Calculated gain of collinear array.

spacing s	feedpoint resistance for each dipole R (ohms)	gain G (dB)
0	94	1.9
$\frac{\lambda}{8}$	87	2.2
$\frac{\lambda}{4}$	76	2.8
$\frac{\lambda}{2}$	70	3.2
λ	73	3.0

is related very simply to the feedpoint resistance by

$$G = 10 \log \frac{73}{R} + 3.01 \text{ dB}$$

This relationship is plotted in **fig. 10**.

The feedpoint resistance and gain for several spacings of collinear dipoles are shown in **table 1**. The resistance values were obtained from *The ARRL Antenna Book*[2] (page 134); the calculated gain values using **fig. 10** agree exactly with those given in the *Antenna Book*.

The *Antenna Book* does not give the feedpoint resistance for the broadside array, but it does give gain as a function of spacing. Again **fig. 10** was used, but this time to determine the feedpoint resistance of each dipole. The results, shown in **table 2**, agree very well with those given in reference 9.

monopoles

A popular vertical radiator is the monopole, or quarter-wavelength half-dipole over ground. At higher frequencies, resonant radials are used if the antenna is elevated well above ground, at least a quarter wavelength. At lower frequencies elaborate radial systems (usually buried a few inches) are often employed. Such a ground system is *required* if optimum radiating efficiency is to be obtained. An infinite, perfectly-conducting ground is assumed in **fig. 11**. The feedpoint resistance of the quarter-wavelength half-dipole, in conjunction with its ground-reflected image is 36.5 ohms, half that of a complete dipole. In this case, the total field strength results from the in-phase addition of the direct radiation from the monopole and the radiation from the ground image. The total field is therefore twice that radiated directly from the monopole. If the reference dipole antenna is

table 2. Calculated feedpoint resistance of the broadside array.

spacing d	feedpoint resistance for each dipole R (ohms)	gain G (dB)
0	146	0
$\frac{\lambda}{8}$	136	0.3
$\frac{\lambda}{4}$	116	1.0
$\frac{\lambda}{2}$	59	3.9
$\frac{3}{4}\lambda$	51	4.6
λ	77	2.8

considered to be in free space then it would appear from **fig. 11** that the monopole has +3.01 dB gain, but this is misleading. The reason for the apparent gain is that the monopole radiates the power only in the "half" space (or half hemisphere) above ground,

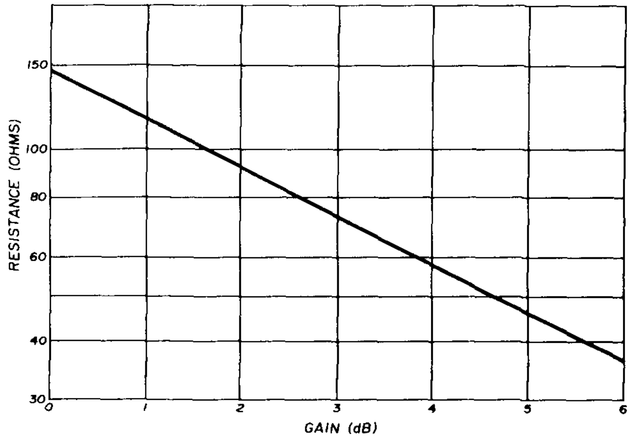


fig. 10. Feedpoint resistance versus gain for parallel half-wavelength dipoles fed in phase.

while a free-space dipole radiates the same amount of power in all space.

A similar calculation can be used to determine the gain of a half-wavelength ground-mounted vertical dipole. Since this antenna plus its image is like a collinear array with zero spacing, **table 1** may be used to obtain the feedpoint resistance of 94 ohms (the feedpoint is at the current maximum, one-quarter wavelength above ground). This calculation yields a gain figure of 4.92 dB over a reference half-wavelength dipole in free space. Thus a half-wavelength ground-mounted dipole has 1.91 dB gain over a monopole ($4.92\ dB - 3.01\ dB = 1.91\ dB$).

full-wave loops

The full-wavelength loop is used in the shape of a square (the quad loop), a diamond, and a triangle (the delta loop). In this section, the dependence of gain on the loop shape will be calculated. Since I don't know the exact feedpoint resistance for each of

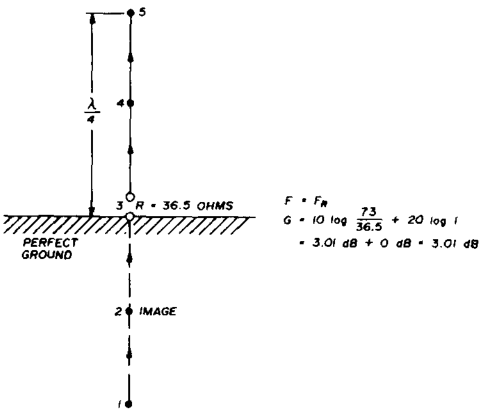


fig. 11. Gain of a monopole over perfectly conducting ground relative to a half-wavelength dipole in free space. The current distribution is the same as that shown in fig. 4.

these forms, a precise analysis is not possible. Some interesting results have been obtained, however, and some suggestions for further work have been identified.

Fig. 12 shows the current distribution on a full-wavelength loop and two special cases. The shorted half-wavelength transmission line will not accept power because $R = 0$, so it is of no interest as an antenna. **Fig. 12C** proves that it is consistent for a

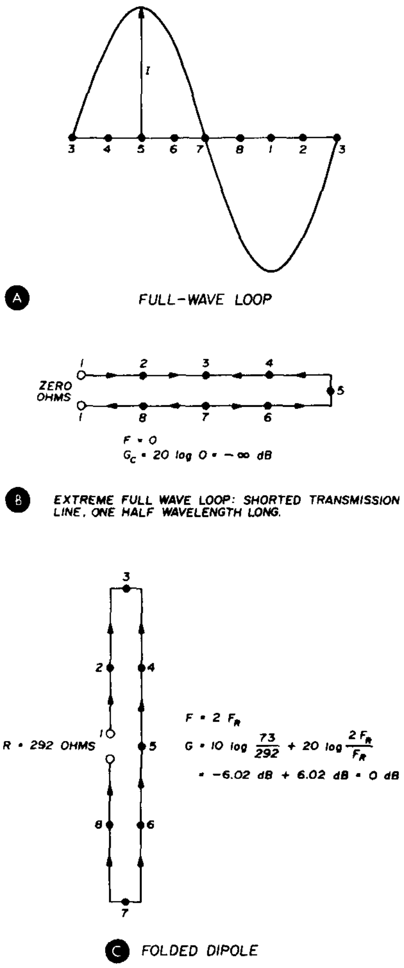
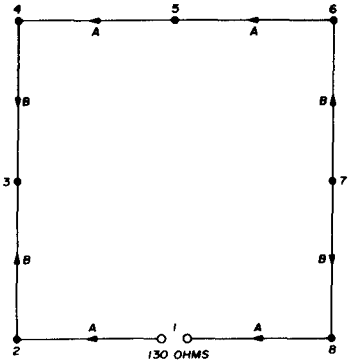


fig. 12. Current distribution of a full-wavelength loop, A, a shorted half-wavelength transmission line, B, and a folded dipole, C. The folded dipole provides the same gain as a reference dipole.

folded dipole to have 0 dB gain and a feedpoint resistance of 292 ohms (4 x 73 ohms).

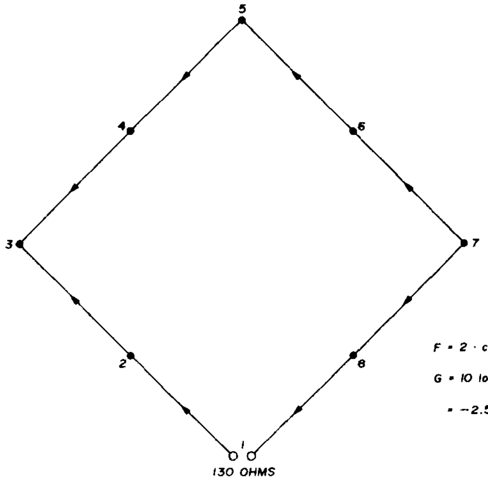
In the following discussion, the feedpoint resistance of symmetrical loops will be assumed to be 130 ohms. This is consistent with reference 10 which gives a value of 125 ohms for the square loop, and reference 11 which reports that the calculated value for a circular loop is 140 ohms. By symmetrical I mean a square, a diamond, an equilateral triangle, or



$$F = 4 \cdot \frac{k\lambda I}{2\pi} \sin \frac{2\pi}{\lambda} \cdot \frac{\lambda}{8} = 2\frac{k\lambda I}{\pi} \sin \frac{\pi}{4} = \sqrt{2} \cdot \frac{k\lambda I}{\pi} = \sqrt{2} \cdot F_R$$

$$G = 10 \log \frac{73}{130} + 20 \log \sqrt{2} = -2.51 \text{ dB} + 3.01 \text{ dB} = +0.5 \text{ dB}$$

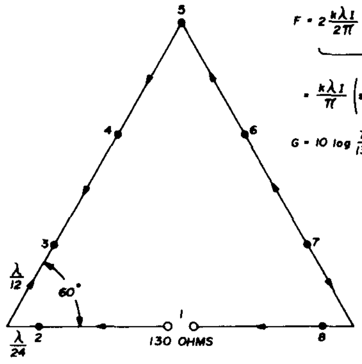
A SQUARE OR QUAD LOOP



$$F = 2 \cdot \cos 45° \cdot F_R = \sqrt{2}\, F_R$$

$$G = 10 \log \frac{73}{130} + 20 \log \sqrt{2} = -2.51 \text{ dB} + 3.01 \text{ dB} = 0.5 \text{ dB}$$

B DIAMOND LOOP

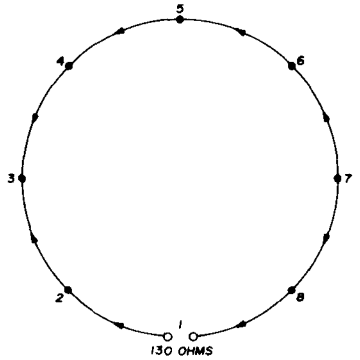


$$F = \underbrace{2\frac{k\lambda I}{2\pi} \sin \frac{2\pi}{\lambda}\left(\frac{\lambda}{8}+\frac{\lambda}{24}\right)}_{\text{BASE}} + \underbrace{2 \cdot \cos 60° \cdot \frac{k\lambda I}{2\pi} \sin \frac{2\pi}{\lambda}\left(\frac{\lambda}{4}-\frac{\lambda}{12}\right)}_{\text{SIDES}}$$

$$= \frac{k\lambda I}{\pi}\left(\sin \frac{\pi}{3} + \frac{1}{2} \sin \frac{\pi}{3}\right) = F_R \cdot \frac{3\sqrt{3}}{4}$$

$$G = 10 \log \frac{73}{130} + 20 \log \frac{3\sqrt{3}}{4} = -2.51 \text{ dB} + 2.27 \text{ dB} = -0.24 \text{ dB}$$

C EQUILATERAL TRIANGLE LOOP



$$F = 4k\int_0^{\frac{\lambda}{4}} I_H \cos \frac{2\pi}{\lambda} X \, dx$$

WHERE I_H = HORIZONTAL COMPONENT OF CURRENT

$$= I \cos \frac{2\pi}{\lambda} X$$

$$F = 4kI\int_0^{\frac{\lambda}{4}} \left(\cos \frac{2\pi}{\lambda} X\right)^2 dx$$

$$= 4kI \cdot \frac{\lambda}{2\pi} \cdot \frac{\pi}{\lambda} \cdot \frac{\lambda}{4} = \frac{k\lambda I}{2} = \frac{\pi}{2} \cdot F_R$$

$$G = 10 \log \frac{73}{130} + 20 \log \frac{\pi}{2} = -2.51 \text{ dB} + 3.92 \text{ dB} = 1.41 \text{ dB}$$

D CIRCULAR LOOP

fig. 13. Gain calculations for symmetrical full-wavelength loops. Note that for the square or quad loop, A, the four A segments add in phase, and the four B segments cancel. The diamond loop has the same gain as the square loop. In the circular loop, D, the variable x is measured clockwise around the loop with $x = 0$ at the feedpoint.

a circle. The gains of symmetrical loops are calculated in **fig. 13**.* All of the loops are fed so that polarization is horizontal. Note that movement of the feedpoint to points **3** or **7** will provide vertical polarization. The gain is the same for either polarization; this is obvious for the square, diamond, and circle, and may be easily shown for the equilateral triangle (a calculation similar to that of **fig. 13C** if you assume that the feedpoint resistance is unchanged).

In reference 13 the top-loaded delta loop is introduced as an efficient vertical radiator for use where height is limited. In **fig. 14** gain calculations are shown for a top-loaded delta loop and an isosceles triangle loop (called here the low delta loop) with the same vertical dimension. Again, a feedpoint resistance of 130 ohms is used, but this time the figure comes from measurements made at W1DTV for 80-meter antennas whose bases are 3 meters (10 feet) above ground. The free-space values of feedpoint resistance are unknown.

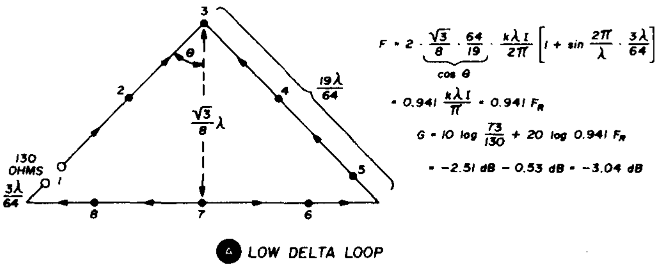
It is interesting to compare the results of the full-wavelength loop gain calculations.

shape	gain
Circle	1.4 dB
Square	0.5 dB
Diamond	0.5 dB
Equilateral triangle	−0.2 dB
Top loaded delta loop	−0.7 dB
Low delta loop ($\frac{\sqrt{3}}{8}$ λ high)	−3.0 dB

All of these results assume the same feedpoint resistance. This is a shaky assumption, and should be examined analytically as was done for the circular loop in reference 11. In any event, the methods given here allow you to revise these results if and when more solid feedpoint resistance data are available.

If the assumed feedpoint resistances are correct, however, there is a significant gain penalty when a low delta loop is used as a vertical radiator. Fortunately, top loading may be used to recover some of this loss. The low gain in a direction perpendicular to

*These gain figures and those in the following table differ from the gain figures given in reference 12 which states that the free-space gain of full-wavelength loops in the shape of squares, diamonds, or circles is approximately 1 dB. The discrepancy may be a result of using incorrect free-space feedpoint resistance values. **Editor.**



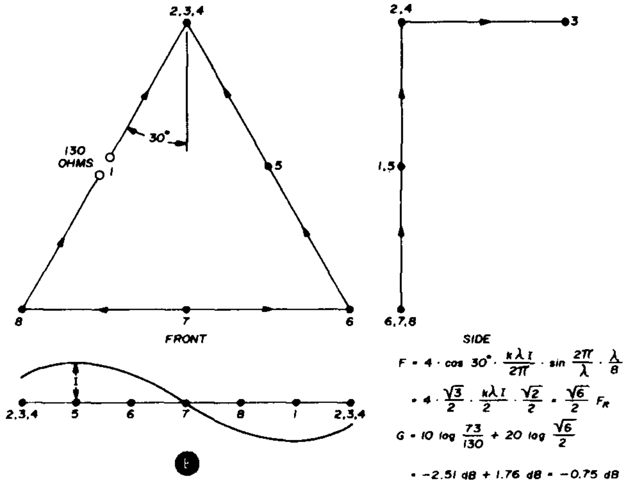


fig. 14. Gain calculation for the low delta loop, A, and the top loaded delta loop, B. The current distribution in the low delta loop is the same as that shown in fig. 12A; note that radiation from the base is cancelled.

the plane of the low delta loop is probably indicative of a pattern somewhat different than that of the more symmetrical configurations. The current distribution suggests more high-angle radiation than the other loops, also.

summary

The results of the gain calculations have been summarized in **table 3**. Also shown is the assumed feedpoint resistance and the size of the antenna in wavelengths. One observation I have made as I became interested in this subject is that there are many conflicting published gain values for various antenna configurations. In particular, the values for full-wavelength loops and the monopole vary considerably. In most cases, however, the method for arriving at a particular published value of gain has not been disclosed. In this paper, a straightforward approach for calculating gain has been presented and applied to a number of examples. The intention is not to present the final word regarding antenna gain figures, but to provide insight and a basis for future work. This has been done by splitting the gain calculation into several parts: the effect of feedpoint resistance, the effect of current distribution, and the effect of the number of driven radiators. Although the approach is not applicable for all antennas, it does apply to enough popularly used antennas to be worthwhile.

It has been pointed out in the past how valuable the knowledge of feedpoint impedance can be for tuning and feeding an antenna; the emphasis here has been to point out another reason for measuring this quantity — the calculation of gain.

table 3. Summary of antenna gains as compared to a half-wavelength dipole in free space.

type	assumed feedpoint (ohms)	size	gain (dB)
Broadside (0.75λ)	51*	0.75λ x 0.5λ	4.6
Broadside (0.5λ)	59*	0.5λ x 0.5λ	3.9
Collinear (0.5λ)	70*	1.5λ x 0λ	3.2
Collinear (0.25λ)	76*	1.25λ x 0λ	2.8
Collinear (0.125λ)	87*	1.125λ x 0λ	2.2
Collinear (0λ)	94*	1.0λ x 0λ	1.9
Circular loop	130	0.32λ x 0.32λ	1.4
Broadside (0.25λ)	116*	0.5λ x 0.25λ	1.0
Square	130	0.25λ x 0.25λ	0.5
Diamond	130	0.35λ x 0.35λ	0.5
Broadside (0.125λ)	136*	0.5λ x 0.125λ	0.3
Half-wavelength dipole	73	0.5λ x 0λ	0
Folded half-wave dipole	292	0.5λ x 0λ	0
Quarter-wave dipole (inductively loaded)	26	0.25λ x 0λ	−0.1
Equilateral triangle loop	130	0.33λ x 0.29λ	−0.2
Inverted V ($\theta = 90°$)	40	0.35λ x .18λ	−0.4
L antenna ($\theta = 90°$)	40	0.25λ x 0.25λ	−0.4
Top-loaded delta loop	130	0.25λ x 0.22λ	−0.7
Low delta loop	130	0.41λ x 0.22λ	−3.0

*Each dipole; spacing shown in parentheses.

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ham radio

the ham notebook

Heath HD-1982 Micoder for low-impedance operation

The new Heathkit HD-1982 microphone with installed *Touch-Tone* pad provides a very convenient way to connect a pad into a transceiver which has no auxiliary audio input jack. There is a problem, however, in using the unit with transceivers such as the Drake TR-22C which is designed for low-impedance microphones. The HD-1982 is designed to operate into a load of 10k-ohms or higher.

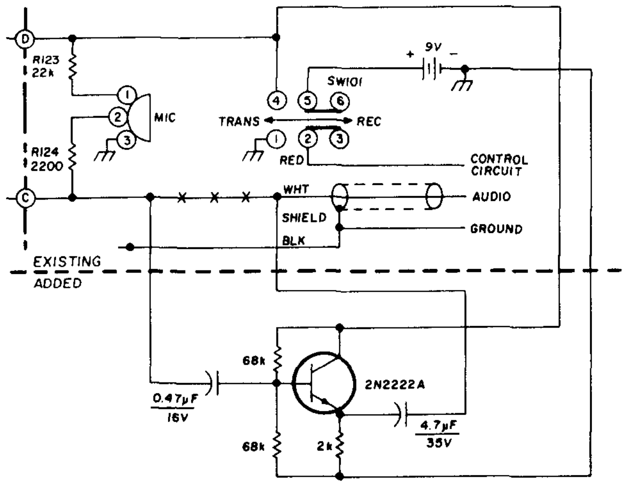

fig. 1. Schematic diagram of the emitter follower used to interface between the HD-1982 Micoder and a low-impedance microphone input.

Thinking that the mismatch might not be too serious, I tried the HD-1982, as designed, with the TR-22C. On-the-air checks were made with four stations. All reported muffled voice quality which cleared up upon changing to the Drake microphone. Some corrective measures were clearly indicated.

An emitter follower circuit was designed and connected into the circuit as shown in **fig. 1**. The input impedance of the circuit consists of the parallel combination of the two 68k resistors and the transistor H_{fe} X (2000), or about 30-k ohms. The output impedance is approximately 2000 ohms. The emitter follower was mounted on a small piece of phenolic board approximately 1-inch (2cm) square. The circuit board fits easily in the top of the microphone case between the two mounting posts. The length and routing of the four leads connected into the existing circuitry is not critical. Because the circuit had to be mounted on a board of limited dimensions, the size of the coupling capacitors is critical; I used Caelectro part numbers A1-302 and A1-306. These capacitors are about the size of the head of a match and fit nicely in the available space.

The modification corrected the problem. On-the-air checks have indicated no essential difference between the performance of the Drake microphone and the modified HD-1982. *Touch-Tone* operation has been very successful with reports of good quality audio.

Wesley Johnson

cleaning teleprinters

My plans to rewire a surplus model 19 Teletype machine and interconnect it to the model 14 TD were hindered by the dirt and oil build-up on the machines. Having an unusually neat and clean shack, it was imperative they be cleaned!

After a few fruitless efforts, the idea to use a commercial degreaser was conceived. A quart can of *Gunk* all-purpose degreaser was sprayed all over the internals (after removing only the motor) using a spray bottle such as a *Fantastik* cleaner applicator. After allowing approximately 20 minutes for the degreaser to work, the machine was rinsed thoroughly using the fine spray of the garden hose hooked to the hot water spigot. Extremely greasy areas were then given a second application of *Gunk*, a light scrubbing with a paint-brush, and more vigorous application of hot-water. Drying can be expedited by setting the unit in the sunshine and/or using your wife's electric hair styler or vacuum cleaner with the hose attached to the exhaust.

After complete drying, the unit should be lubricated to prevent rust. The model 14 TD was also cleaned in this manner. The motors, however, were cleaned with heavy shop-rags and solvent. I have since cleaned several motors using this technique (with *Gunk* and hot water) so I suspect the TTY motors could be cleaned while on their mountings.

As can be imagined, this is a very messy operation. The best place to do it is outside on the gravel driveway. The results, however, are nothing short of fantastic! The aluminum frame sparkles like new; the levers and gears look like the day they were assembled.

It may only be my imagination, but the model 19 seems to run quieter and better since cleaning. It's definitely easier to work on.

F. Neil Urban, W8CD

remote rf current readout

The amount of current flowing in an antenna or a feedline is often more indicative of efficient system operation than vswr. Moreover, having such a current readout conveniently located at the operating position greatly simplifies transmitter adjustment and rapid frequency changes.

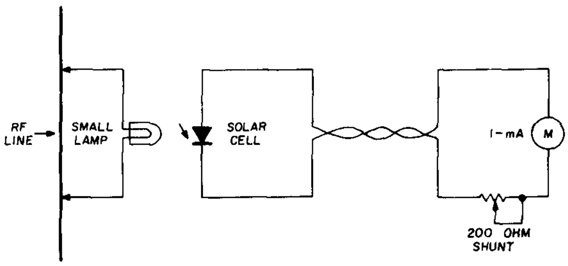

fig. 2. Simple circuit for remote readout of rf current on your transmission line.

The rf current readout described is based on the use of a simple optically-coupled isolator, as illustrated in **fig. 2**. A suitable pilot lamp is illuminated by a small sample of rf and energizes an inexpensive solar cell; the dc current generated by the cell is a measure of relative rf power, and may be routed to a low-current meter located at any convenient point.

A sensitive, low-current pilot lamp is desirable to cause minimum disturbance to normal rf circuit conditions. The Number 48 or 49, 60 mA lamp is suitable for use with transmitters above 1-watt output. The solar cell may be an International Rectifier B2M or any similar device. A meter, reading 1 milliampere or less is suitable. A variable current-limiting control, although not absolutely necessary, will add convenience to the system.

The solar cell and lamp may be taped together, using dark tape to prevent light from other sources reaching the cell. The lamp filament should parallel the solar cell surface for maximum sensitivity.

The lamp may be connected directly in series with a conductor which carries very low rf power. For higher current, the lamp is connected across a suitable portion of the line, which then serves as a shunt to limit lamp current to a safe value. Initially, the lamp shunt should have low resistance, and be gradually increased until only useful brilliance is obtained at maximum rf power output.

The rf readout system provides complete isolation between rf and dc circuits, allowing efficient placement of the rf pickup and convenient location of the relative rf-reading meter. During initial adjustment of a new antenna system, several inexpensive pickups may be temporarily installed to monitor rf current in various components.

Gene Brizendine, W4ATE

multiplexed counter displays

We have received many questions from readers concerning the digital display for the counter shown in the article on page 22 of February, 1978, issue of *ham radio*. The author made two references to the type of display used with the counter. In one he stated that, "the multiplexer in the 7208 energizes each LED in sequence." This means that the lines coming from the 7208 (pin 5 for example) are used to turn on (enable) the appropriate digit.

Segment information is also obtained from the 7208 IC. In this case, the same information goes to all the LEDs. Pin 28 of the 7208, would go to the *a* segment of each LED, and pin 17 to the *b* segments, etc. During operation, the segment information appears at the same time as the digit enable line. For additional information, refer to the article by John Bordelon, K4JIU, on page 30 of the same issue. He uses the same technique in another version of the same counter.

At the present time, the ICs can be obtained from at least three sources: Circuit Specialists, James Electronics, and Poly-Paks.

Charles Carroll, K1XX

emergency quad antenna repairs

The worst enemy of the cubical quad antenna is the kind of winter storm that simultaneously subjects it to ice and wind loading. Here in the Chicago area we don't get that kind of storm too often, but when we do most quad owners seem to end up with some antenna damage. A quad antenna with a broken spreader, besides being useless as an antenna, is also extremely vulnerable to further damage if not promptly repaired.

A few winters back, I arose one morning to find three broken spreaders on my two-element quad. Not having any replacement spreaders on hand, I nevertheless managed to have the antenna fully operational and structurally sound again within a few hours. I went to the hardware store and purchased a length of 19 x 19 mm x 3 mm (3/4 x 3/4 inch x 1/8 inch) thick aluminum angle stock. I cut it into 30 cm (12 inch) lengths and used two pieces as splints to repair each break, one on each side of the spreader. The splints were secured by wrapping them tightly at each end with no. 14 (1.6mm) solid wire, and a layer of tape was added over the wire for good measure. The antenna was used in this condition for several months with no apparent effect on its performance due to the aluminum splints.

John E. Becker, K9MM

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NEW products

For literature on any of the new products, use our *Check-Off* service on page 150.

new CDE rotor for super antennas

Cornell-Dubilier Electric Company has introduced a new heavy-duty rotor, the *Tail Twister*, to handle antennas with up to 2.6 square meters (28 ft^2) of wind load area. A new control box was designed for the rotor to complete the system.

The rotor incorporates the highly successful HAM II design with a new, thicker, cast-aluminum bell housing. Wider reinforced webs of the housing permit easy support of large antennas. On this model the upper mast-support is predrilled to have a bolt-through installation for positive locking. Also new is a three-ring ball-bearing assembly to provide increased side thrust control and vertical load-carrying capacities. The motor is a new design with an automatic coast-down prebrake action and a metal pinion gear to guard against stripping.

The control box features a full metered indication of the antenna direction with front panel control for calibration and brake. A separate on/off switch is provided for instant antenna location and brake operation. LEDs provide a positive signal for rotational power and brake operation. The unit is attractively housed in a black satin case. Low-voltage control assures safe operation for the user and installer.

The *Tail Twister* system is designed for tower mounting as required for most "super" communications antennas. Weighing slightly over 8 kg (18 pounds), with a height of 36 cm (14-1/16 in), and a diameter of 2.3 cm (9-5/16 in), the unit is secured with six bolts provided. The mast diameter is a hefty 5 cm (2 in). For further information, please contact Mr. W. Carlson, Cornell-Dubilier Electric Co., 150 Avenue L, Newark, New Jersey 07101, or call (201) 589-7500.

SST ultra tuner

SST Electronics has introduced the SST T-2 ultra tuner to tune out the swr on any coax-fed or random-wire antenna. The T-2 will work on 160 through 10 meters and is capable of

handling 200 watts. It uses a toroid inductor and specially made capacitors for small size. Its compact, rugged, attractively finished, bronze enclosure is 134 x 57.5 x 63.5 cm (5-1/4 x 2-1/4 x 2-1/2 inches). SO-239 type connectors are used for the transmitter input and coax-fed antennas, while convenient binding posts are used for the random wire and ground connections.

The SST T-3 impedance transformer matches 52-ohm coax to the low impedance of a mobile whip or vertical. Using a 12-position switch, with taps between 3 and 52 ohms, this impedance transformer provides broadband matching between 1 and 30 MHz. The SST T-3 is also housed in an attractive, bronze-finished cabinet [70 x 51 x 57.5 cm (2-3/4 x 2 x 2-1/4 inches)], with a toroid inductor accounting for its small size.

All SST products are guaranteed for one year. In addition, they may be returned for a full refund within 10 days. The SST T-2 costs $49.95, while the T-3 is priced at $19.95. For additional information, contact SST Electronics, P.O. Box 1, Lawndale, California 90260.

mobile communications antennas

A full line, general catalog featuring communications antennas for all frequencies normally used for mobile-to-mobile and mobile-to-base operations, has just been published by Larsen Electronics. Both quarter-wave and gain type antennas are featured, with a variety of both permanent and temporary mounts included. In all, over 200 antenna types, frequency ranges, and mounting styles are detailed. The catalog is fully indexed by both number designations and description for easy finding of any specific style or model.

Two special features of the catalog are an explanation of the easy-memory designation system used by Larsen and the complete series of copyrighted "Select-a-Charts." As explained in the catalog, the Larsen

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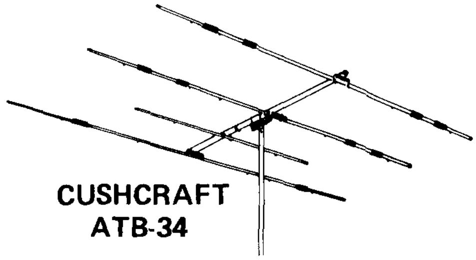

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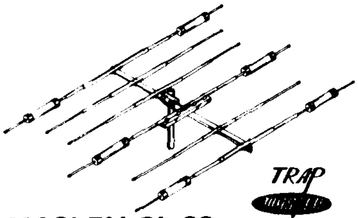

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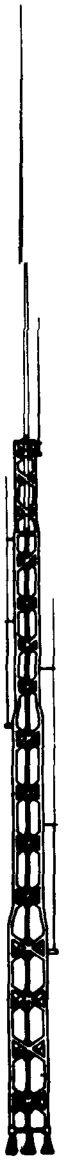

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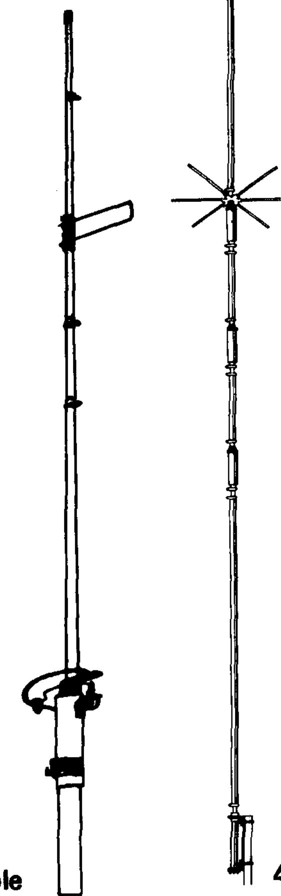

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numerical system uses a combination of letter and number designations. The letters tie in with the origin or common term for the antenna (such as NMO for new Motorola-mount, MM for magnetic mount, etc.). The numerals always relate to the frequency.

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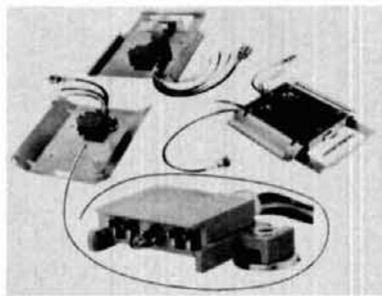
Copies of the Larsen catalog are available by writing to Larsen Electronics, P.O. Box 1686, Vancouver, Washington 98663.

quick-disconnect mobile mount bracket



Scientific Dimensions, manufacturer of the STOP-A-THIEF "Quick Disconnect" radio slide mount system, has just introduced a new unit to the line: the SDI-700. This slide mount provides easy removal of two-way radios for antitheft and/or convenience in transferring and interchanging radios from one vehicle to another. The new SDI-700, with its AMP *Coaxicon* contact will handle up to 500 MHz so it can be used for amateur and small business radios at the economical price of $19.95 each. It is the same high quality mechanical design as the popular model SDI-1000 with its gold-plated AMP *Coaxicon* contact, which will handle up to 1000 MHz.

All models in the "Quick Discon-

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nect" mobile radio slide mount line feature the modified AMP *Dualatch* connector system which houses the AMP *Miniature Coaxicon* coaxial cable contact. This assures maximum performance and absolute contact for thousands of connections. The radio is easily removed by the operator when leaving his vehicle, to prevent theft of the radio and damage to the vehicle.

The SDI-700 unit, like the SDI-1000, is precision made of 16-gauge steel with durable chrome finish. It has a three-way spring lock to guarantee positive connector contact every time and to eliminate rattles while the vehicle is in motion, yet the lock is designed to enable easy and quick removal of the radio from its mount in the vehicle with just a slight finger pressure on the release lever. All wire leads in the slide mount unit are securely clamped into place to prevent breakage. The no. 18 AWG stranded power and accessory leads have 10 ampere capacity, and the coaxial cable is RG-58C/U with UHF connectors attached.

To provide for easy transfer of the radio to another vehicle, Scientific Dimensions offers extra individual stationary mounts for each unit. The company has also designed and produced the universal TILT-N-TURN, model SDI-500 mounting bracket for all its slide mount units to provide easy visibility and operation of the radio. The SDI-500 unit is easily mounted on a vehicle's hump, floor, roof, or dash and similar locations on boats and farm equipment. It provides 50 degree tilting and turning flexibility, making head-on viewing possible from the driver's seat, thus reducing distraction of attention from operating the vehicle.

For more information on the complete line of Scientific Dimensions' products, use check-off on page 150, or write to them at 309 McKnight NE, Albuquerque, New Mexico 87102.

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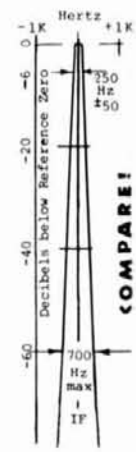

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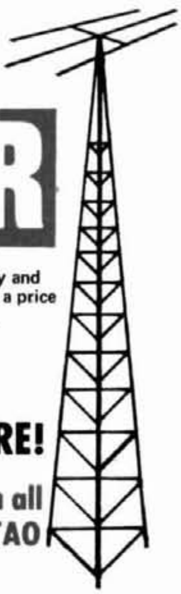

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MOSLEY TA-33-JR	3-Element/1KW	$151.85	$136.50	$247.50
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MOSLEY TA-36	6-Element/2KW	$335.25	$299.95	$399.95
MOSLEY CL-33	3-Element/2KW	$232.50	$209.00	$317.25
MOSLEY CL-36	6-Element/2KW	$310.65	$279.50	$384.50
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70 KHz to 160 MHz

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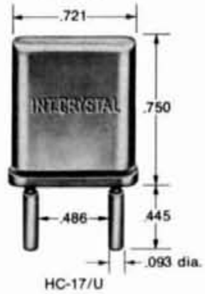
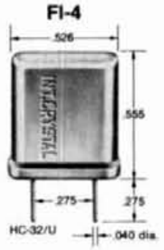
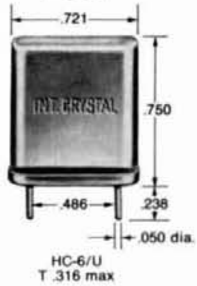
F-609



FI-4

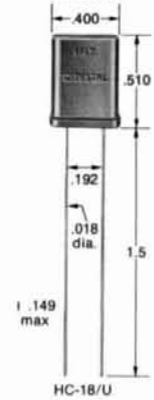


F-605

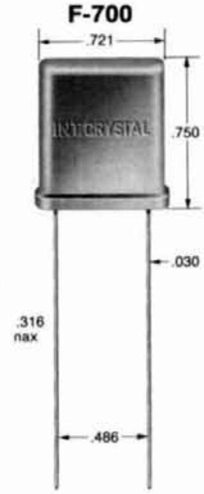


F605-SL

FM-1

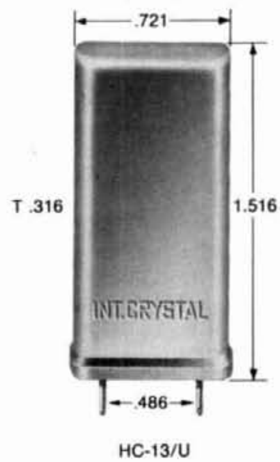


F-700

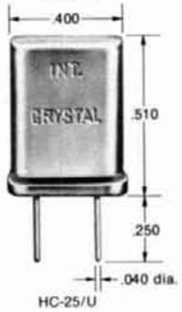


International Crystal Manufacturing Co., Inc. guarantees every crystal against defective materials and workmanship for an unlimited time, when used in equipment for which they were specifically made.

F-13



FM-2



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12 Channels
15 Watts
$229.00 Amateur Net

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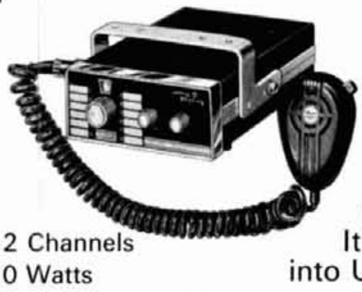

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CES Model 230 has Snap Action Tactile Keys and is only 3-3/4 x 2 inches. A 500 ohm Dynamic Mic with built in Tone Pad has adjustable balance and level controls and can be used with any transceiver using a 500 ohm mic.
Model 225 with non-tactile keys $44.95

Battery Powered Tone Encoder with Snap Action Keys

The CES Model 300 is powered by a 9v self-contained battery and includes an integral, rear mounted speaker with external volume control. This acoustically coupled tone pad is small enough to be carried in shirt pocket.

$49.95

Autodialer

CES Model 251 stores up to 10 twenty-two digit phone numbers including access and clear codes. Autodialer is programmed from the keyboard. Internal battery holds memory when disconnected from 12v supply. Numbers not in memory can be dialed from keyboard. Available in tone or pulse dialing.

$99.95

Thin Tone Encoders

2-7/8 x 2-1/4 x 7/16 inches with Snap Action Keys

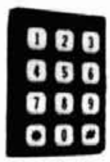
CES Model 211 has a Snap Action Keyboard with standard dual tone frequencies. Other features include a .2% accuracy and is crystal controlled. MOS Digital IC provides high immunity to R.F. and voltage fields regulated for supplies of 7 to 35v. Easily mounted to Transceivers.

$39.95

Model 210 same as 211 but with non-tactile keys.

Miniature Tone Encoders

2-3/8 x 1-11/16 x 3/8 inches Smallest



Model 215

CES Model 215 is the smallest the industry offers — with all the features of the above Model 210. Order Model 216 with Snap Action Tactile Keys.

$39.95

Model 216 with Snap Action Keys $44.95

Pad Mounted on Standard 146 A or Motorola HT

CES Model 220 offers complete back assembly with Touch Tone Encoder already mounted and ready to plug into the private channel connector. With an LED Tone indicator.

Model 220 for Standard 146A
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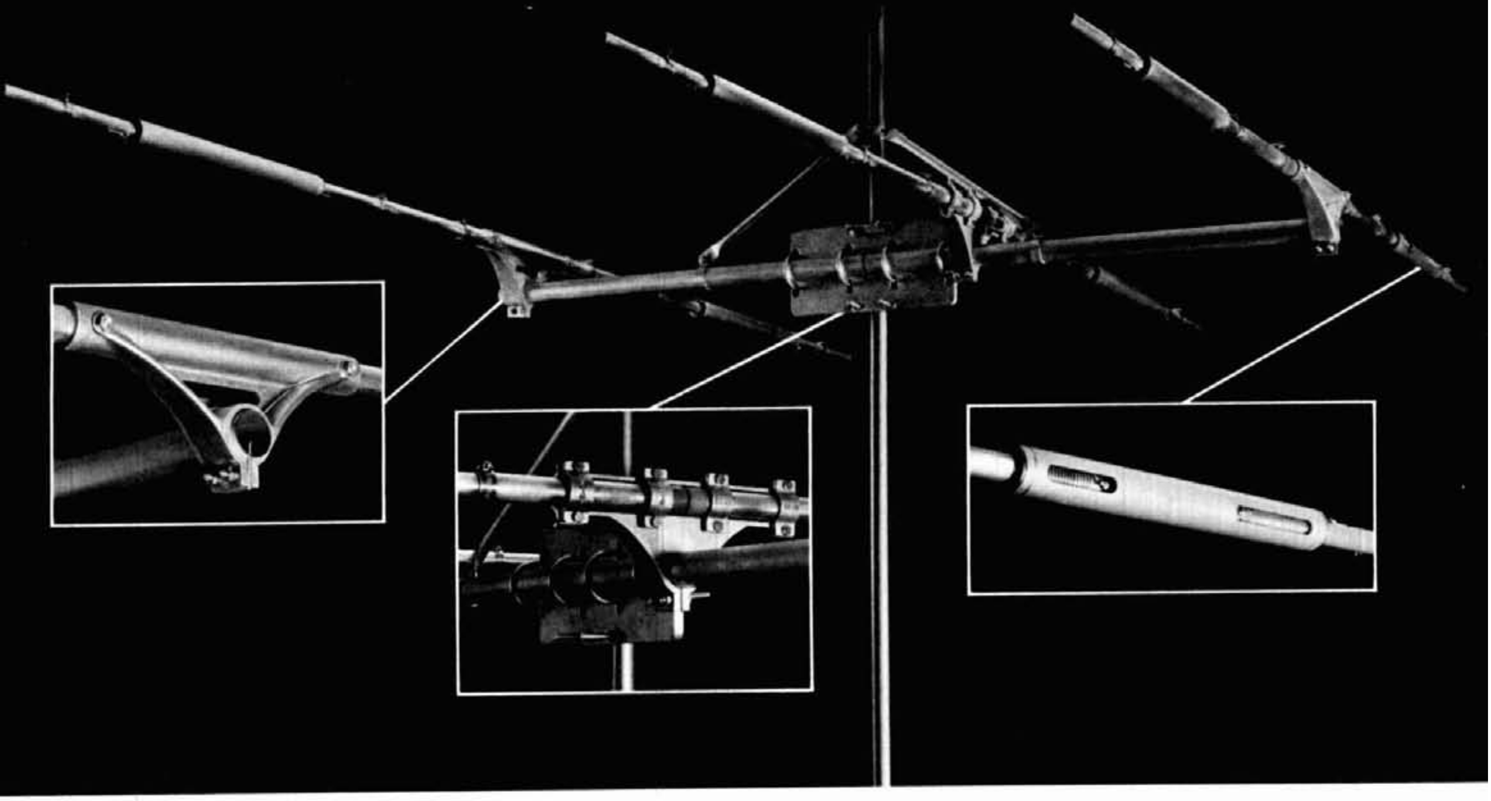
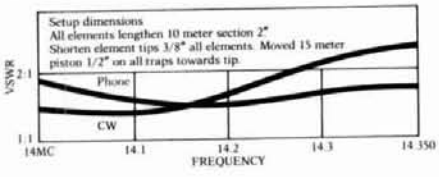
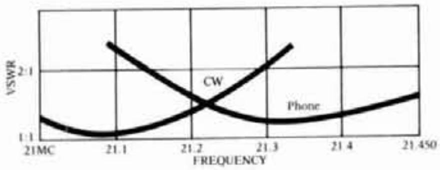
TOO STRONG TO GIVE THE WIND A BREAK!

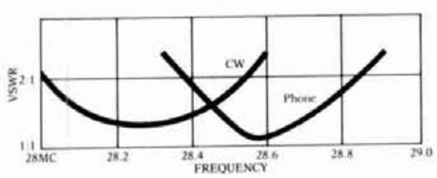


Braced against the elements as only Swan can do it! Even hurricane winds to 100 MPH can't lower the boom on your operations.

The Swan TB3HA Tri-Bander: a really heavy-duty 20-15-10 meter beam. 3 solid elements, all work-







ing on all bands. With a VSWR of 1.5:1 or better at resonance, plus a full 2000-watt PEP rating, our TB3HA is built to work up a storm!

Strongest fittings available: this you've got to see. Exclusive cast-aluminum braces grasp tubes at every joint, spreading stress over an 11¾" span. Compared to slipshod U-bolt plates—no contest!

Reinforced by super specs, TB3HA's one tough competitor:

- ☐ 8dBd average forward gain.
- ☐ 20-22 dB front to back ratio.
- ☐ 16' turning radius.
- ☐ longest element: 28'2".
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- ☐ direct 52 Ohm coaxial feed.
- ☐ wind load @ 80 mph: 110 lbs.
- ☐ 44 lbs. net weight.
- ☐ $199.95 complete, Swan Credit Card accepted.

For $199.95 at your local Swan dealer you can start operating tri-band from a position of real strength—because TB3HA is Swan-engineered to work under pressure!

Please rush full information on Swan's Heavy Duty Tri-Band Beam Antennas:

☐ 3-element TB3HA
☐ 4-element TB4HA

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2¼" x 4" with stand, no charge.
☐ Please send my plaque enscribed with my station call __________

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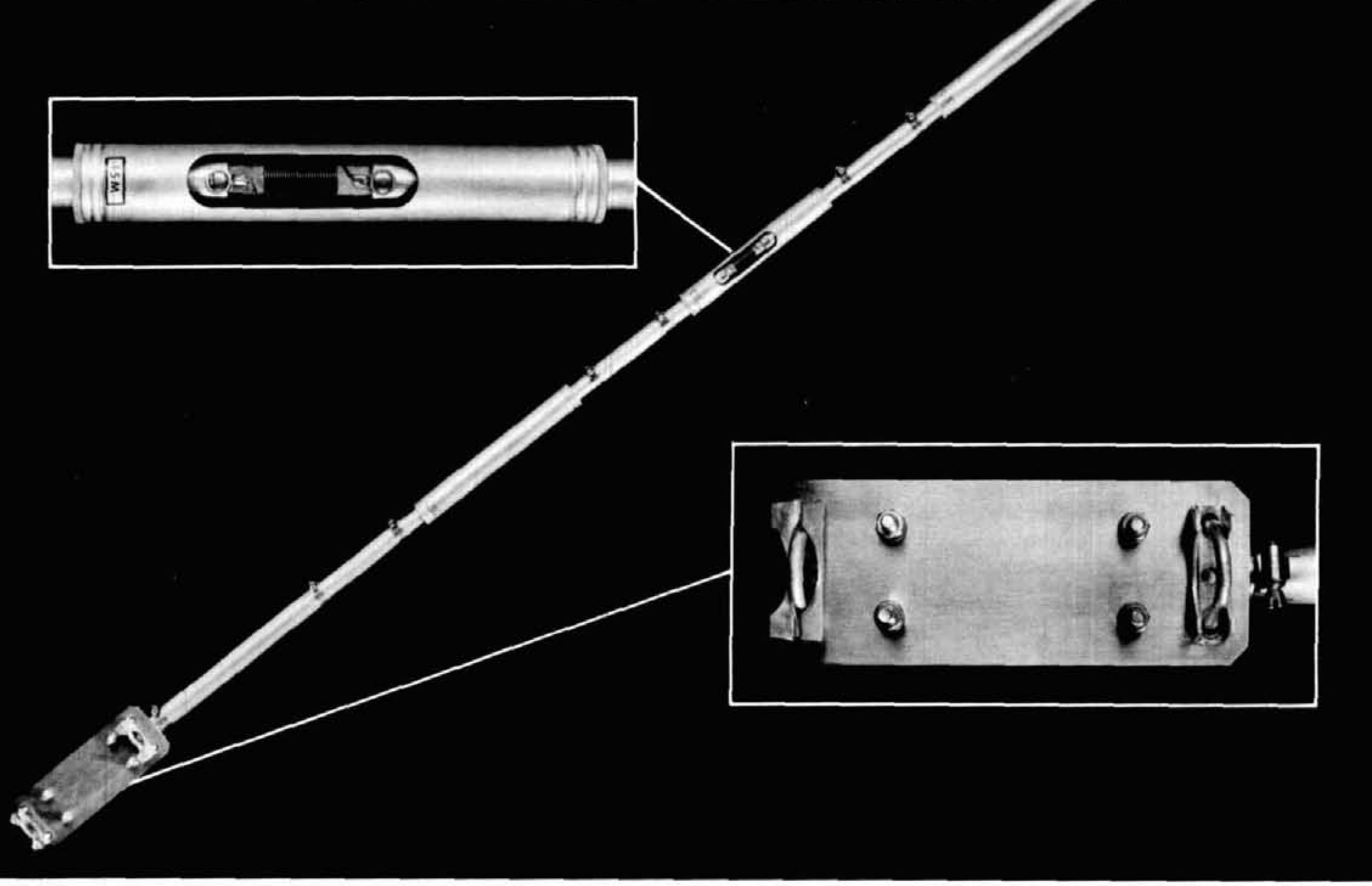
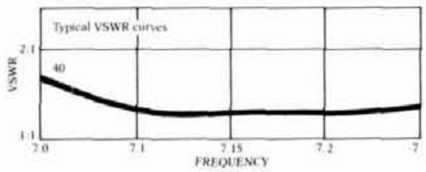
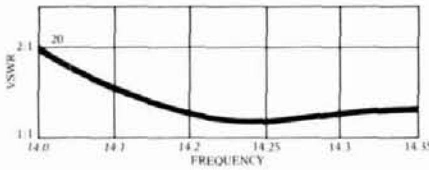
BUILT FOR ACTION FROM THE GROUND UP

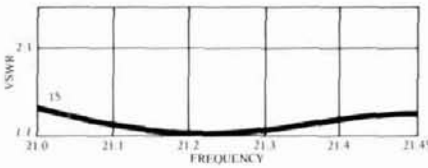


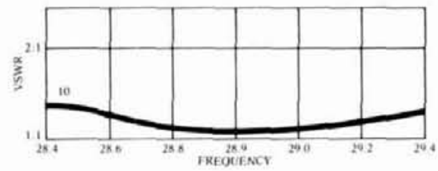
If you've got a calling for far-reaching action, have we got a number for you!

Swan's 4010V: precision engineered for 40-20-15-10 meters.









With slim-line traps, this 4-band vertical offers advanced light-weight construction *with* heavy weight performance.

4010V's fine-tuned to handle 2000 PEP. With a typical VSWR of 1.5:1 at resonance.

Powerfully designed—yet powerfully simple to set up.

No-hassle installation. The 21′ *vertical comes in short, easy to* assemble lengths. Complete with *mounting* hardware. You're up and running in record time.

Expandable too. No trick at all to stretch your reach into 75 meters, *with the optional* Swan 75 AK Kit.

Just $74.95 for an outstanding 4-band trap vertical, the 4010V (and $39.95 for the 75-meter add-*on kit*), *at your* Swan *dealer.* And by all means, use your Swan Credit Card.

Please rush full specs for Swan's
☐ 4010V 4-band trap vertical antenna
☐ 75 AK 75-meter Kit

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Address__________

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TTP-1 (12-digit) **$59.00**
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MODEL DTM — Completely self-contained miniature encoder for hand-held portables. Only 5/16″ thick. Three wire connection. Automatic PTT keying optional. With your choice of keyboards. **Price DTM - $39.00, DTM-PTT - $49.00.**

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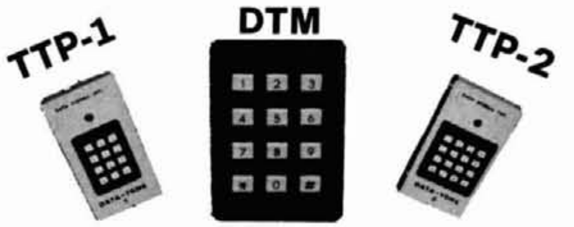
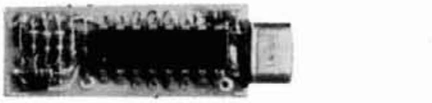
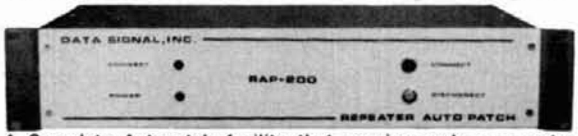


SUB-MINIATURE ENCODERS



MODEL SME — Smallest available Touch Tone Encoder. Thin, only .05″ thick, keyboard mounts directly to front of hand-held portable, while sub-miniature tone module fits inside. This keyboard allows use of battery chargers. **Price $29.00,** with your choice of keyboards. **SME** (less keyboard) **$24.00**

AUTOPATCH — Ready to go!



A Complete Autopatch facility that requires only a repeater and a telephone line. Features include single-digit access/disconnect, direct dialing from mobile or hand-held radios, adjustable amplifiers for transmitter and telephone audio, and tone-burst transponder for acknowledgement of patch disconnect.

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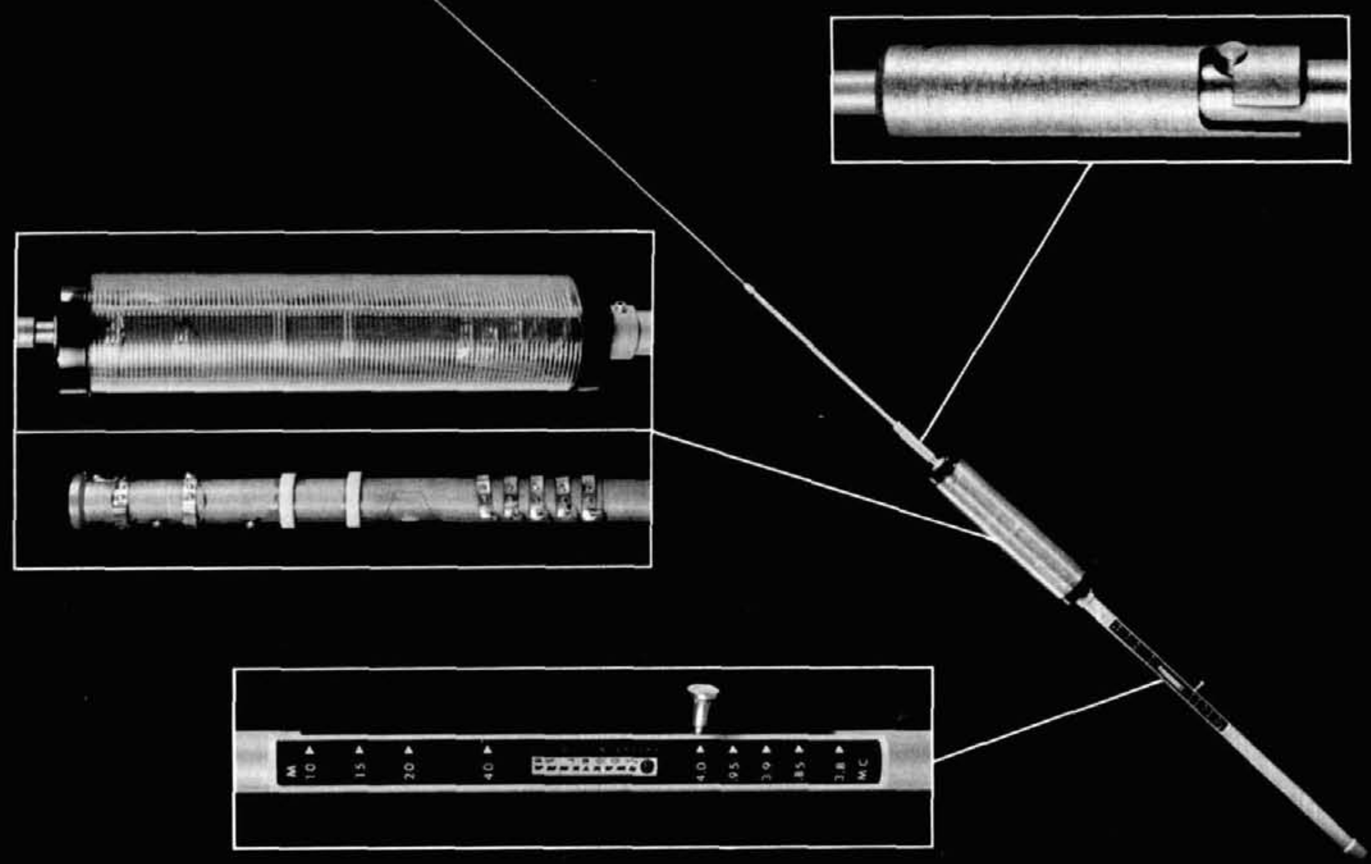
GETTING OUR STRONG POINTS OUT IN THE OPEN



Before you take off for the wide open spaces, close in on the top performer in the field:

Swan's 5-band Mobile 45, loaded with high-engineering specs unavailable from any other antenna source today.

Wait till you see how we've got band-switching in hand. This has got to turn you on!

No more coil changing. At the flick of the positive-stop switch, skip across all 5 bands freely using our *High-Q tapped coil.* Nine positions to shift to:

☐ one each for 10, 15, 20 and 40 meters

☐ five settings on 75 meters

And gold smooths the way. *Only Mobile 45's rods slide* through switch contacts that we've *gold-plated,* to deliver a swan-song *to corrosion.*

All for $119.95, with 1000 watts PEP, and the same extra rugged, *heavy duty construction you* always count on from Swan.

Check out Mobile 45's exclu-*sive strong points at your Swan* dealer, where—your Swan Credit Card's good anytime.

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City __________

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FREE! Personalized call-letter plaque

$2\frac{1}{2}$" x 4" with stand, no charge

☐ Please send my plaque imprinted with my station call __________

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50 ohm COAX RG 174 U 50 FT. $2.75

.1 ohm FIVE WATT RESISTOR 5/$1.00

SO-239 65¢ PL-259 65¢ 10/$6.00 MIX/MATCH

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THIS FINE QUALITY KEYBOARD, WITH ORIGINAL COST OF $96.00, IS NOW OFFERED TO YOU FOR ONLY.... $19.95

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MJ1000	NPN	90	8	60	$1.25
MJ900	PNP	90	8	60	$1.65

BOXER FANS 100CFM 4.5 INCH FAN OPENING 1.5 INCHS DEEP EXHAUST OR BLOW 115 VOLT A.C. USED BUT GUARANTEED $6.95
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SOLDER TAIL IC SOCKETS 8 PIN $.20 14 PIN $.18 16 PIN $.18 24 PIN $.35 40 PIN $.85

MDA 970-2 BRIDGE 100 PIV 4 AMP. $.75

THIS ASSORTMENT OF POTS ARE NEW. WE JUST HAVE NOT HAD TIME TO SORT THEM. ALL ARE PC TYPE. 10/$1.00

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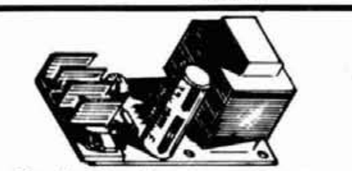

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* 5-12V ADJUSTABLE POWER SUPPLY AT 12V 500MA ADJUSTABLE POT.

ALL PARTS INCLUDED

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PS2 12-28V $5.95

MARLIN P JONES & ASSOCIATES
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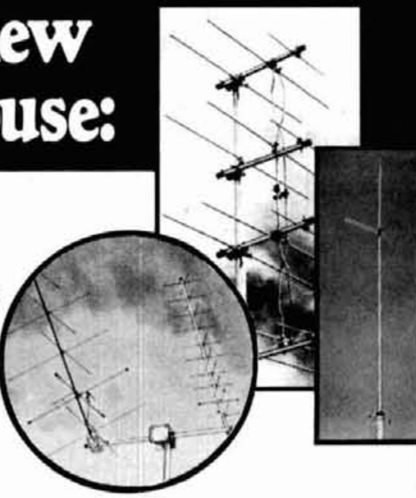

cushcraft

DX-ARRAYS 20 ELEMENTS

		List	OUR PRICE
DX-120	144 MHz	$42.95	$39.95

OSCAR ANTENNAS

A144-10T 145 MHz 10 El. Twist $34.95
A432-20T 430-436 MHz 20 El. Twist $49.95
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Total List Price $108.85
OUR PRICE $94.95

RINGO RANGER for FM

ARX 2 135-170 MHz
ARX 220 220-225 MHz
ARX 450 435-450 MHz
YOUR CHOICE **$30.00** Delivered in USA Reg. List $32.95

Make 2 Meters all the fun you had hoped for... put more signal into the other guy's receiver with a new Cushcraft 2 meter antenna.

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Move those 220 and 430 MHz signals in and out with the quality you deserve. A Cushcraft antenna from G.R. Whitehouse & Co.

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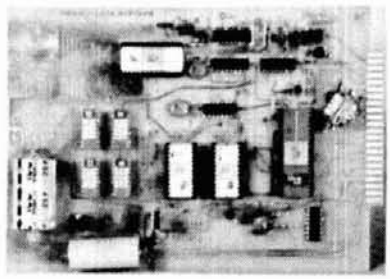
SEND FOR CATALOG with First Class Stamp

C. W. KEYBOARD MODULE



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BY POPULAR DEMAND —
we are continuing to offer with any purchase of $99 or more from ad or flyer,

FREE

*a Fairchild **clock module** FCS-8100A (suggested retail price $20).*

COMPLETE KITS: CONSISTING OF EVERY ESSENTIAL PART NEEDED TO MAKE YOUR COUNTER COMPLETE. **HAL-600A** 7-DIGIT COUNTER WITH FREQUENCY RANGE OF ZERO TO 600 MHz. FEATURES TWO INPUTS: ONE FOR LOW FREQUENCY AND ONE FOR HIGH FREQUENCY; AUTOMATIC ZERO SUPPRESSION. TIME BASE IS 1.0 SEC OR .1 SEC GATE WITH OPTIONAL 10 SEC GATE AVAILABLE. ACCURACY ± .001%, UTILIZES 10-MHz CRYSTAL 5 PPM.

COMPLETE KIT $149.00

HAL-300A 7-DIGIT COUNTER WITH FREQUENCY RANGE OF ZERO TO 300 MHz. FEATURES TWO INPUTS: ONE FOR LOW FREQUENCY AND ONE FOR HIGH FREQUENCY; AUTOMATIC ZERO SUPPRESSION. TIME BASE IS 1.0 SEC OR .1 SEC GATE WITH OPTIONAL 10 SEC GATE AVAILABLE. ACCURACY ±.001%, UTILIZES 10-MHz CRYSTAL 5 PPM.

COMPLETE KIT $124.00

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THE FOLLOWING MATERIAL DOES **NOT** COME WITH THE BASIC KIT: THE CABINET, TRANSFORMER, SWITCHES, COAX FITTINGS, FILTER LENS, FUSE HOLDER, T-03 SOCKET, POWER CORD AND MOUNTING HARDWARE.

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(Same Specifications as HAL-600A)
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HAL-0-300PRE $19.95
(Pre-drilled G10 board and all components)
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(Same as above but with preamp)
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PRE-BUILT COUNTERS AVAILABLE

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ALLOW 4- TO 6-WEEK DELIVERY ON PRE-BUILT UNITS.

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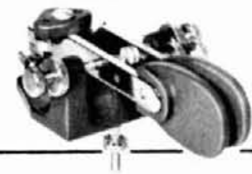
Model HK-1
- Dual-lever squeeze paddle
- Use with HK-5 or any electronic keyer
- Heavy base with non-slip rubber feet
- Paddles reversible for wide- or close-finger spacing

$29.95



Model HK-2
- Same as HK-1 less base for incorporation in own keyer

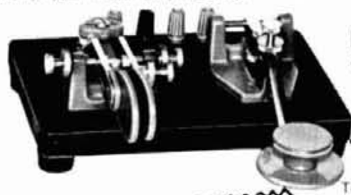
$19.95

Model HK-3
- Deluxe straight key
- Heavy base no need to attach to desk
- Velvet smooth action

$16.95

Model HK-3A
- Same as above less base $9.95

Navy type knob only $2.75



Model HK-4
- Combination of HK-1 and HK-3 on same base

$44.95

- Base only with rubber feet $12.00

Terminals red or black $.75 each

Model HK-5A Electronic Keyer
- New Cabinet Colored-Keyed to Match most modern radio equipment
- Iambic Circuit for squeeze keying
- Self-completing dots and dashes
- Dot memory
- Battery operated with provision for external power
- Built-in side-tone monitor
- Grid block or direct keying

- Speed, volume, tone and weight controls all mounted on front panel
- For use with external paddle, such as HK-1 or HK-4
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Same day shipment . . . PREPAID

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the super-compact alda 103

only 3¼" high x 9" wide x 12½" deep • less than 8¼ pounds

ALDA 103, the trim little powerhouse with incredible performance for the price! ALDA 103 provides a full 250 watts PEP input for SSB operation, and 250 watts DC input for CW. And when it comes to performance, ALDA 103 is the hottest little transceiver going — all solid state, totally broadbanded and super-stable VFO.

Ideal first transceiver for brand new novices! You'll want a full-capability CW/USB/LSB unit with all the power and performance you can use. ALDA 103 gives you 250 watts DC input for CW, the maximum allowable power for your novice license. When you upgrade to technician, you've got 2 bands for CW operation. And with your general license, just plug in your mic and use the ALDA 103's full 250 watts PEP on SSB!

Perfect second or mobile unit for seasoned hams! If you're looking for a super-sharp, compact unit to use in your car or boat, ALDA 103 will live up to your expectations. Absolute worst case sensitivity 0.5 uV for 10 dB S+N/N — a must for mobile operation. Receiver audio output of 3 watts minimum — another must. Also, *very low receiver power drain* of only 5.5 watts — that's 0.4 amps at nominal 13.8 VDC *including power for dial and meter lamps!*

$495

including microphone and mobile mount, too.

GENERAL SPECIFICATIONS

Semiconductors:	39 diodes, 23 transistors; 11 integrated circuits
Power Requirements:	Nominal 13.8 VDC input at 15 amps, negative ground only
Power Consumption:	Receive — 5.5 watts (includes dial and meter lamps); Transmit — 260 watts
Dimensions:	*3-1/4" high x 9" wide x 12-1/2" deep* (82.55 mm x 228.6 mm x 317.5 mm)
Weight:	8-1/4 lbs. (3.66 kg)

PERFORMANCE SPECIFICATIONS

Frequency Range:	80 meter band — 3.5 to 4.0 MHz 40 meter band — 7.0 to 7.5 MHz 20 meter band — 14.0 to 14.5 MHz
Modes:	CW; USB; LSB
RF Input Power:	SSB — 250 watts PEP nominal CW — 250 watts DC maximum (adjustable)
Transmitter:	
Antenna Impedance:	50 ohm, unbalanced
Carrier Suppression:	Better than −45 dB
Side-Band Suppression:	Better than −55 dB at 1000 Hz
Distortion Products:	Better than −26 dB
AF Response:	*500 to 2500 Hz*
Spurious Radiation:	Harmonics better than −45 dB below 30 MHz; better than −60 dB above *30 MHz*
Frequency Stability:	Less than 100 Hz drift per hour (from a cold start at room temperature)
Microphone:	High impedance 3000 ohm
Receiver:	
Sensitivity:	*Better than 0.5 watts audio output for* 0.5 μV input
Signal-to-Noise Ratio:	Better than 10 dB S+N/N for *0.5 μV input*
Image Ratio:	Better than −60 dB (typical with respect to 0.5 μV input: 80 meters — −130 dB; 40 meters — −100 dB; 20 meters — −75 dB).
IF Rejection:	Better than −70 dB (typical with respect to 0.5 μV input: 80 meters — 110 dB; 40 meters — 80 dB; 20 meters — 75 dB).
Intermodulation Intercept Point:	Better than 10 dBM
Selectivity:	2.5 kHz — 6 dB; 5.0 kHz — 60 dB
Audio Output Power:	More than 3 watts
Audio Distortion:	Less than 5% at 3 watts

OPTIONS & ACCESSORIES

Noise Blanker — Model No. PC 701 .... **$29.95**

100 kHz and 25 kHz *Dual Crystal Calibrator —* Model No. PC 801 .... **$14.95**

Portable Power Supply — Model No. ALDA PS 115: average duty 15 amp unregulated; input — 115/230 VAC, 50/60 Hz; output — 13.8 V nominal at 15 amps **$79.95**

Heavy Duty Power Supply — Model No. ALDA PS 130: output — *regulated 30 amp at 13.8 VDC; input —* 115/230 VAC, 50/60 Hz **$149.95**

alda communications, inc. 215 Via El Centro Oceanside, CA 92054 (714) 433-6123

ALDA 103 is completely manufactured in the U.S.A.

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flea market

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QUAD BUILDERS — Blizzard/Hurricane proof your antenna. Fiberglass vaulting poles. Incredible strength. Guaranteed new, 8-for $240.00. S.A.S.E. for info. K5WSE, Box 20-AA, San Antonio, Texas 78201. (512) 699-9260.

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QUAGI/YAGI BUILDERS — Insulated fiberglass booms to 16-feet. Straight, strong, vaulting poles. S.A.S.E. for info. Guaranteed new $30.00 K5WSE, Box 20-AA, San Antonio, Texas 78201. (512) 699-9260.

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LOWEST PRICES: Crystal Certificates, Two-Meter Certificates $2.75 each. Scanner-Monitor Certificates $2.25 each. Minimum order 10 assorted pieces. Bob Anderson W1LBA, 428 Central Ave., Milton, MA 02186.

WANTED: Schematic and Operating Manual for Hallicrafters Model SBT-20 Xcvr. Loaned material will be Returned, and Postage Expense Reimbursed. Ralph Irish WA8GDT, P.O. Box 122, Utica, Mich. 48087. 313-731-6760.

PC BOARDS HR ARTICLES, Synthesized HF Converter 12/77 $16.00. Digital 2M Scanner 2/78 $8.50. Glass, Plated, Drilled. RTC Electronics, Box 2514, Lincoln, Nebraska 68502.

WANTED — Jennings vacuum variables: UCSL500, UCSL1000, UCSL2000. State price first letter. McMullen, Box 211, Greenville, NH 03048.

SEE OUR AD in this issue, Pyramid Data, page 112.

B&K TEST EQUIPMENT. Free catalog. Free shipping. Dinosaur discounts. Spacetron-CE, 948 Prospect, Elmhurst, IL 60126.

MOTOROLA HT220, HT200, and Pageboy service and modifications performed at reasonable rates. WA4FRV (804) 320-4439, evenings.

AUTHORIZED DEALER for DenTron, KLM, Larsen, Bearcat, etc., Big Catalog 201-962-4695 Narwid Electronics, 61 Bellot Road, Ringwood, N.J. 07456.

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TELETYPEWRITER PARTS WANTED: for all machines manufactured by: Klienschmidt Corp., Teletype Corp. and Mite. Any quantity, top prices paid send list for my quote. Phil Rickson, W4LNW, Rt. 6, Box 1103G2, Brooksville, Fl. 33512.

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QSL CARDS 500/$10. 400 illustrations, sample. Bowman Printing, Dept. HR, 743 Harvard, St. Louis, MO 63130.

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FAST PROFESSIONAL HAM REPAIR Service. New York City area. Amateur extra; FCC Commercial License. Fully equipped shop. Reasonable. Rich Tashner N2EO. 212-352-1214.

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PROP PITCH gasket sets for smaller prop pitch series. Same material as original. $5.00/set. H. Landskov, N7RT, 1050 So. Stanley, #40A, Tempe, AZ 85281.

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THE "CADILLAC" of QSL's! New! Samples: $1.00 (Refundable) — MAC'S SHACK; Box #1171-D; Garland, Texas 75040.

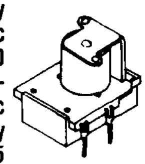
TELETYPEWRITER PARTS, gears, manuals, supplies, tools, toroids. SASE list. Typetronics, Box 8873, Ft. Lauderdale, FL. 33310. N4TT Buy parts, late machines.

EXCLUSIVELY HAM TELETYPE 24th year, RTTY Journal, articles, news, DX, VHF, classified ads. Sample 35¢. $3.50 per year. 1155 Arden Drive, Encinitas, Calif. 92024.

FIGHT TVI with the RSO Low Pass Filter. For brochure write: Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada MIS 3B4.

MANUALS for most ham gear made 1937/1970. Send only 25¢ coin for list of manuals, postpaid. HI, Inc., Box H864, Council Bluffs, Iowa 51501.

NEW ITEMS



Sigma Relay — Factory new 18 ohm coil. 6 Vdc nominal coil voltage @ 60 mA. S.P.D.T. 3 amp contacts 115 Vac or 28 Vdc rating. We find this relay will pull-in as low as 1.5 volts. A real jewel. $1.25 ea.

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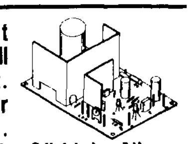
Protect your family — home — and business. These U.L. Approved, ALL METAL units were removed from large apartment complex being torn down. 115 Volt AC input. Sensitivity to smoke is 2% per ft. max. Paint may be scratched but all units tested before shipment. Buy now — Don't delay longer. $8.95 ea. ppd.

Transformer: 115V AC Primary, Secondary 17-0-17V @ 7 Amps. We tested and find good for 10 Amps intermittent duty. Ideal for 2M rigs! $8.00 ea. ppd.



High-gain 8 watt audio amp. 20 mV will drive it to 8 watts out. Rectifiers and filter cap on the board. Size approx. 3" x 4" x 3" high. All you need is 24-0-24 volts ac. Of course we supply schematic. $3.25 ppd.

SO239 Coax Fittings . . . . 50¢ ea.
PL259 Coax Fittings . . . . 50¢ ea.

88 mHy unpotted toroids 5 for $3.50

400 MW Zener diodes. Factory new — full leads. All 5% 3.3 volts, 5.1 volts, 9.1 volts, 11 volts. 10¢ ea. or 100 for $8.00

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Computer Grade Capacitor. 5100 mfd @ 50 volts. Size: 2" dia. x 2-1/4" high. $1.90 ea. ppd.

MA1003 Clock Module $18.95

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ATLAS 210X With Noise Blanker; Deluxe Console; Mike; VOX; Digital Frequency Display; Mobile Connectors. Staley Keener, Rt. 10, Box 782, Hickory, NC. 704-264-5510.

TELETYPES, Models 15-35 $75-up. SASE for list. Goodman, 5454 South Shore, Chicago, IL 60615.

RTTY — NS-1A PLL demodulator. Board $3.50; Parts $15.00; W/T $24.95, all postpaid. SASE for info. Nat Stinnette Electronics, Tavares, Fl. 32778.

FREE Electronics Parts Catalog. OK Electronics, Box 291, Onalaska, WI 54650.

EZ does it best. Deals, that is, on Yaesu, ICOM, Drake, Swan, Cushcraft, Larsen, KLM, DenTron, VHF Engineering and Wilson. For new or used gear call, see or write WØEZ, Bob Smith Electronics, 12 So. 21st St., Fort Dodge, Iowa 50501. (515) 576-3886.

TECH MANUALS for Govt. surplus gear — $6.50 each: SP-600JX, URM-25D, OS-8A/U, TS-173/UR. Thousands more available. Send 50¢ (coin) for 22-page list. W3IHD, 7218 Roanne Drive, Washington, DC 20021.

QSL FORWARDING SERVICE — 30 cards per dollar. Write: QSL Express, 30 Lockwood Lane, West Chester, PA. 19380.

WANTED: Heathkit Oscilloscope OL-1 Schematic. Robert Dole, Route 1, Almena, Kansas 67622.

51J4, Rack Mount, Three Filters, very good $350. Also new 4D32 — $10, 4-125A — $10, 866A — $2, 872 — $5. K6AVF, Box 566, LaMesa, CA 92041.

SELLING OUT — MINT 2KW SSB STATION. Hallicrafters SX117/HT-44/PS-150AC will transceive or separate, 3-400Z (pair) Linear matches above, Turner 254C SSB microphone, Heath HO-13 monitor scope, SWR bridge, Heath Cantenna, B&W electronic T-R switch, grid dipper, all manuals. First $700 takes all. Send $1.00 for color photo. Wade Calvert WA9EZY, Box 276, Port Byron, IL 61275.

TRAVEL-PAK QSL KIT — Send call and 25¢; receive your call sample kit in return. Samco, Box 203, Wynantskill, NY 12198.

THE MEASUREMENT SHOP has used/reconditioned test equipment at sensible prices; catalog. 2 West 22nd St., Baltimore, MD 21218.

RADIO MUSEUM NOW OPEN. Free admission, 15,000 pieces of equipment from 1850 telegraph instruments to amateur and commercial transmitters of the 1920s. Amateur station W2AN. Write for information: Antique Wireless Assn., Main St., Holcomb, NY 14469.

AMATEUR MICROPROCESSOR EXPERIMENTERS: 10 MHz ± 20 ppm Coldweld crystals. 1 ppm/yr. 32 pF. C_0 6 pF. $4.25 ea. postpaid. Savoy Electronics, Inc., P.O. Box 5727, Ft. Lauderdale, FL 33310. 305-563-1333.

WE BUY ELECTRON TUBES, diodes, transistors, integrated circuits, semiconductors. ASTRAL ELECTRONICS, 321 Pennsylvania Avenue, Linden, NJ 07036. 201-486-3300.

AUTHORIZED DISTRIBUTOR F9FT Antennas, Microwave Modules, RIW Products' new tandem reflector, 19 element, 432 MHz Yagi — Radio Clinic — N2MB (formerly WA2BIT) 212-327-4952.

STOP LOOKING for a good deal on amateur radio equipment — you've found it here — at your amateur radio headquarters in the heart of the Midwest. We are factory-authorized dealers for Yaesu, Kenwood, Drake, Collins, ICOM, Ten-Tec, Atlas, Hallicrafters, KLM, Wilson, Regency, Tempo, Swan, Midland, Alpha, Standard, DenTron, Hy-Gain, Mosley, Cushcraft, and CDE, plus accessories. Thousands of thrifty hams from coast to coast already know us and we invite you to join them by writing or calling us today for our low quote and trying our personal and friendly Hoosier service. HOOSIER ELECTRONICS, P.O. Box 2001, Terre Haute, Indiana 47802. 812-238-1456.

More Details? CHECK—OFF Page 150

More Details? CHECK—OFF Page 150

Coming Events

SPRING SWAPFEST OF THE MILWAUKEE UHF SOCIETY, Sunday, May 14, 1978, starting at 7 AM on the grounds of the Waukesha Co. Expo Center. Indoor space available on an advanced reservation basis at $3.00 per table. Admission to the grounds: $1.50 Advance, $2.00 at Gate. Prizes! Beer and Brats! Directions: I-94 to Waukesha Co. F, South to FT, west to Expo. Write: Swapfest, P.O. Box 49, North Prairie, Wisconsin 53153.

7TH ANNUAL MARC (MILTON AMATEUR RADIO CLUB) HAMFEST, 8 AM to 5 PM, June 4, 1978, rain or shine, at the Allenwood Fireman's Fairground on U.S. Rt. 15, 4 miles north of Interstate 80. Advanced registration for sellers is $2.50; at the gate $3.00, XYLs and children free. Flea market, auction and contests. Cash door prizes, free portable and mobile FM clinic, and supervised childrens activities. Indoor area available, food and beverages at reasonable prices, talk-in on 37/97, 34/94, and 52 simplex. Camping and motels nearby. For further details contact Jerry Williamson WA3SXQ, 10 Old Farm Ln., Milton, PA 17847. 717-742-3027.

HAMPDEN COUNTY RADIO ASSOCIATION ANNUAL FLEA MARKET, Friday, May 5th, at the Feeding Hills Congregational Church, Feeding Hills, Mass. Doors open at 7:30 PM. Free table to members of ARRL affiliated clubs. $1.00 for non-members, $2.00 for a business. .25¢ admission charge puts you in a raffle! For information: Jeffrey J. Duquette, K1BE, P.O. Box 346, Southwick, MA 01077. 413-569-6739.

THE CHAMPAIGN LOGAN AMATEUR RADIO CLUB'S ANNUAL HAMFEST, Sunday, May 14, 1978 at West Liberty Lions Park, West Liberty, Ohio. Free admission. Trunk and table sales $1.00. Door prizes. Talk-in will be on 146.52.

23RD ANNUAL WEST COAST VHF CONFERENCE, May 12-14, 1978, Stardust Hotel, Las Vegas, Nevada. Technical program by the San Bernardino Microwave Society. Hospitality Room, informal technical and operating sessions, noise-figure measurements contest, prize drawing, entertainment! Advance registration $4.00 ($5.00 at the door). Reservation cards and one-night's deposit must be received before Friday, April 14, 1978. Details from West Coast VHF Conference, 510 South Rose Street, Las Vegas, Nevada 89106.

CLARK COUNTY AMATEUR RADIO CLUB'S Annual Ft. Vancouver Hamfair, May 13-14th, at the Clark County Fairgrounds off Interstate 5, just north of Vancouver, Washington. Registration $3.00 per person. Preregistration by May 5th nets extra drawing ticket. Grand prize: Kenwood TS 820S transceiver. Dinner: $4.25 for adult, $2.00 for children under 12. Pancake breakfast served on Sunday for donation only. Activities for hams and families. Camping with electricity available $2.50 per night. Make checks payable to Ft. Vancouver Hamfair for registration and dinner tickets. Mail to Jack Ellis, K7SUQ, 9610 SE 6th St., Vancouver, Washington 98664.

24TH ANNUAL BREEZE SHOOTERS HAMFEST, Sunday, May 21st, White Swan Park, Parkway West (Rt. 60), near the Greater Pittsburgh International Airport. Prizes, homebrew contest, refreshments, amusement park for harmonics, (discount ride tickets available at hamfest). Western Pennsylvania's largest ham event. Admission, flea market and parking FREE! Talk-in on 29.0 and 28/88. Contact WA3LUM, 311 Evergreen Ave., Pittsburgh, PA 15209.

1978 DELTA A.R.R.L. CONVENTION, May 6-7, 1978, Baton Rouge, LA. Riverside Centroplex 30,000 Square feet of air conditioned exhibition area. 12,000 Square feet of meeting room. Convention Hotel is The Chateau Capitol, 1 1/2 block walk. Forums on antennas, microprocessor-repeater application, beginning help and novices, FCC exams and much more. A.R.R.L. Guest Harry Dannals, Perry Williams and Gerald Hall. For details, write: Baton Rouge Amateur Radio Club, 10715 Waverland, Baton Rouge, Louisiana 70815.

SATELLITE AMATEUR RADIO CLUB Annual Swap/Fun fest and Santa Maria BBQ on Sunday, June 18. Best steak and biggest hamfest in the west. Fantastic prizes! Swap tables available. All you can eat dinner — $6.00 adults; $3.00 children under 12. Contact W2KVA/6 at (805) 925-0398, or write SWAPFEST, P.O. Box 2531, Orcutt, CA 93454.

AMATEUR FAIR '78: SWAPFEST & EXPOSITION for Amateur Radio Operators and Computer Hobbyists. Saturday, June 3rd. Minnesota State Fairgrounds. Free overnight parking for self-contained campers, June 2 only. Talk-in on 16/76 and 52/52. Sell from your car in the GIANT FLEA MARKET. Inside space available. Great prizes, scheduled forums on FM and microprocessors. Admission: $2.00. For information or reservations for commercial exhibit space, call (612) 933-2823.

More Details? CHECK—OFF Page 150

flea market

TRI-COUNTY ARC ANNUAL HAMFEST May 7. 9:00 A.M. to 4:00 P.M. Rain or Shine — Indoors at Stirling, NJ Youth Center, just off Valley Road. Tables $5.00 Buyers $1.00. Door Prize — Bird Wattmeter. Talk-in on 146.52 and 147.855-147.255. Info from Tri-County ARC P.O. Box 412, Scotch Plains, NJ 07076 or call Herb W2CHA 201-647-3461.

HOSSTRADERS NET: Fifth annual Tailgate Swapfest Saturday, May 13th at Deerfield, New Hampshire Fairgrounds. (Covered buildings in case of rain.) Admission one dollar, no commission or percentage. Commercial dealers welcome at same rate. Excess revenues benefit Boston Burns Unit of Shriner's Hospital for Crippled Children. Last year we donated $430.80. Talk-in .52, 146.40-147.00, 3940 kHz. Questions: S.A.S.E. to Joe Demaso K1RQG, Star Rt. Box 56, Bucksport, ME 04416 or Norm Blake WA1IVB, P.O. Box 32, Cornish, ME 04020 or Check Hosstraders Net on Sundays 4 PM 3940 kHz.

KENTUCKY HAM-O-RAMA — Sunday, May 28 (Memorial Day Weekend). 7 minutes south of Cincinnati. Erlanger Lions Park, Erlanger, Kentucky. Donaldson Road exit, I-75 South. Prizes, exhibits, flea market. NKARC, Box 31, Ft. Mitchell, Kentucky 41017.

F.M. B*A*S*H, DAYTON, OHIO, April 28, 1978, Friday night of DAYTON HAMVENTION. Social evening for hams and friends, 8 P.M til midnight. *NEW LOCATION*: Downtown Dayton Convention Center, Main @ Fifth. Admission is free. Sandwiches, beverages, snacks and C.O.D. bar available. Live entertainment by TV personality Rob Reider (WA8GFF) and his group. Fabulous prize drawing featuring a complete Drake UV-3, including 144, 220 & 440 MHz synthesized modules, power supply, encoder mike and antenna plus other prizes. Winner of first prize need not be present. For further information contact: Miami Valley F.M. Assn. c/o Sue Hagedon, WB8GWQ, 1340 Brainard Woods Drive, Dayton, Ohio 45459.

THE 4TH ANNUAL NORTHWESTERN PENNSYLVANIA HAMFEST, May 6th, Crawford County Fairgrounds, Meadville, PA. Gates open at 8:00 $2 prize ticket required for admission — $1 to display. Children FREE. Hourly door prizes, refreshments, commercial displays welcome. Indoors if rain. Talk-in 04/64 and 52. Details CARS, P.O. Box 653, Meadville, PA. 16335.

THE WABASH COUNTY AMATEUR RADIO CLUB'S 10th Annual Hamfest is Sunday, May 21, 1978, rain or shine, at the Wabash County 4-H Fairgrounds in Wabash. Large flea market (no table or setup charge). Technical forums, bingo, free parking, good food at reasonable prices. Advance admission, $2.00; $2.50 at gate. Children under 12 free. Write Dave Nagel WD9BDZ, 555 Valley Brook Lane, Wabash, IN 46992.

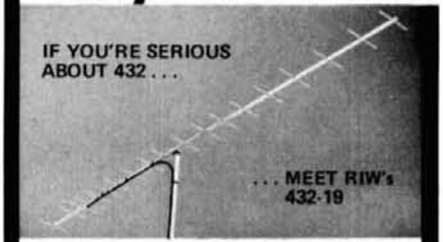
WARMINSTER AMATEUR RADIO CLUB'S Fourth Annual "HAMMART", Flea Market and Auction, Sunday, May 14th, from 9 to 4 at William Tennent Senior High School, Street Road (Route 132) 2 miles East of York Road (Route 263), Warminster, Bucks County, PA. Registration $1.00, Tailgating $2.00 additional. No indoor selling, bring your own tables. Talk-in on 146.16-76 and 146.52. Further information write Horace Carter K3KT, 38 Hickory Lane, Doylestown, PA 18901 or call 215-345-6816.

THE POTOMAC AREA VHF SOCIETY Seventh Annual Hamfest, Sunday, May 7, 1978, from 8 AM to 5 PM at the Howard County Fairgrounds, approximately 25 miles north of Washington, D.C. at the intersection of I-70 and Maryland Route 32. Registration of $3 includes flea market or tailgate sales. Professional food and beverage catering and unlimited parking will be available. Talk-in on 146.52. For further information contact K3DUA, K4LHB or WA3NZL.

MARYLAND: Fourth Annual Easton ARS hamfest, May 14th, 10 AM to 4 PM at the Talbot County Agricultural Center located 5 miles north of Easton on Route 50, between mile markers 60 and 61. From Baltimore or D.C., cross Chesapeake Bay Bridge and follow Route 50 East for 21 miles. Talk-in on 52 simplex or 146.445/147.045. Some tables available inside and outside, or bring own. Lots of space for tables and tailgaters at $2 in addition to regular Donation of $2 to enter. Write or call K3ONU, Box 781, Easton, MD 21601; 301-822-0943 after 6 PM.

VACATIONLAND HAMFEST, Erie County Fairgrounds, Sandusky, Ohio, May 21, 1978. Tables $4 each, $1 for trunk sales (8 acres for trunk sales). Advance tickets $1.50, $2.00 at gate. Talk-in on 52/52. Write EARS, P.O. Box 2037, Sandusky, OH 44870.

TENNESSEE: The Annual Humboldt Amateur Radio Club Hamfest Sunday, May 21 at Shady Acres City Park in Trenton, TN. Talk-in on 37/97. Flea Market, Prizes, Ladies activities, light lunches. For further information, contact Ed Holmes W4IGW, 501 N. 18 Ave., Humboldt, TN 38343.



More Details? CHECK—OFF Page 150

DIODES/ZENERS

1N914	100v	10mA	.05
1N4005	600v	1A	.08
1N4007	1000v	1A	.15
1N4148	75v	10mA	.05
1N753A	6.2v	z	.25
1N758A	10v	z	.25
1N759A	12v	z	.25
1N4733	5.1v	z	.25
1N5243	13v	z	.25
1N5244B	14v	z	.25
1N5245B	15v	z	.25

SOCKETS/BRIDGES

8-pin	pcb .25	ww	.45
14-pin	pcb .25	ww	.40
16-pin	pcb .25	ww	.40
18-pin	pcb .25	ww	.75
22-pin	pcb .45	ww	1.25
24-pin	pcb .35	ww	1.10
28-pin	pcb .35	ww	1.45
40-pin	pcb .50	ww	1.25

Molex pins	.01	To-3 Sockets	.45
2 Amp Bridge		100-prv	1.20
25 Amp Bridge		200-prv	1.95

TRANSISTORS, LEDS, etc.

2N2222A	NPN (2N2222 Plastic .10)	.15
2N2907A	PNP	.15
2N3906	PNP (Plastic)	.10
2N3904	NPN (Plastic)	.10
2N3054	NPN	.35
2N3055	NPN 15A 60v	.50
T1P125	PNP Darlington	.35
LED Green, Red, Clear, Yellow		.15
D.L.747	7 seg 5/8" High com-anode	1.95
XAN72	7 seg com-anode (Red)	1.25
MAN71	7 seg com-anode (Red)	1.25
MAN3610	7 seg com-anode (Orange)	1.25
MAN82A	7 seg com-anode (Yellow)	1.25
MAN74A	7 seg com-cathode (Red)	1.50
FND359	7 seg com-cathode (Red)	1.25

C MOS

4000	.15
4001	.15
4002	.20
4004	3.95
4006	.95
4007	.35
4008	.95
4009	.45
4010	.45
4011	.20
4012	.20
4013	.40
4014	.95
4015	.90
4016	.35
4017	1.10
4018	1.10
4019	.50
4020	.85
4021	1.00
4022	.85
4023	.25
4024	.75
4025	.30
4026	1.95
4027	.50
4028	.95
4030	.35
4033	1.50
4034	2.45
4035	1.25
4040	1.35
4041	.69
4042	.95
4043	.95
4044	.95
4046	1.75
4049	.45
4050	.45
4066	.95
4069	.40
4071	.35
4081	.70
4082	.45
MC 14409	14.50
MC 14419	4.85

– T T L –

7400	.15
7401	.15
7402	.20
7403	.20
7404	.15
7405	.25
7406	.35
7407	.55
7408	.25
7409	.15
7410	.10
7411	.25
7412	.30
7413	.35
7414	1.10
7416	.25
7417	.40
7420	.15
7426	.30
7427	.45
7430	.15
7432	.30
7437	.30
7438	.35
7440	.25
7441	1.15
7442	.45
7443	.65
7444	.45
7445	.65
7446	.95
7447	.95
7448	.65
7450	.25
7451	.25
7453	.20
7454	.25
7460	.40
7470	.45
7472	.40

7473	.25
7474	.30
7475	.35
7476	.40
7480	.55
7481	.75
7483	.95
7485	.75
7486	.25
7489	1.35
7490	.55
7491	.95
7492	.95
7493	.35
7494	.75
7495	.60
7496	.80
74100	1.15
74107	.35
74121	.35
74122	.55
74123	.55
74125	.45
74126	.35
74132	1.35
74141	.90
74150	.85
74151	.65
74153	.75
74154	.95
74156	.95
74157	.65
74161	.85
74163	.85
74164	.60
74165	1.50
74166	1.35
74175	.80

74176	1.25
74180	.75
74181	2.25
74182	.95
74190	1.75
74191	1.05
74192	.75
74193	.85
74194	1.25
74195	.95
74196	1.25
74197	1.25
74198	2.35
74221	1.00
74367	.85
75108A	.35
75110	.35
75491	.50
75492	.50
74H00	.15
74H01	.25
74H04	.20
74H05	.20
74H08	.35
74H10	.35
74H11	.35
74H15	.45
74H20	.30
74H21	.25
74H22	.40
74H30	.20
74H40	.25
74H50	.25
74H51	.25
74H52	.15
74H53J	.25
74H55	.20

74H72	.45
74H101	.75
74H103	.75
74H106	.95
74L00	.25
74L02	.25
74L03	.30
74L04	.30
74L10	.30
74L20	.35
74L30	.45
74L47	1.95
74L51	.45
74L55	.65
74L72	.45
74L73	.40
74L74	.45
74L75	.55
74L93	.55
74L123	.85
74S00	.35
74S02	.35
74S03	.30
74S04	.30
74S05	.35
74S08	.35
74S10	.35
74S11	.35
74S20	.35
74S40	.20
74S50	.20
74S51	.25
74S64	.20
74S74	.35
74S112	.60
74S114	.65

74S133	.40
74S140	.55
74S151	.30
74S153	.35
74S157	.75
74S158	.30
74S194	1.05
74S257 (8123)	1.05
74LS00	.25
74LS01	.35
74LS02	.35
74LS04	.30
74LS05	.45
74LS08	.25
74LS09	.35
74LS10	.35
74LS11	.35
74LS20	.25
74LS21	.25
74LS22	.25
74LS32	.40
74LS37	.35
74LS40	.45
74LS42	1.10
74LS51	.50
74LS74	.65
74LS86	.65
74LS90	.95
74LS93	.95
74LS107	.85
74LS123	1.00
74LS151	.95
74LS153	1.20
74LS157	.85
74LS164	1.90
74LS367	.75
74LS368	.75
74C04	.25
74C151	2.25

LINEARS, REGULATORS, etc.

MCT2	.95
8038	3.95
LM201	.75
LM301	.45
LM308 (Mini)	.95
LM309H	.65
LM309K (340K-5)	.85
LM310	1.15
LM311D (Mini)	.75
LM318 (Mini)	.95
LM320K5 (7905)	1.65
LM320K12	1.65

LM320T5	1.65
LM320T12	1.65
LM320T15	1.65
LM324N	.95
LM339	.95
7805 (340T5)	.95
LM340T12	1.00
LM340T15	1.00
LM340T18	1.00
LM340T24	.95
LM340K12	1.65

LM340K15	1.25
LM340K18	1.25
LM340K24	.95
78L05	.75
78L12	.75
78L15	.75
78M05	.75
LM373	2.95
LM380 (8-14 PIN)	.95
LM709 (8, 14 PIN)	.25
LM711	.45

LM723	.50
LM725N	2.50
LM739	1.50
LM741 (8-14)	.25
LM747	1.10
LM1307	1.25
LM1458	.95
LM3900	.50
LM75451	.65
NE555	.50
NE556	.95
NE565	.95
NE566	1.75
NE567	1.35

9000 SERIES

9301	.85	95H03	1.10
9309	.35	9601	.45
9322	.75	9602	.45

MICRO'S, RAMS, CPU'S, ETC.

74S188	3.00
1702A	4.50
MM5314	3.00
MM5316	3.50
2102-1	1.45
2102L-1	1.75
TR1602B	4.50
TMS 4044-45NL	14.50
8080AD	12.00
8T13	1.50
8T23	1.50
8T24	2.00
8T97	1.00
2107B-4, A	4.00
2708	11.50

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Total Order	Deduct
$35 - $99	5%
$100 - $300	10%
$301 - $1000	15%
$1000 - Up	20%

flea market

CASS COUNTY AMATEUR RADIO CLUB HAMFEST Sunday, May 7, 1978 7 AM to 4 PM at the 4-H Fairgrounds. Go north of Logansport on highway 25 approximately one mile, turn right, follow the QSY signs. Advance tickets $1.50, $2.00 at the gate. Outside setup free, undercover $1.00. Bring your own tables. Talk-in 146.52 & Logansport repeater 147.78/.18. Write to K9DVL Dave Rothermel, RFD 4, Box 146 G, Logansport, Indiana 46947.

CENTRAL MICHIGAN Amateur Repeater Association Fourth Annual Swap & Shop in Midland, Michigan, June 3rd at the Midland County Fairgrounds. Camping & Program Friday evening. Computer Demonstrations, Door Prizes. Donation: $2.00 at door, $1.50 in advance. Talk-in: 146.07 - 146.67 WR8AKN, 146.13 - 146.73 WR8AHM and 146.52 Simplex. Info & Tickets, SASE to Don Zahm, WB8SDJ, 3871 Monroe R#8, Midland, MI 48640.

SOCIETY OF WIRELESS PIONEERS (SOWP) 10th birthday on-the-air CW QSO Party during the full GMT period of May 4 and 5, 1978. Call CQ SOWP, all bands, 55 kHz up from the low end. Novice members should use the center portion of each novice band. Part-time participants should call CQ on the even hours. Exchange of information should include handle, SOWP membership number and QTH. Additional information is optional. A special certificate will be awarded to all members who make a minimum of 10 CW contacts with fellow members. Send a list of contacts (with SASE) showing date, time, call and SOWP numbers to the Society's Vice President for Awards, Pete Fernandez W4SM, 129 Hialeah Road, Greenville, South Carolina 29607.

THE TERRY COUNTY AMATEUR RADIO CLUB annual swapfest May 7, 1978 in the National Guard Armory, Brownfield, Texas. For more information contact Viola Simmons, W5FBM, 1603 East Tate, Brownfield, Texas 79316.

INDIANA: THE MIDWEST REPEATER ASSOCIATION'S FIRST ANNUAL HAMFEST Sunday, May 21, at the National Guard Armory, 2530 173rd St., Hammond, Indiana. Doors open 7 AM. Advance tickets $2.00, $2.50 at the door. Reserved flea market area $1.00 otherwise first come, first served. Door prizes, raffles, equipment displays. Talk-in 146.31/91 and 146.52 Simplex. Write to M.R.A., P.O. Box 342, Griffith, IN 46319.

ANNUAL TEXAS VHF-FM SOCIETY SUMMER CONVENTION, hosted by the Houston Echo Society, August 4, 5, 6, 1978 at the Galleria Plaza Hotel off Interstate Loop 610 at Westheimer Road. Microprocessors/microcomputers, hidden transmitter hunt, OSCAR communications, VHF-FM activities. ARRL & FCC forums, open hospitality suite, ladies' activities, Astrodome-Astroworld tours for the kids, Exhibitors, and prizes. Saturday night banquet featuring Bill Tynan, W3XO, editor of *QST*'s "World Above 50 MHz", as guest speaker. For information and reservations write FM Society Summer Convention, P.O. Box 717, Tomball, Texas 77375.

JUNE 4, 1978, STARVED ROCK RADIO CLUB HAMFEST, Bureau County Fairgrounds, Princeton, Illinois. Advance registrations $1.50 before May 25, after $2.00. Large SASE please for registrations, map, information, etc. W9MKS/WR9AFG/SRRC, RFD #1, Box 171, Oglesby, Illinois 61348. 815-667-4614.

MISSOURI: ST. LOUIS REPEATER CLUB HAMFEST, Sunday, May 21, 1978, at Jefferson Barracks Park. Talk-in 146.34/.94 and 146.52/.52. Refreshments available, prizes, Advance tickets $1.50, donations $2.00 at door. Info, contact WB0SVS, Bob Evans, 22 Brixworth, Florissant, MO 63033.

MANASSAS HAMFEST SPONSORED BY The "Ole Virginia Hams" A.R.C. June 4, 1978 at the Prince William County Fairgrounds one-half mile south of Manassas, Virginia on Rt. 234. Gates open 7 AM for tailgating, 8 AM for general admission. Fantastic Prizes. Admission $3.00 adult, under 12 free. Tailgating $2.00 per vehicle, over 300 spaces available. Refreshments, YL Program, Children's entertainment. FM Clinic; QSL Bureaus: learn how they work. CW Proficiency awards: 5 wpm up. Indoor exhibit space available for Dealers: for info contact Sam Lebowich, 9512 Sudley Manor Dr., Manassas, VA 22110. Talk-in on 146.37/146.97, 147.84/147.24 and CB Ch. 1. Accommodations: Olde Towne Inn in Manassas, Holiday Inn at I-66 and Rt. 234 Interchange. Camping at Prince William County Forest (on Rt. 234 near the intersection of 234 and I-66).

HAMFESTERS 44TH ANNUAL PICNIC AND HAMFEST, Sunday, August 13, 1978 at Santa Fe Park, 91st and Wolf Road, Willow Springs, Illinois, Southwest suburb of Chicago. Exhibits for OM's and XYL's, FAMOUS SWAPPERS ROW. Tickets at gate $2.00, Advance $1.50. For Hamfest info or Advance Tickets (send check or money order — SASE appreciated) to Bob Hayes W9KXW, 18931 Cedar Ave., Country Club Hills, IL 60477.

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NEW OMNI-J 2-meter mobile or portable antenna. 3/8" thread, 5-dB gain (1.5-dB gain over conventional 5/8-wave mobile whip antenna). **$29.95**; 220-MHz **$27.95**. 450 MHz **$27.95**. Guaranteed results.

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* will operate with slight degradation at 142-150 MHz.

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SSB-30 Amps @ 300W PEP

DUTY CYCLE:
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SSB 60% @ 150W 50% @ 250W

Model 2002 $479.00

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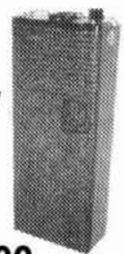

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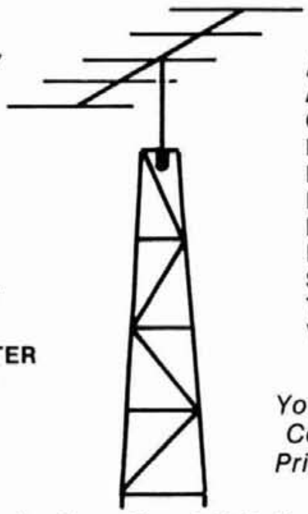

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25 watts	—	25A	25C	25D	25E
50 watts	50H	50A	50C	50D	50E
100 watts	100H	100A	100C	100D	100E
250 watts	250H	250A	250C	250D	250E
500 watts	500H	500A	500C	500D	500E
1000 watts	1000H	1000A	1000C	1000D	1000E
2500 watts	2500H				
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Headphone Impedance	32 20 ohms	2000 ohms	32 20 ohms	32 20 ohms	32 20 ohms	32 20 ohms	32 20 ohms	32 20 ohms
Microphone Frequency Response					50 8000 Hz	50 8000 Hz	50 8000 Hz	50 8000 Hz
Microphone Impedance					High	High	High	High
Microphone Sensitivity Below 1 volt/microbar at 1kHz					51dB ±5dB	51dB ±5dB	51dB ±5dB	51dB ±5dB
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Model C 610 Model C 1320 Model CM 610 Model C 1210

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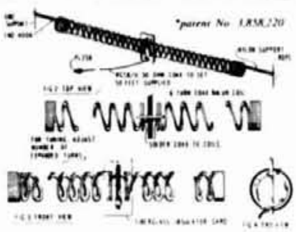

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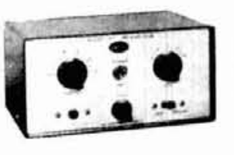

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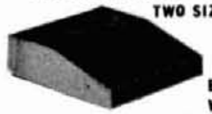

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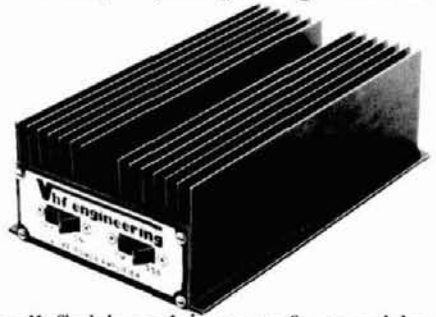

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1 200 pfd wide-spaced variable with isolantite insulation rated 3,000 volts

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1 pvc insulated shaft couplings ¼ to ¼

3 SO-239 coax chassis connectors. Tunes 52 ohm or 52-300-600* or random wires

1 heavy inductance for 10-15-20-40-80 meters
6 pvc stand-offs, 4 for condensers and 2 for inductance
1 HD switch for band catching 10 thru 80 meter coverage
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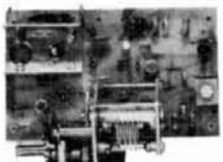

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12 VAC XFMR for 110 VAC 1.50
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- ½" 4 DIGIT LED READOUT
- Adjustable tone & duration on chime
- Simulated swinging LED Pendulum
- All CMOS Construction
- Quality plated boards
- All electronics switches and XFMR included

39.95 KIT **59.95** ASEMB

Optional Solid Hardwood Case for MG-01 Clock over 10 inches tall. Includes Ruby front. Unassembled.
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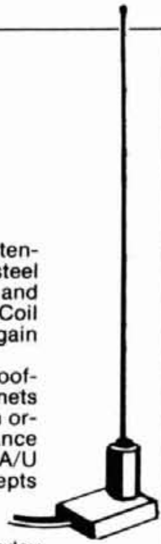

LARSEN MM-LM-150K

The ⅝ wave length 2m antenna has a 49" stainless steel rod, quadruple plated, and coil for 144-174 MHz. Coil threads: 5/16" x 24. 3 dB gain over ¼ wave.
The mount has a low, roof-hugging silhouette, magnets guaranteed permanent in ordinary use. Full capacitance coupling, 12 ft. of RG-58 A/U coax and connector, accepts 5/16" x 24 threads.

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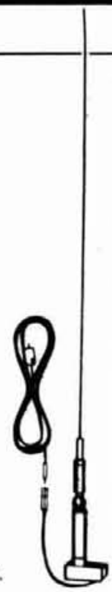

HUSTLER BBLT-144 ⅝ antenna

BBLT-144 features: • 3.4 dB gain over ¼ wave antennas • 200 watts FM power rating • Whip & mount, 52" high • 1.1:1 typical SWR at resonance • Sturdy stainless steel impact spring • 17-7 stainless radiator • Trunk lip mount, easy installation, no drilling required

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Features: • Taper swaged seamless aluminum construction • Omnidirectional performance • Auto band switching • Wide band performance with one setting • Hy-Q traps with large diameter coils • 12" double-grip mast bracket • Low angle radiation • SO-239 connector • 50 ohm input impedance.

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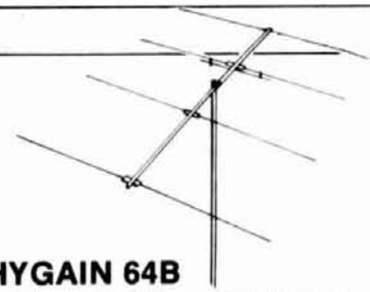
HyGain 14 RMQ roof mount kit for 14AQV/WB **26.06**



HYGAIN 64B 4-element 6 meter beam

The 64B has a unidirectional pattern and special VHF Beta Match configuration, plus: • 12.7 dB gain • 20-25 dB F/B ratio • 1 KW input • SWR less than 1.5:1 • 52 ohm impedance • 9'11" longest element • 12' boom length • 8' turning radius • 80 MPH wind load, 28.2 lbs. • 100 MPH max. wind survival • Net wt. 10 lbs. • 1.1 sq. ft. surface area.

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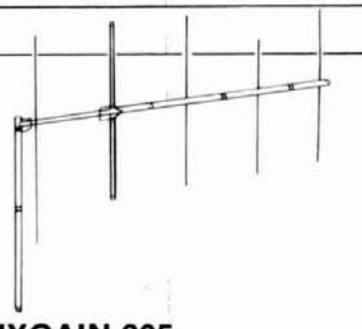

HYGAIN 205 5-element 2m beam

The 205 has 5 elements, optimum spaced on a 75" boom, plux: • Gain (average), 11.3 dBi & 9.1 dBd • F/B ratio, 20 dB • Max. SWR, 2:1 • 4 MHz band width • Max. power, 500 watts PEP • Longest element, 39⅝" • Net wt. 2.9 lbs. • Turning radius, 73".

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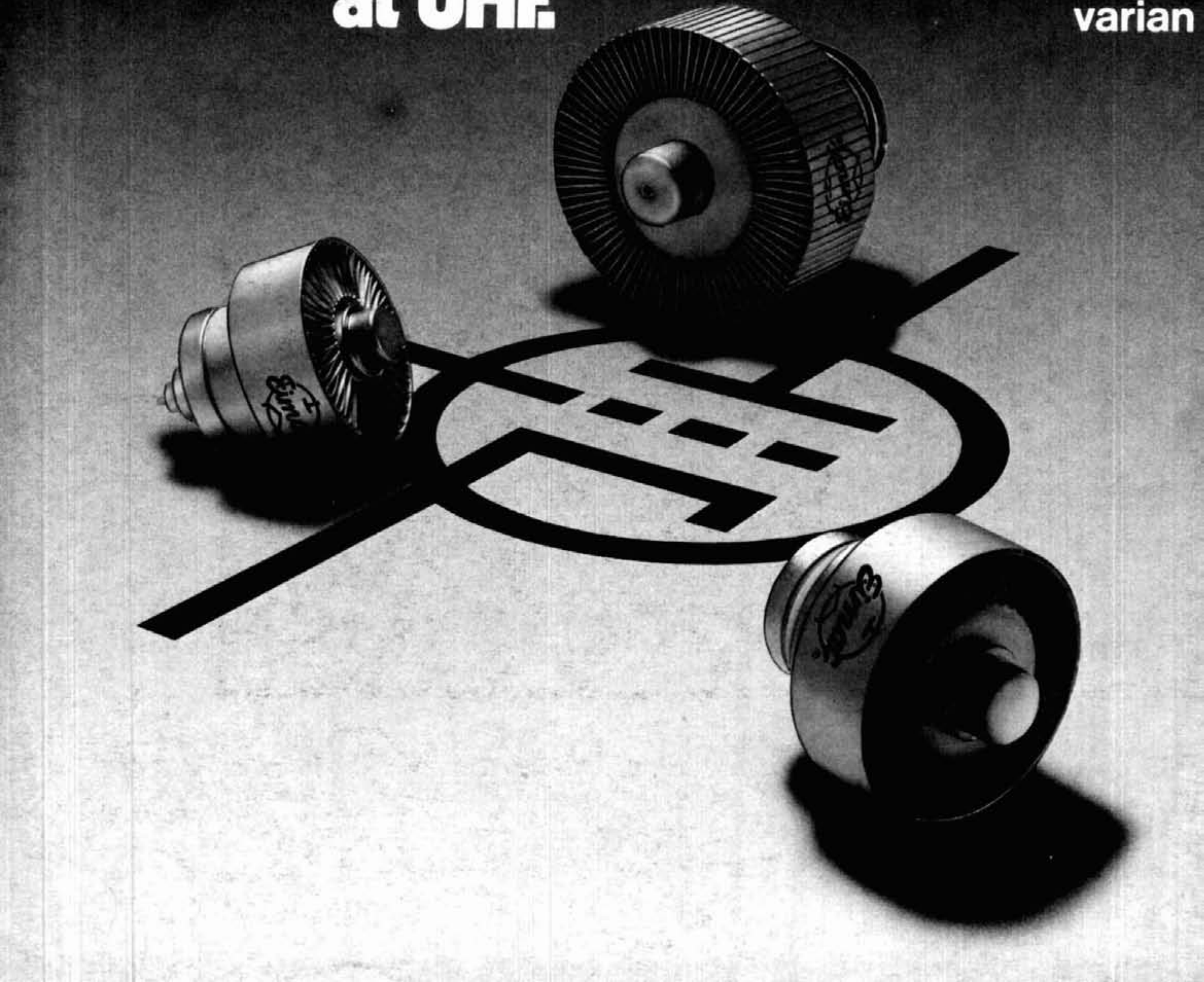


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