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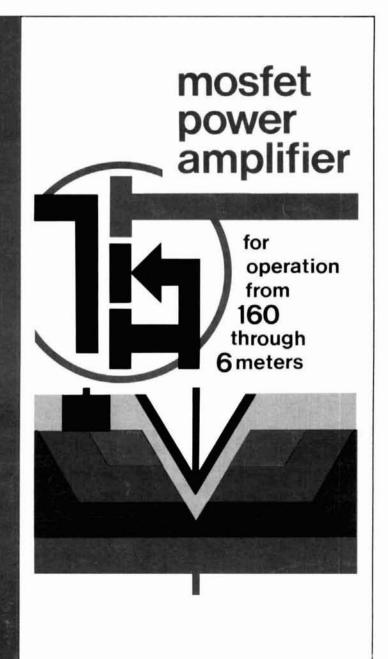
magazine



NOVEMBER 1978

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and much more . . .





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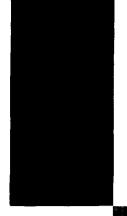
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NOVEMBER 1978

volume 11, number 11

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ham radio magazine is published monthly by Communications Technology, Inc Greenville, New Hampshire 03048 Telephone: 603-878-1441

subscription rates

United States: one year, \$12.00 two years, \$22.00; three years, \$30.00

Canada: one year, \$14.00 two years, \$26.00; three years, \$36.00

Europe, Japan, Africa: (via Air Forwarding Service) one year, \$25.00

North America, South America, Australia and Asia (except Japan): (via Surface Mail) one year, \$18.00

foreign subscription agents

Foreign subscription agents are fisted on page 117

Microfilm cooles university Microfilms, International
Ann Arbor, Michigan 48106
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Cassette tapes of selected articles from ham radio are available to the blind and physically handicapped from Recorded Periodicals 919 Walnut Street, 8th Floor Philadelphia, Pennsylvania 19107

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Second-class postage paid at Greenville, N.H. 03048 and at additional mailing offices Publication number 233340

ham radio

magazine

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November 27th marks the 55th anniversary of one of Amateur Radio's most memorable events - the first two-way amateur communications across the Atlantic Ocean. It was a hard-won goal, its path marked with failure and frustration, but when the Atlantic, at last, had been spanned, it was conquered by short-wave Amateur Radio, on wavelengths that were considered by the "professionals" to be useless.

The first transatlantic tests, in December, 1920, were a dismal failure, as were another series of tests conducted in February, 1921. The 250 or so British stations which were listening for prearranged signals from the United States on a wavelength of 200 meters (1.5 MHz) jammed each other so badly with radiations from their own regenerative receivers that they couldn't hear any signals from across the pond!

A second transatlantic test was scheduled for December, 1921. In November, Paul Godley, 2XE, designer of the famous Paragon receiver, sailed from New York with two receivers under his arm one a standard regenerative set with two stages of audio amplification, the other a 10-tube Armstrong superheterodyne built especially for the tests.

With this superhet and a wire antenna 390 meters long installed on the Androssan moor on the coast of Scotland, not far from Glasgow, Godley heard the first stateside signals coming through in the wee hours of the morning on December 8th. Over the next few days Godley copied the signals of 27 different American amateur stations; on December 12th he received the first complete message from the United States to Europe on the "short waves." British amateurs also participated in the test, and when all the reports were analyzed it was discovered that W.F. Burne, G2KW, had actually made the *first* positive identification of an American amateur signal.

During similar tests a year later, two European stations, F8AB in Nice and G5WS in London, were heard by amateurs along the east coast of the United States. Many American signals were logged in Europe, but two-way communications were not established (but probably only because no one took the initiative to try).

A fourth series of transatlantic tests were scheduled for late 1923. However, these carefully laid plans were totally upset by the enterprise of one man, Leon Deloy, F8AB. Deloy came to the States during the summer of 1923 when he met with John Reinartz, 1XAM, and Fred Schnell, 1MO. Deloy picked up much valuable advice from his talks with Reinartz and Schnell, and before returning to France he acquired a new Grebe receiver and the details of a "trick" circuit which, he was told, would "go down to about 100 meters."

Deloy put his new 100-meter station on the air in early autumn, and, having satisfied himself that everything was in working order, cabled Schnell that he would transmit on 100 meters between 0200 and 0300 GMT on November 26, 1923. The signals from F8AB were heard by Schnell and Reinartz almost from the first dot he transmitted, but the Americans were not ready to transmit back. Unlike Deloy, who presumably did not think it was necessary to seek official permission to transmit on such a short wavelength. Schnell had to obtain the authority from the Radio Supervisor in Boston.

On November 27th, Schnell received special permits from Boston for himself and Reinartz; late that night they were both on the air. For an hour Deloy called the United States and then sent two messages. At 0330 GMT he signed off and asked for acknowledgment. Long calls followed from 1MO and 1XAM. Then came the eagerly awaited reply — Deloy had clearly heard both stations. Reinartz was asked to stand by as Deloy transmitted to Schnell, "R R QRK UR SIGS QSA VERY ONE FOOT FROM PHONES ON GREBE FB OM HEARTY CONGRATULATIONS THIS IS A FINE DAY - PSE QSL." It was, indeed, a fine day.

> Jim Fisk, W1HR editor-in-chief

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Donley, N6CD

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Clark, WD6AVT

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Horace, K3KT

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Gary, WD8CAO

"Very pleased with unit."

Kenneth, W3EMT

"Excellent Equipment."

Tom, WB7SAB

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Edward, W6OSV

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radiation hazards

Dear HR:

In the early 1960s I was civilian chief of a Radiation Hazards and Measurements Branch of European GEEIA - an arm of USAF. The branch had about 15 contract engineers and an equal number of service personnel checking RFI complaints, siting, and orientation accuracy of tropo and radar antennas, and noise measurements for prospective receiver sites at bases from Ireland to India. I seem to remember one or two radars rated at 3 megawatts!

Other than our concern with the possible (and to us tenuous) effects on personnel, there was the very real possibility that by some freakish concatenation of circumstances a radar beam might set off a squib on a bomb in an adjacent munitions dump, and touch off an international incident.

USAF used the 10 mW/cm² minimum, of course, but in deference to host governments, we had fences and warning signs installed at distances such that field intensity was perhaps one-half that minimum. Thus, we hoped that some peasant would not herd his sheep too close to a parabola and claim their milk had been soured.

Brodeur's article, which I read in the New Yorker, shook me somewhat. I was conditioned by an earlier series, also in the New Yorker, on the hazards posed by monster 200,000ton supertankers - a hazard which has been brought to our attention rather frequently of late. Over a period of a month I pored over the

composition of a letter to Mr. Brodeur; I didn't want to write anything challenging or in any way derogatory. I even had the local library try to ascertain whether Mr. Brodeur has academic credentials.

After reading W1HR's editorial in the August issue. I feel that Mr. Brodeur's articles and book must be relegated to the category of "The Bermuda Triangle." Both have been written more for shock effect than for any aim at veracity or objectivity.

> Josef Darmento, W4SXK Merritt Island, Florida

realizable set of design specifications. Note that until these pole frequencies are known, nothing can be designed. It is the only published program I know of which solves for these frequencies for any network. Computer hobbyists may wish to translate it to home computer BASIC.

I would be happy to send copies of the original program listing to ham radio readers upon receipt of a selfaddressed, stamped envelope.

> Allan Llovd, W2ESH 15 Greenwood Avenue Hawthorne, New Jersey 07506

phasing networks Dear HR:

In reference to the article on "SSB Phasing Techniques" in the January 1978, issue of ham radio, and WB9YEM's comments in the August issue, please note that I have published two articles on 90-degree phase-difference networks which may be of interest to your readers. The first appeared in Electronic Design in 1971* and described a simple, four-pole, 90-degree network consisting of four capacitors, six resistors, and one dual op amp; this is an improvement over WB9YEM's circuit by four resistors and seven years. It is also easier to calculate the element values from a given set of pole frequencies for this circuit.

In 1976, Electronic Design published a program in BASIC† which calculate the required pole frequencies for the A and B networks for any

windom antenna

Dear HR:

The article on the Windom antenna by K4KJ was of interest because I had written some articles on this antenna for RADIO magazine back in the 1930s. There was one variation I always used because I just don't like to cut wire into little pieces. I'd put my two meters in a jumper about 2 meters (6 feet) long, and jump this across the section of antenna to be measured; it can be moved along without cutting the wire. For a time I tied string from the feedline to the antenna to support the weight of the meters and feedline, so the jumper wire would stay reasonably close to the antenna.

I tended to use a wire about 85 meters (280 feet) long, and had no problems on 10, 20, 40, and 80 meters. With modern rigs, however, there might have to be some arrangement for coupling the single-wire feeder to the exciter or linear amplifier, especially for band switching without complications.

> Bill Conklin, K6KA La Canada, California

^{*}Allan Lloyd, "Here's a better way to design a 90° phase-difference network," Electronic Design 15, July 22, 1971, page 78.

[†]Allan Lloyd, "90-degree phase-difference networks are simply designed with a program in Basic," Electronic Design 19, September 13, 1976.



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THE TRAGIC LATE SEPTEMBER air crash in San Diego found Amateurs on the scene almost immediately, providing much-needed communications through area repeaters. WR6AFC bore the brunt of the action, serving both the Red Cross and County Emergency Medical, with WD6LUD doing a superb job as net control. W6INI handled the lion's share of the traffic, while W6OTE operated from Red Cross Headquarters. The special 34-94 repeater that had been set up for the ARRL convention was also pressed into service, operating port-

able from a nearby ridge.

Several Dozen Amateurs participated in the operation, while discipline among the hundreds listening was excellent. Amateurs at or near the scene found themselves directing traffic, hauling spares and emergency equipment, and in general working shoulder-to-shoulder with the professionals at the scene. The Red Cross relieved its Amateur Radio assistants about 1:00 PM, almost four hours after the crash, but the County Emergency Medical teams contined to depend on Amateur communications — particularly WD6AHX, an AREC member with extensive emergency equipment — for another two hours.

CANADA'S NO-CODE AMATEUR LICENSE went into effect September 30, with the first exams ember 15. The "Amateur Digital Radio Operator's Certificate" gives the holder operating privileges on 2 meters and up and includes packet (computer-to-computer data bursts)

and pulse as well as more conventional modulation modes.

Holders Of The New no-code Canadian certificate may upgrade to the Canadian "Advanced" certificate simply by passing a 15 WPM code test, but must wait one year after initial licensing before becoming eligible to do so. Canadian Amateurs may now also retain credit for passing a CW test for up to one year after it's taken, even though the written test was failed. Amateur exams will now be given four times annually, in January, April, July, and October.

Packet-Radio Transmissions will be permitted on 220.1-220.5 MHz, shared with other modulation modes, while 221-223 and 433-434 MHz will be reserved exclusively for packet radio use. Packet would also be permitted on 24.00-24.01 GHz. Each packet transmission would require a "header" that would include the ASCII-coded call sign of the transmission would require a "header" that would include the ASCII-coded call sign of the transmission would require a "header" that would include the ASCII-coded call sign of the transmission will be permitted to the state of the transmission will be permitted to the state of the transmission will be permitted to the state of the transmission will be permitted to the state of the transmission will be permitted to the state of the transmission will be permitted to the state of the transmission will be permitted to the state of the transmission will be permitted to the state of the state mitting station, and no secret codes or ciphers would be permitted. Technical characteristics and modulation techniques for packet radio are to be determined by "Amateur

experimentation.

THE EARLY SEPTEMBER IARU Region 2 meeting in Panama City was one of the most successful and encouraging Amateur Radio conferences ever held in Latin America. The amount of preparation, both for next year's World Administrative Radio Conference and for the Panama City meeting itself, was a pleasant surprise, with very positive implications for the Amateur Service's future. Support for new Amateur HF bands by their governments was reported by a number of delegates, but it appears that not one country south of the United States favors expansion of HF broadcasting.

CINCINNATI WAS THE SITE of a late-scheduled hearing on the Communications Act rewrite, and, thanks to WB8FGE and a group of area Amateurs, the Amateur Service made a very good showing before the three attending congressmen. Following the suggestions that resulted from the report of the Chicago hearing, Peter submitted a number of letters and citations praising the Amateur Service for various public service contributions and made numerous suggestions (including definitions of the Amateur Service and various Amateur operations) in his testimony. He also pointed out that the rewrite as proposed was oriented toward profit-making radio services, and thus was quite inappropriate for Amateur Radio in several specific areas.

An On-The-Air Demonstration climaxed WB8FGE's testimony. He keyed up a local repeater with his hand-held and made an auto-patch call to the head of the local National Weather Service office, who told those attending the hearing how important Amateur Radio had been

AMATEURS PUTTING HR REPORT or other items of "general interest to Amateurs" on the air as a "QST" are generally well within the FCC's rules, but should be discreet with respect to some types of stories. True news stories pose no problem, but job offerings should probably not be aired at all; announcements of new products, forthcoming books, and the like should be treated strictly as news without price or "where to buy" details, to avoid any appearance of commercialism. The rules do not appear to be as restrictive for 2-way contacts, however, as they are for QSTs.

LARRY LEKASHMAN, W2IOP, passed away in late September after being ill for some months. Larry, ex-W9IOP and W8IOP, had been editor of CQ back in the 1950s and had held toplevel positions at Electro Voice, Olson, Lafayette, and Gladding. He was also well known on the air as a top DXer and contest operator.

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8-POLE CRYSTAL FILTER. 2.4 kHz bandwidth, 1.8 shape factor. SEPARATE MODE SWITCH. Permits using all filters in any mode. 2-SPEED BREAK-IN. Switch to "fast" or "slow" receiver muting to accomodate any band condition or mobile operating.

2-RANGE OFFSET TUNING. Switch-select the ±5 kHz range for off-frequency DX work or the ±0.5 kHz range for fine tuning.

OPTIMIZED RECEIVER SENSITIVITY. Ranges from 2 uV on 160 m to 0.3 uV on 10 m (10 dB S+N/N) to achieve ideal balance between dynamic range and sensitivity

GREATER DYNAMIC RANGE. Typically exceeds 90 dB to reduce possible overload from nearby stations. Also includes switchable 18 dB PIN diode attenuator for additional overload prevention.

WWV RECEPTION. On the 10 MHz band switch position

FRONT PANEL CONTROL OF LINEAR/ANTENNA BAND-SWITCHING. Auxiliary bandswitch terminals on back panel for simultaneous control of external relays or circuits with the OMNI bandswitch.

BUILT-IN PHONE PATCH JACKS. Provide interface to speaker and microphone audio signals for phone patch connection.

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BUILT-IN ZERO BEAT SWITCH. Permits placing your transmitted signal exactly on the listening frequencies of CW stations.

BUILT-IN SWR BRIDGE. The "S" meter electronically switches to read SWR every time you transmit to provide a continuous antenna

FRONT PANEL MICROPHONE AND PHONE JACKS.

ADJUSTABLE AUTOMATIC LEVEL CONTROL. For setting output power level from low power to full output, for retaining low distortion at desired drive power to linear amplifier.

SEPARATE RECEIVING ANTENNA CAPABILITY. Rear panel switch and jack connect receiving section to common antenna or separate receiving antenna. Also acts as receiving antenna by-pass when used with instant break-in linear amplifiers.

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- Receiver RESONATE control for peak sensitivity.
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- 6-Digit LED FREQUENCY READOUT for 100 Hz accuracy OFFSET TUNING LED indicates OT switch is "on"
- Combination "S" and SWR METER, switches automatically
- MAIN TUNING KNOB, big. easy-to-grip with integral spinner.
 AUTOMATIC LEVEL CONTROL LED indicates ALC-region operation.
- 10 DRIVE control for final stage 11 SQUELCH combination on/off switch and control 12 4-Position SELECTIVITY switch for SSB and CW. 13 4-Position MODE switch: automatic SSB Normal, Reverse, CW,

9 Combination ALC control and NOISE BLANKER on/off switch

- andd Lock (key down) 14 Combination push-pull POWER switch and AUDIO LEVEL control 15 Combination RF ATTENUATOR prioff switch and control.
- 16 VOX GAIN control

- 17 VOX DELAY control
- 18 VOX ANTI-TRIP control
- 19 11-Position BAND SWITCH 20 MICROPHONE jack, ht-2 input
- HEADPHONES jack RECEIVER OFF-SET TUNING SWITCH, 3-position, Max-Min-Off
- VOX-PTT SWITCH
- 24 QSK (full break-in) SWITCH, 2-position, Fast-Slow

mosfet power amplifier

for operation from 160 — 6 meters

The potential of the recently developed vertical mosfet^{1,2} to simplify and to improve rf power amplifiers has been shown in several recent papers.³⁻⁶ These circuits usually employ class-A amplification, however, and so do not achieve the efficiency possible with class-B or class-D operation.^{7,8} Where mosfets have been used in high-efficiency modes of operation, their power output has been relatively low.^{9,10}

The circuit described here uses two Siliconix VMP-4 *Mospower** fets in a push-pull configuration. The same circuit may be operated in either class B (for linear power amplification) or in class D (for highly efficient power amplification) by use of the proper output filter. The 160- to 6-meter frequency range and the 16-watt output make this circuit attractive for use in amateur equipment, particularly when multiband and multimode operation is desired.

The vertical mosfet has many advantages over its bipolar counterpart for both class-B and class-D rf power amplification. In particular, a complicated bias supply is unnecessary because of the high input impedance and the negative temperature coefficient of the fets; drive power is reduced and input matching is simplified by the high input impedance. Because of the negative temperature characteristic, secondary breakdown does not occur; hence the fets can withstand reactive loads in class-B operation.

In addition, the absence of storage time facilitates rapid switching in class-D operation. The absence of storage time eliminates the possibility of subharmonic oscillation in either class-D or saturated class-B operation. The ability of mosfets to pass current in both directions (i.e., they are bilateral) allows them to operate in class D with reactive loads without

*Mospower is the registered trademark of Siliconix Incorporated, Santa Clara, California.

This circuit was the runner-up in the 1977 EDN-Siliconix VMOS design contest, and is being published concurrently in EDN Magazine.

diode protection (since suitable diodes for high-power high-frequency operation are not available, this is quite significant). The saturation resistance of an fet (in contrast to the saturation voltage of a bipolar transistor) allows the efficiency in class-D operation to remain high even at low power levels. This is also significant, since ssb voice signals¹⁰ have relatively high peak-to-average power ratios.

The low gate-drain capacitance of the mosfet reduces feedback, making the amplifier easier to stabilize. The low gate-drain capacitance also reduces feedthrough, hence improving linearity (by supply voltage modulation) in class-D operation. And finally, the essentially constant gain of the vertical mosfet over the entire high-frequency band eliminates the need for gain-leveling circuitry.

theory

Class-A operation means that the fets are biased to remain in their active regions at all times.^{7,8} While this mode of operation promotes high gain in bipolar transistor power amplifiers, it limits the circuit power efficiency to a maximum of 50 per cent. Efficiency is even less at less than full output (**fig. 1**), and attainment of full output is prevented in real circuits by the nonzero saturation resistance of real fets (or by nonzero saturation voltage in bipolar amplifiers).

A broadband class-B amplifier (fig. 2A) uses a pair of fets (or bipolar transistors or vacuum tubes) as current sources, controlled by the driving signal. For maximum linearity, the fets are biased so they have only a very small drain current when no driving signal is applied; therefore, application of drive alternately causes one transistor to be active while the other is cutoff (fig. 2B).

The parallel-tuned output circuit bypasses any harmonic components in the current, allowing only the fundamental (carrier) frequency component to reach the load. The sinusoidal voltage generated there becomes the drain voltage waveform with the addition of the supply voltage. The power output is then

$$P_o = \frac{V^2_{eff}}{2R} \tag{1}$$

where R is the drain load resistance and $V_{eff} = V_{DD} \ R/(R+R_{on})$ accounts for the effects of fet on (saturation) resistance R_{on} .

At peak power output the efficiency (rf power out-

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put divided by dc power input) of a class B poweramplifier is

$$n \cong (\pi/4) R/(R + R_{on}) \le 78.5\%$$

At lower output levels, class B efficiency is significantly higher than class A efficiency.

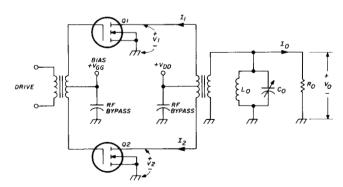
Saturation of the fets produces only a small increase in the output power and efficiency, because the sinusoidal drain voltage is maintained by the parallel-tuned circuit. During saturation, the fets are roughly equivalent to constant resistances, and the drain current is determined by application of the drain voltage to this resistance.

A class-D amplifier also uses a pair of fets, but they are controlled by the driving signal to act as switches. The voltage-switching configuration (fig. 3A) has square-wave drain voltage (fig. 3B) produced by the alternate switching of the two fets. The series-tuned output circuit passes current only at the fundamental (switching) frequency, and the sinusoidal output voltage is equal to the fundamental frequency component of the square-wave on the transformer secondary winding. Alternate half-cycles of the transformed output current flow through each fet. (If the load is reactive, the output current is outof-phase with the switching, and at times must flow in the reverse direction through each fet.) The amplitude of the output is a function of the supply voltage, and is not affected by the amplitude of the driving signal if the drive signal is sufficient to cause switching. The power output of a class-D power amplifier is

$$P_o = \frac{8}{\pi^2} \cdot \frac{V^2_{eff}}{R} \tag{2}$$

A class-D amplifier can ideally achieve an efficiency of 100 per cent, but in practice achieves $\eta \cong R/(R+R_{on})$ at most output levels. A gate bias voltage is not required, but is easily implemented with

fig. 2. Operation of vertical mosfets (vmos fets) in class B. The fets act as current sources. Waveforms of the class-B amplifier are shown at right.



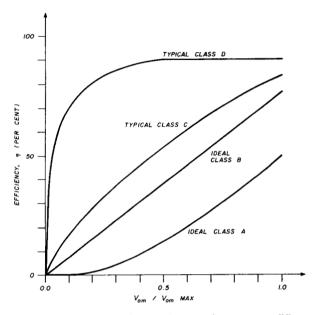
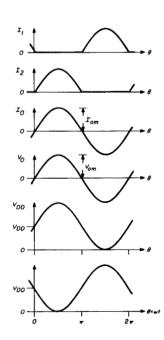


fig. 1. Efficiency of different classes of power amplifiers. Maximum theoretical efficiency of a class-A power amplifier is 50 per cent, while class B efficiency is 78.5 per cent. The efficiency of a class-D (switching) amplifier is much higher, particularly at lower output levels.

fets and can reduce the required amplitude of the driving signal.

Class-B operation of rf power amplifiers is typically used in the high-frequency bands for linear amplification of single-sideband suppressed-carrier signals. Class-B power amplifiers may be driven just into saturation to ensure maximum output and efficiency for CW and FSK transmissions. In real applications, a set of lowpass filters (with shunt capacitor inputs) is



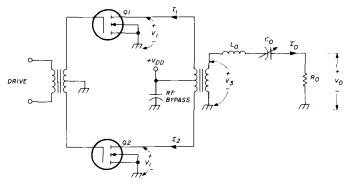


fig. 3. In class-D operation the fets operate as switches. Waveforms of the class-D amplifier are shown at right.

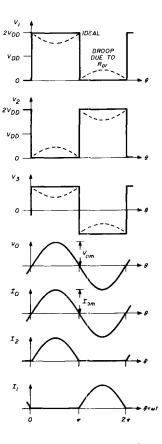
used instead of a simple parallel-tuned circuit. The cutoff frequencies of the filters are chosen to be slightly less than one octave from each other to provide adequate harmonic rejection for any frequency of operation.

Class-D amplifiers for high frequency use are just now becoming practical. Circuits such as this would be used directly for CW, FSK, and fm transmissions. Linear amplitude modulation is readily accomplished by variation of drain supply voltage. Single-sideband signals can be amplified through the envelope elimination and restoration technique, 7,8,10 which uses amplitude modulation to restore the envelope of the ssb signal and a limiter to ensure adequate drive for the power amplifier. A set of lowpass filters might similarly be used to span the range of operating frequencies; however, these would have series-inductor inputs, rather than the shunt capacitor inputs of the filters used with class-B amplifiers.

design

Since input requirements are based on the output current, design begins with the output portion of the circuit.

Output design. Inspection of the VMP-4 Mospower data sheet shows a maximum drain current of 1.6 amps and a maximum drain voltage of 60 volts; the latter allows a +30 Vac variation around a supply voltage of +30 volts. Were there no saturation voltage, outputs of 24 watts (class B) and 30.4 watts (class D) could be obtained. For these power outputs, load resistances of 30/1.6 = 18.75 ohms (class B) and $(4/\pi)(30/1.6) = 23.87$ ohms would be reguired. The load resistance required to obtain the maximum output power when the saturation (on) resistance is non-zero is determined by subtracting the saturation resistance from the load resistance determined above for zero saturation resistance. Thus, for a typical value of $R_{on} = 2.6 \text{ ohms}$, the maximum power loadlines are 18.75 - 2.6 = 16.15



ohms and 23.87 - 2.6 = 21.27 ohms for class-B and class-D operation, respectively.

Obtaining optimum broadband performance requires the use of transmission-line transformers; 8,11 this limits the choice of impedance transformations. A convenient transformation is 4:1; with a 50-ohm load, the load line is $R=12.5\ ohms$. This transformation is accomplished by a pair of transformers as shown in **fig. 4**. The hybrid (T2) requires a 25-ohm transmission line; it is built by winding two turns of transmission line through two parallel stacks of Ceramic Magnetics CN-20 ferrite* (each stack is 2.86 cm long). The 25-ohm transmission line may conveniently be made by connecting two 50-ohm miniature coaxial cables in parallel.

The balun (T3) requires a 50-ohm characteristic impedance line, and is fabricated by winding two turns of miniature coax through two parallel stacks of CN-20 material (also 2.86 cm long). The combination of T2 and T3 provides a mostly resistive input impedance over the entire 1.6 to 54 MHz band when connected to a 50-ohm load.

With a load line of 12.5 ohms, the output power is limited by the maximum drain current, thus

^{*}If CN-20 material is not available, other rf ferrite material can be used. The reader can use the design procedures in references 8 or 11 or can borrow a design from another published PA circuit.

$$P_{omax} = \frac{I_{Dmax}^2 R}{2} = 16 \text{ watts}$$
 (3)

(An output of 25 watts would be possible with the VMP-1, which has a 2 amps rating.)

The 1.6 amp peak drain current requires a dc supply current of $1.6(2/\pi) = 1.02$ amps. In class-B operation, the drain voltage swing is simply $1.6 \cdot 12.5 = 20 \text{ volts}$ at peak output. To avoid saturation, the drain voltage must be greater than $1.6 \cdot 2.6 = 4.2 \text{ volts}$; therefore, the supply voltage must be 24.2 volts to allow maximum output power (higher supply voltages allow maximum output power, but reduce the efficiency).

For class-D operation, the 4.2 volt peak drop across the saturation resistance acts in opposition to the fundamental frequency component of the ideal square-wave, requiring a supply voltage of 24.2 $(\pi/4) = 19.0 \text{ volts}$ for maximum output. With $R_{on} = 2.6 \text{ ohms}$, you can expect efficiency of 82.8 per cent for class D, and 65 per cent for class B.

The values of the bypass and blocking capacitors and chokes are not especially critical. However, care should be taken to make sure there are no self-resonances within the band of operation. For testing purposes, simple tuned circuits with a loaded Q of 3 (parallel-tuned) or 5 (series-tuned) were used.

Input design. The input circuitry of a power amplifier using fets can differ considerably from the input circuitry of a bipolar amplifier. A relatively high voltage across a relatively high capacitive reactance is required. Both gate capacitances remain in the circuit at all times, in contrast to the reverse-biased base being effectively out of the circuit. Inspection of the VMP-1 data sheet suggests that a bias voltage of about 2 volts and a sinusoidal voltage of about 6 volts peak will be required to produce the 1.6-amp peak

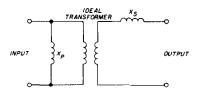


fig. 5. Equivalent circuit of the wirewound transformer used at the input of the amplifier (see text).

drain current. The VMP-1 data sheet also shows a gate-to-source capacitance of about 50 pF.

A 4:1 voltage reduction produces a more convenient output for the driver. Since the input impedance of a Mospower fet varies from about 2 kilohms at 1.6 MHz to 59 ohms at 54 MHz, true broadband matching is difficult. Since the drive power is relatively small in comparison to the output power, however, matching of the input to the power amplifier is not really necessary in most applications. This philosophy makes possible the use of a simple "wirewound" transformer. Fig. 5 depicts the equivalent circuit of such a transformer; low-frequency performance is degraded by the shunt (magnetizing) inductance X_{p} , while high-frequency performance is degraded by the series inductance X_s . If the driver acts as a current source, only the shunt inductance reduces the current reaching the gates. If the driver acts as a voltage source, only the series reactance reduces the voltage applied to the gates.

The selected design for input transformer T1 is therefore a single-turn primary and a 4+4 turn secondary wound on a Ferroxcube 3B7 core (1 cm diameter). At midband, a driving voltage of 2.5 volts (peak) with a 70 mA (peak current) should be sufficient. The required driving current will be higher at the low end of the band, and both the driving current and the driving voltage increase at the high end of

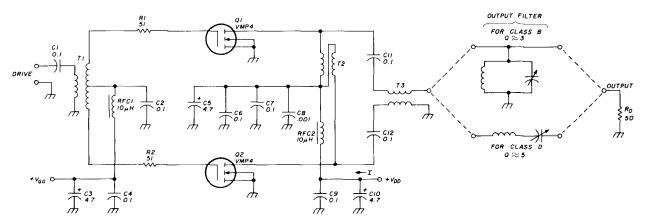


fig. 4. Schematic of a vmos fet power amplifier for 160 through 6 meters. The design of transformers T1, T2, and T3 is discussed in the text. Capacitors C5, C6, C7, and C8 provide an rf ground; combination may vary with different capacitors. RFC1 and RFC2 consist of 40 turns of no. 26 (0.4mm) wire on a Permacore 57-1677 form.

the band. Better overall performance might be achieved by using a core made from material with a lower permeability (3D3, for example). The 51 ohm resistors R1 and R2 dampen the tuned circuit formed by fet capacitance and the inductance of transformer T1, thus preventing high-frequency oscillations.

Construction is straightforward, not critical, and guite similar to that for other high-frequency power amplifiers. Fig. 6 shows the physical layout of the amplifier. The fets are in the middle, with input-related circuitry on their left and output-related circuitry on their right. Transformer T3 is on top of transformer T2, and both can be supported by pieces of PC board. The circuit was assembled on a piece of double-sided PC board to ensure a good ground plane; holes were cut in the board to allow the fets to be mounted directly on a small heatsink.

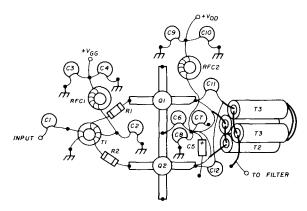


fig. 6. Parts layout of the mosfet power amplifier. Construction is straightforward, not critical, and similar to that used for other high-frequency amplifiers.

performance

Performance of this PA is measured in terms of its linearity and efficiency in both class B and class D operation, as well as by its maximum output and efficiency at various frequencies.

Class B linearity. Measurements of the linearity (output voltage) and efficiency as functions of the drive amplitude were made at 3.5 MHz and with $V_{DD} = 24 \text{ volts}$. A gate bias voltage of 1.65 volts was used; this produced a total quiescent drain current of about 70 mA. Since a spectrum analyzer was not available, this value of V_{GG} was selected by increasing V_{GG} until cross-over distortion vanished from the combined current waveform at the output from T3.

Variation of the output voltage and efficiency with drive (peak voltages) is shown in fig. 7. It is evident in both curves that saturation has an effect at a lower output than expected. Efficiency increases roughly in proportion to output amplitude, reaching the expect-

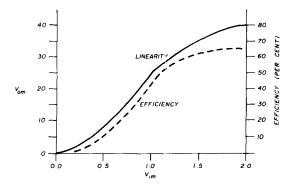


fig. 7. Class B performance of the mosfet power amplifier. Efficiency increases approximately with the output amplitude, reaching the expected 65 per cent efficiency at peak output.

ed 65 per cent at peak output. A simple numerical analysis shows that this curve is roughly equivalent to a third-order IMD ratio of -20 dB with the peak output power at 16 watts. Decreasing the peak output power or increasing V_{DD} will keep the fets out of saturation and thus decrease the IMD. However, it is very possible that a different choice of V_{GG} will improve the IMD; this can easily be determined with the aid of a spectrum analyzer. Should this not produce a level of -30 dB or better, feedback might be added to the circuit.

Class D modulation linearity. Measurements of performance in class-D operation at various amplitudes were also made at 3.5 MHz. A driving voltage of 2.5 volts peak was sufficient to ensure saturation. Gate bias was reduced to 1.2 volts, since drive linearity was not required.

Variations of the output voltage and efficiency as functions of the supply voltage are shown in fig. 8. The very linear relationship of the output voltage to the supply voltage indicates excellent amplitude

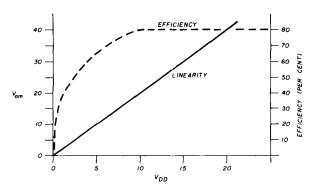


fig. 8. Class D performance of the high-frequency mosfet power amplifier. The very linear relationship of output voltage to supply voltage indicates excellent amplitude modulation characteristics. Efficiency remains close to 80 per cent except for low output levels.

modulation characteristics. The feedthrough (V_{om} with $V_{DD} = 0$) measured only 34 mV, which is - 61.6 dB relative to peak output! Efficiency remains close to 80 per cent except for low output levels, where it decreases but is still several times greater than that for class-B operation.

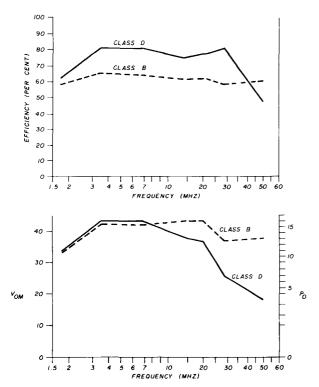


fig. 9. Performance of the mosfet power amplifier vs. frequency.

This amplifier can be tuned at full output power without fear of damage to the fets. This contrasts with the use of bipolar transistors, 10 with which tuning at other than very low power can destroy the transistors. Because of the exceptional stability of the Mospower fets, there is no oscillation when the drive is reduced to less than that required to maintain saturation; the output simply decreases gradually as the fets become current sources instead of switches. This may make possible a simplification of the class BD rf amplifier. 12

frequency characteristics

Measurement of circuit performance in class-B operation was made with the supply voltage set at 24 volts; drive was then increased until maximum output was obtained. The resulting variations of efficiency and maximum output with frequency are shown

by dashed lines in fig. 9. In the 80 to 15 meter bands, the output and efficiency are nearly constant at about 16 watts and 64 per cent, respectively. At both the high and low ends of the 160 to 6 meter range, both output and efficiency decrease as the performance of the output transformer degrades.

The performance of the circuit in class-D operation was measured with the supply voltage set at 20 volts and the drive increased until no further increases in output were observed. The resulting efficiency and maximum output curves are shown in fig. 9 as solid lines. An efficiency of about 80 per cent is obtained in the 80 and 40 meter bands with an output of about 16 watts. Slight decreases in the output and efficiency can be seen in the 20- and 15-meter bands. More significant decreases in both output and efficiency occur near the high and low band edges.* It is not surprising that these are more pronounced than in class-B operation, since class D is a more exacting mode of operation.

This circuit has demonstrated the use of Mospower fets in both class-B and class-D operation in the high frequency range. Performance in class-B operation was generally similar to that obtained with comparable bipolar transistors; performance in class-D service was generally superior to that which can be obtained with bipolar transistors. In either type of operation, the VMOS fet offers many advantages stated previously that do not show up in the measurements made above, but do show up in the operation of real circuits. Similar design techniques can be used with other VMOS fets to achieve other power outputs.

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ham radio

^{*}Because of the limitations of the author's test equipment, the 10-meter and 6-meter measurements should be regarded as approximate.

basics of the digital VFO a tunable synthesizer

Basics of the digital synthesizer, from simple switch-controlled dividers to a full optical-encoder controlled system

It has been many years since something technically innovative has come to ham radio. Two meters has allowed development of repeaters, scanning, and Touch-Tone* phone patching, but these developments are generally extensions of already wellknown and previously applied principles. The digital synthesizer is somewhat new to amateur radio, but has been in use in commercial and military equipment for many years. The digital synthesizer was a natural for two-meter operation because of the cur-

*Touch-Tone is a registered trademark of the American Telephone and Telegraph Company.

rent practice of operating on fixed channels. By and large, amateurs have contributed very little to the synthesizer art.

On the high-frequency bands, due to nonchannel operation, the knob- or push-button-operated synthesizer has found little favor. Those of you who have tried to tune the hf ham bands using synthesized, military receiving equipment will readily understand why. Yet the advantages of the synthesizer over the VFO are tremendous. Apart from its inherent frequency stability and accurate readout, the synthesizer allows the transceiver designer to convert to an i-f frequency higher than the signal frequency, eliminating the need for multiple conversion and the attendant barrage of spurious and birdie signals that usually accompany multiple conversion techniques.

Additionally, each mixer placed at the front end of a receiver increases the potential for cross modulation, intermodulation, and image products.

The principal advantages of up-conversion design have been well documented, particularly in recent articles by Ulrich L. Rohde, DJ2LR,1 and Wayne Ryder, W6URH,² So, no more will be said here except to point out that a high-frequency i-f amplifier requires a high-frequency local oscillator signal. Unlike the VFO, the synthesizer readily operates at the higher frequencies.

Recently, a new transceiver, the Astro 200, appeared on the ham market. To the best of my

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knowledge, the Astro 200 was an amateur first. Its design has probably set the pace for all new designs in ham radio, using a tunable synthesizer. To tune down in frequency a momentary switch is depressed; to tune up in frequency the switch is lifted. A digital readout keeps track of the frequency. The company's literature claims that this method of tuning will eventually be used in all ham transceivers and, with some reservations, I agree. However, I do not agree about the switch method of performing the tuning. In my opinion, the method lacks feel and does not control the speed with which the signal is tuned. But this is a matter of application, not principle. There are various variable-speed methods that overcome this objection. The point is, it is only a matter of time before the continuously tunable synthesizer does to the VFO what transistors have done to tubes.

Surprisingly, the manufacturers of the Astro 200 stayed with the conventional low-frequency i-f, reducing by perhaps 75 per cent the effectiveness of the system. To have "tamed" the multitude of harmonics and birdies that this low-frequency system must have generated is a task I am pleased was not allotted to me!

There has been little information in the amateur literature, or any literature for that matter, concerning tunable synthesizers, or, as I have chosen to call them here, digitally tuned VFOs (DTVFOs). Anyone experimenting with DTVFOs is virtually on his own. It is hoped that the information presented here, which has come mostly from breadboards and experiments, will be of some assistance. I have tried to present the information in a progressive way, much as I learned it myself. However, I should point out that the devices are not necessarily the latest or the best available. Most of the ICs are readily available junk box surplus.

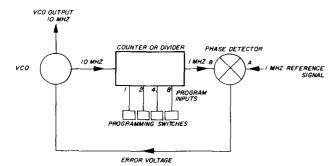


fig. 1. Block diagram of a basic digital synthesizer. The phase detector is used to determine the difference between the reference frequency and that supplied by the divider. The error voltage represents the difference and is used to change the frequency of the VCO until there is zero difference. The programming switches can alter the divider, and hence will change the output frequency of the VCO,

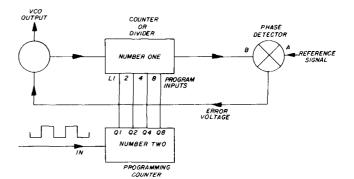


fig. 2. In this example, the programming switches have been replaced with another counter. Since the outputs of the second counter are in BCD, they are used to directly drive the load inputs of the first counter.

Amateurs with a technical background will have little trouble duplicating and improving the circuitry.

synthesizers

A DTVFO is actually a modification of a regular synthesizer which uses a number of panel-mounted BCD switches or pushbuttons to program counters to control the desired frequency. The DTVFO, instead of using switches, uses the Q outputs from a second parallel counter to do the programming. This is more clearly illustrated in the block diagrams of **figs. 1** and **2. Fig. 1** is a conventional, simplified digital counter using a phase-locked loop system and panel-mounted switches to select the frequency.

In this instance, let the reference signal applied to the phase detector be 1 MHz and the VCO output be 10 MHz. If the counter or divider has been set by the knobs to divide by 10, the second signal applied to the phase detector will also be 1 MHz. There will be no error voltage from the phase detector. Now, let the VCO be 11 MHz; the signal applied to the phase detector will now be 1.1 MHz. The phase detector will develop an error voltage and retune the VCO to 10 MHz so that the input to the phase detector is again 1 MHz. When no error voltage is developed the system is in lock.

If the divide ratio is set to 9, the VCO will retune to 9 MHz so that once again the phase detector input is 1 MHz. Thus, by changing the divide ratio, a number of VCO output frequencies, each locked to the crystal-derived reference frequency, may be obtained. Note that the counter cannot divide by fractions. Since the reference frequency in the example just discussed is 1 MHz, the VCO output can only be in 1-MHz increments. If smaller increments are required, the reference frequency must be lowered and the number of counters and synthesizer frequency-selector knobs increased. Obviously, it requires only one knob to tune a frequency synthesizer from 1 MHz to 9 MHz in 1-MHz steps, but it requires four

knobs to cover the same frequency range in 1-kHz steps.

Fig. 3 shows a truth table for a 74192 programmable decade counter commonly used in frequency synthesizers. The program input pins (often called load inputs, program inputs, or jam inputs) are normally held low. By connecting specific inputs to +5 volts, a predetermined counter divide-ratio may be obtained. Highs (+5 volts) are shown as 1 on the truth table. To divide by 1, for example, input L1 (pin 15 on the 74192) is taken high. To divide by 2, input L2 (pin 1) is made high. And to divide by 3, both L1 and L2 are made high. These functions are normally carried out by the programming switches. It should be readily apparent that if a method of progressively raising the appropriate inputs can be devised, the VCO will sweep across a band of frequencies. If a smaller reference frequency is used, the VCO will sweep in smaller steps. When the steps are sufficiently small - less than 100 Hz - it will seem as if the VCO has tuned across a signal in a normal VFO manner.

To accomplish the tuning or sweeping action yet stop on command, a second counter may be substituted for the BCD switches. This is shown in fig. 2 and, for the purpose of clarity, this counter is referred to as No. 2 counter. If we feed a square wave, one pulse at a time, into the input of counter No. 2, assuming the counter had been set to zero, the first pulse will cause Q1 output to go high. Since Q1 is connected to L1 of the No. 1 counter, the synthesizer will divide by 1. When the second pulse is entered into the counter, L1 will go to ground and L2

DIVIDE	BY	1	<u>L</u> 2	4	<u>L</u>
	/	/			
,	2		1		
;	3	/	/		
	4			/	
:	5	/		/	
	6		/	/	
	7	/	1	1	
	8				1
	9	1			1

fig. 3. Truth table for the programming inputs of a 74192 programmable divider.

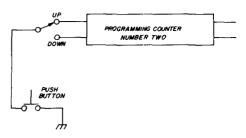


fig. 4. The single-pole, double-throw switch allows either an up or down count each time the button is pressed. Normally though, a low-frequency pulse train would be fed into the switches.

will go high. The counter will divide by 2. Similarly, a third pulse will cause both L1 and L2 to go high for a count of 3. And so on. If only one set of counters is used, the division will be from 1 to 9. If two sets of series-connected counters are used, the division will be 1 to 99; three counters, 1 to 999; and so forth.

To tune or sweep a synthesizer, it is merely necessary to feed a number of pulses into the No. 2 counter. The pulses may be created by sequential action of a pushbutton-type switch. But, as this is a slow process, a square-wave oscillator may be turned on when the button is pressed and the output entered into the counter. This is the same method used to tune the Astro 200 transceiver.

One problem with feeding the oscillator into the counter is that it becomes necessary to tune across the whole band if the counter is at one end and the desired frequency at the other. This can be quite a chore on 10 meters, especially when the square-wave oscillator has a fixed speed. Fortunately, this problem may be overcome by presetting the load inputs. A *Touch-Tone* pad will allow the operator to preset the frequency, with the fine tuning accomplished by feeding the square-wave into the clock input.

With the counter connected as shown in fig. 2, it will sweep only in one direction. In the case of the single-decade unit, the counter will count to 9, reset to 0, and start over again. To move the frequency in either direction a special kind of counter is required. This is known as an up/down counter. The 74192 has this facility. The counter has two input terminals. one for counting up and one for counting down. Fig. 4 shows how a double-pole, single-throw switch may be connected to route the tuning pulses to either counter. In practice, pulse conditioners, also known as debouncers, would have to follow the switches. (No matter how quickly and firmly the buttons or switches are activated, there will be bounce.) For those who wish to experiment with pushbuttons, a simple debouncing circuit is given in fig. 5.3

The pushbutton makes it relatively easy to observe the sequence of events, but you must realize that in an actual system with a 100-Hz reference, the VCO

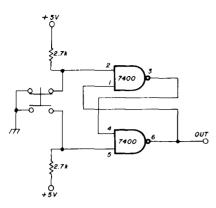


fig. 5. Example of a simple switch debouncing circuit.

will move only 100 Hz each time the button is pressed. Be prepared for an all-night session if you wish to tune across 10 meters!

As previously mentioned, a low-frequency, square-wave oscillator may be used to speed up the tuning; it may be switched between input terminals, or two separate oscillators can be used. Of course, only one oscillator may be activated at one time. It is important to note that when feeding a square wave into one input terminal, the other terminal must remain high.

Several switching methods for the square-wave oscillators have been used. The first used two separate pushbuttons, one for up and one for down. Better yet was a potentiometer with a center off position. When turned counterclockwise, the oscillator slowly started up, increasing frequency as the knob was turned. When the potentiometer was turned clockwise, past center, the other oscillator started.

frequency readouts

A major disadvantage of the DTVFO is that there is no simple, conventional, dial-type readout that can be used. Fortunately, the DTVFO also contains the essentials of a frequency counter that can be used instead of a conventional dial. Referring to fig. 2, the Q outputs of the No. 2 programming counter are exactly like those obtained from a frequency counter. Fig. 6 illustrates the connections necessary to convert the BCD information to a digital readout. Only a few extra components are required. The system is simplified by the fact that no latching is required. Since the counter does not update, no flickering occurs and the figures change only when the VCO is tuned.

With the frequency counter measuring the local oscillator, an offset is required so that the display indicates the actual transmitted frequency. If the signal frequency is 14000.0 kHz and the VCO is 5000.0 kHz, a 9000.0 offset is needed to make the readouts indicate 14000.0 kHz. In an amateur-band transceiver, the MHz readout is often hardwired and activated by the bandswitch because the bands only cover a small part of the spectrum. In a complicated system,

such as a general-coverage receiver, a PROM may well provide the programming link between the bandswitch and the readout.

A digital VFO is subject to all the pitfalls and problems normally encountered in a regular synthesizer. Of prime interest is error-voltage filtering and lock-up time, plus a couple of other hazards applicable to the tunable system. A tunable system requires a low-frequency reference, 100 Hz or less, preferably less. Even a 100-Hz system produces observable steps, especially noticeable on CW. Unfortunately, a lowreference signal makes it very difficult to adequately filter the reference spikes from the error-voltage line without unduly affecting the settling and lock-up time. The reference signal will show up as sidebands, removed from the carrier by the amount of the reference frequency. When a low-frequency reference is used, lock-up time takes considerably longer, and if the VCO frequency is high, many cycles of operation will take place before correction can occur. The VCO may then exhibit jitter or burble.

Sideband content is also highly dependent upon the type of phase detector used. Rohde has thoroughly discussed the advantages and the disadvantages of different types of phase detectors. There is little I can add except to reiterate that the sample-and-hold-type detectors and the RCA CD4046 are preferred over most of their counterparts.

In many synthesizers, when the frequency is being changed the oscillator output is muted. With the oscillator muted, it will be impossible to recognize that a station is tuned in until the tuning stops. How-

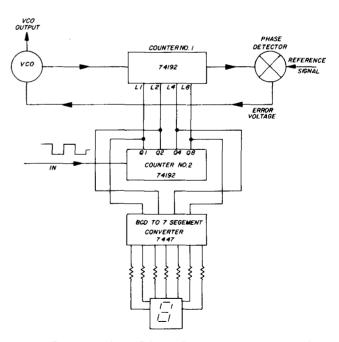


fig. 6. By connecting a BCD-to-7-segment converter to the outputs of the first counter, the frequency of the VCO can be directly read. This, in most cases, will not be the frequency to which the receiver is tuned.

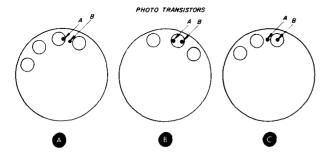


fig. 7. Diagram of the basic system for using a rotating disc to control the frequency of the VCO. Depending upon the direction of rotation, the voltages derived from the photo transistors will cause the counter to count either up or down, hence, changing the frequency of the VCO.

ever, if presetting is employed (a large VCO frequency change), muting may be necessary.

In a 100-Hz reference system, you will notice that the tuning action imparts a small fm effect to the signal during the tuning process. This is due to the fact the lock-up process is trying to follow the tuning, and unless lock-up is extremely fast, it must have a period of time in which there is an out-of-lock condition.

A problem encountered in tunable synthesizer systems that is not seen in the regular synthesizer occurs whenever the VCO approaches the end of the band. Assume a DTVFO is designed to cover the frequency range of 5 to 6 MHz, with the counter hardwired to 5 MHz. The counter counts up to 5997, 5998, 5999, and suddenly resets back to 5000 kHz. This can be somewhat upsetting to the operator trying to tune in a signal near the band edge. Fortunately, the problem is readily overcome by using a few inexpensive gates. For every frequency, the outputs of the No. 2 counter will have a unique combination of highs and lows. For example, when the counter reaches 9, the Q1 and Q4 outputs will be high. Therefore, if the two highs are sampled by a two-input NAND gate, its output may be used to prevent any further pulses from reaching the counter.

optical encoder

After having built and operated a number of DTVFO systems, I soon realized that something better than spring-loaded switches or pushbuttons was required if the tuning is to have the right feel. Using the button- or switch-activated square-wave oscillator, you either run right over the station or you stop short of it. Then it becomes necessary to "milk" the switch in the manner of an aircraft flap system, to creep up on the station. Then, in the middle of the milking process, the other person stops speaking! The potentiometer-controlled oscillator was better, but still lacked feel.

The next logical step called for the incorporation of the rotating disc tuning system.* In this system, a series of holes was punched around the perimeter of a 8.9-cm (3-1/2-inch) diameter disc. The disc was mounted on a flywheel, which in turn was connected to a 2.8:1 Jackson Brothers drive. The entire mechanism was then mounted on the receiver front panel and turned by an ordinary, large-diameter tuning knob. The flywheel not only gives the tuning a nice heavy feel, but also allows the knob to be quickly spun.

Using the same principles as the commercial versions (light sources and photo transistors), this tuning system creates a series of pulses as the knob is rotated. The problem was then one of sensing direction of travel, or tuning the DTVFO up or down in frequency as the knob is turned one direction or the other. The answer turned out to be quite simple, and is shown in fig. 7. Initial conditions start with photo transistor A receiving light and delivering an output (see fig. 7A). As the disc is turned clockwise, fig. 7B, both transistors receive light and deliver output voltages. In fig. 7C, transistor A has lost its light, accompanied by the resultant decrease in output voltage. Thus if the disc is turned clockwise, transistor A always receives and loses its light ahead of B. However, if the disc is turned in the opposite direc-

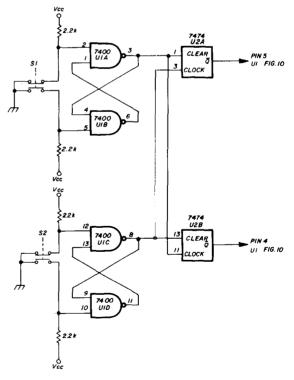


fig. 8. Pushbuttons have been used to simulate the pulses received from the photo transistors in the rotating disc, or optical encoder, system. The NAND gates are used to debounce the switches. The two 7474s determine in which direction the disc is rotating, thereby feeding the pulses to the appropriate input of the up/down counter.

^{*}Rohde & Schwartz, Munich, applied for a patent on the shaft-encoder, controlled-frequency synthesizer in May of 1961. Patent 977,780 was issued on July 13, 1962.

tion, B will receive and lose its light ahead of A. Therefore, the voltage pulses from the photo transistors, in conjunction with the appropriate logic, can be used to control an UP/DOWN counter. By detecting which pulse occurs first, A or B, the second pulse can clock the counter, either up or down.

To prove the theory, I constructed the actual circuit, with pushbuttons used to simulate the pulses from the photo transistors (see fig. 8). As previously mentioned, it is necessary to use debouncing circuits to ensure that only one pulse is generated when the button is pressed; hence, the 7400 NAND gates with normally low outputs. U2A drives the count-up input of the 74192 counter, with U2B driving the count-

the holes should be at least twice the diameter of one of the photo transistors. The transistors should receive an equal amount of light when the hole is exactly above the two. There must be no stray light, which causes the photo transistors to bias up into an intermediate state. When the light is shaded, the collector voltage should be close to +5 volts. When the photo transistor receives full light, its collector voltage should drop to less than 1 volt.

Just as the pushbuttons were followed by debouncing circuits, so must the photo transistors be followed by similar circuits. In this instance, the debouncing circuits are schmitt triggers fashioned from CMOS gates. CMOS was used to prevent

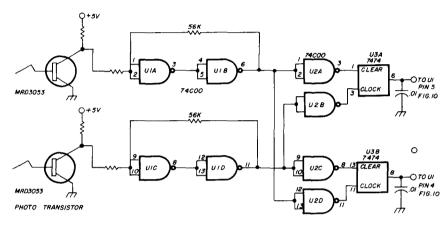


fig. 9. Final circuit for the homemade shaft encoder. Each time the light path is interrupted, a pulse is generated. This pulse causes the counter to increment or decrement by one count, depending upon the direction of disc rotation.

down terminal. Using the switches, I was able to simulate the sequence of events as the disc is turned.

- 1. S1 is pressed and held down. U2A is preset by the high level on pin 1.
- 2. S2 is pressed and held down. U2A is clocked, with pin 3 going high and pin 6 low.
- **3.** S1 is released. Pin 1 of U2A goes low. Pin 6 of U2A goes high causing the 74192 counter to advance by one count. (The counter operates only on positive-going pulses.)
 - S2 is released. Pin 3 of U2A goes low.

If the buttons are pressed in the opposite order (S2 pressed and held down, S1 pressed and held down, S2 released and then S1 released), U2B will be preset and clocked. The output pulse applied to the 74192 will cause the counter to decrement one count.

After the pushbutton system was operating correctly, the circuit was modified to that shown in **fig. 9.** A small light bulb was positioned so as to shine through only one disc hole into the two photo transistors. The hole diameter and the spacing between

loading of the photo transistors. Further gates were necessary to prevent the 7474 from loading the schmitt triggers and causing interaction between the inputs. Before the gates were added, the 7474 flipflops were delivering a stepped-output waveform which was causing the counter to misfire. Since the 74192 counters are edge triggered, a stepped waveform edge can cause havoc. If the CMOS version of the 7474 is used, the buffering may be unnecessary.

practical DTVFO

A complete, tunable synthesizer with 100-Hz resolution is shown in figs. 10 and 11; although I earlier mentioned that a DTVFO readily allowed up-conversion operation, the unit to be described operates in the low-frequency range, 5 to 6 MHz. This frequency was chosen so that its output could be used to operate an ALDA 103 transceiver. The frequency and output level are both compatible with this particular transceiver. Fig. 10 contains the two counter chains and readouts, with fig. 11 showing the reference oscillator, dividers, phase detector, VCO, and buffers.

Referring to fig. 10, the synthesizer divider chain

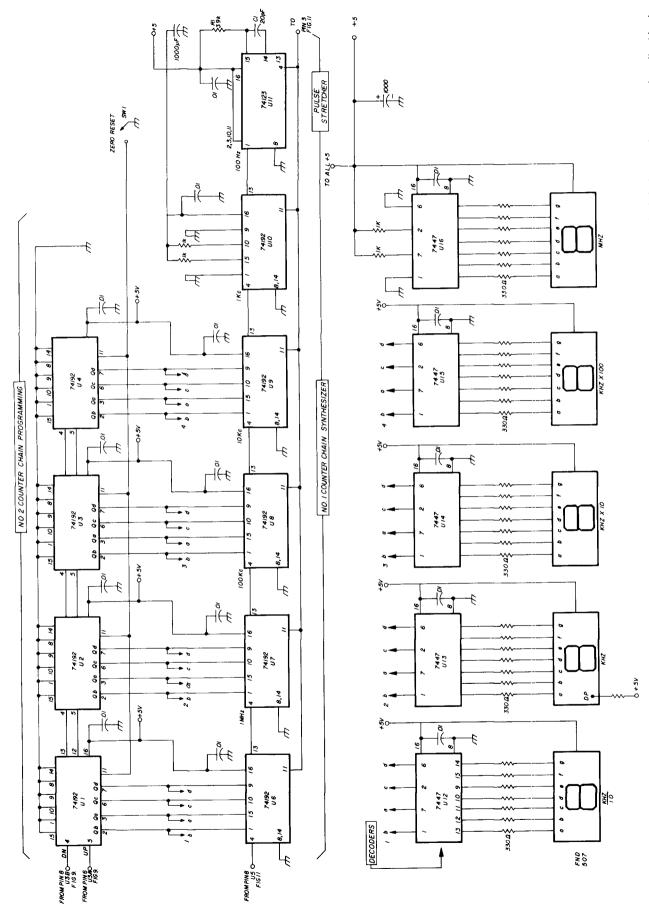


fig. 10. Schematic diagram of the counter chains used in the 5-MHz digitally tuned VFO. The up/down inputs for U1 come from the optical shaft encoder system described in the text. In addition to driving the load inputs of the first counter chain, the Q outputs also drive the BCD-to-7-segment converters. The load inputs for U9 have been hardwired since this counter always divides by five.

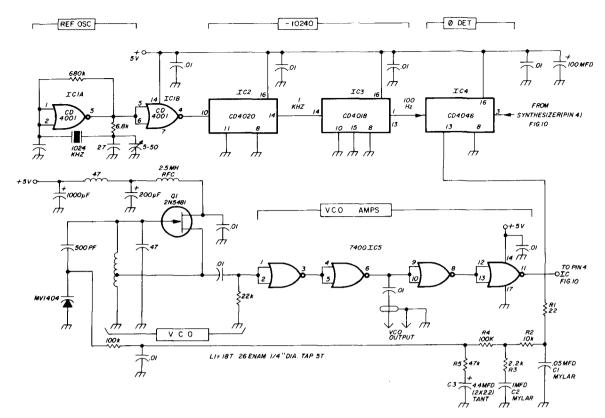


fig. 11. Diagram of the VCO and reference oscillator portion of the digitally tuned VFO. Using standard CMOS ICs, the oscillator is divided down to 100 Hz prior to being applied to the phase detector. Transistor Q1 is a Hartley oscillator which covers the 5 to 6 MHz range under control of the phase detector. L1 is 18 turns of no. 26 AWG (0.4 mm) wire wound on a 6.5-mm (1/4-inch) diameter form. The tap is located five turns from the bottom of the coil.

consists of U6 through U10. Pulse stretcher U11, in addition to supplying the reset pulse for the counters, also drives the phase comparator in **fig. 11**. U1 through U4 constitute the programming chain, with pulses from the optical encoder used to provide the count-up and count-down inputs. The Q outputs from the four respective counters are used to program the load inputs of the synthesizer divider chain and also to drive the BCD-to-7-segment converter.

U10 has been hardwired to divide by five so that the synthesizer output covers from 5 to 5.9999 MHz. Provided the VCO-tuned circuit is adjusted accordingly, U10 may be hardwired to cover other 1-MHz increments.

U16 may be dispensed with and the appropriate segment resistors grounded to obtain the digit 5, eliminating the 7447 BCD-to-7 segment decoder.

When first turned on, the counter outputs cause odd readout displays; momentarily grounding the load line (S1) will set all counters and displays to zero. Of course, this function could be made automatic or gated to the bandswitch so that each band is started at the low-frequency end.

In fig. 11, U1A is connected as a 1024-kHz crystal oscillator, followed by buffer U1B. U2 and U3 subsequently divide the reference oscillator output down

to 100 Hz prior to application to pin 14 of the phase detector.

The resultant error voltage from the phase detector is filtered by the following RC network. While this filter does not represent the ultimate in filtering, it is satisfactory for normal listening. A Singer spectrum analyzer shows the sidebands well down into the noise. However, when the VCO is beat against a crystal oscillator to create a low-frequency audio beat note, a slight "burble" can be detected. As the DTVFO was an experimental project rather than a construction project, no further effort was made to improve the filtering.

I found in the breadboarded circuit that considerable sidebands were created not by poor lock-line filtering, but by lock-line pick up. The line was changed to coax and the sideband content dropped considerably.

The VCO tuning slug should be adjusted for the middle of the range to be covered. This is more readily accomplished if the lock line is disconnected from the phase detector and connected to a variable source of voltage. The voltage should be set to 2.5 volts and the slug set to the center of the frequency range to be covered, or in this case 5.5 MHz. When the lock line is reconnected and the circuit tuned to

5.5 MHz, the lock-line dc voltage should read close to 2.5 volts.

The VCO must be capable of swinging at least 20 per cent beyond the frequency range to be covered when the lock line is varied from 1 to 4 volts. Except for the photo transistor part of the encoder, the circuit is pretty well foolproof and, provided the correct connections are made, should work right off. Because my junk box yielded two different photo transistors, some difficulty was encountered obtaining a match from the two devices and one of them was somewhat temperature conscious. However, it was felt that proper devices would solve this problem so it was shelved until better devices were obtained.

When tuning ssb signals, the DTVFO will feel exactly like a conventional VFO. After using the DTVFO for a few days I was amazed at how sloppy my previous listening habits had been. I'm quite sure that I'd been listening to stations with the receiver 100 Hz or more off frequency. Consequently, the fact that the DTVFO comes in 100 Hz steps made no difference. On CW, however, as the signal is tuned, the steps are guite noticeable, but make little difference to the copy.

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ham radio

printed-circuit layout using the longhand method

Electronic circuits require some method of connecting the parts together. This can be done by the printed-circuit board, which is made using photographic methods, or by point-to-point wiring. Obtaining the photo negative for the PC board is usually beyond most home workshop capabilities. I'd like to describe what I call the "longhand circuit layout" method. This method allows you to make PC boards without using a photo negative. With it, you can quickly lay out and build a PC board. My method differs from the photo approach in that you draw the circuit pattern on the copper-clad board.

materials

Fig. 1 shows the materials required for laying out a circuit. The graph paper should have 10 squares per 25.4 mm (10 squares per inch). This allows for IC pin spacing of 2.54 mm (0.1 inch). The copper-clad board can be cut to size by scoring a line on both sides and snapping it in two. Of the three types of etchant resist shown, I prefer the felt-tipped pen because it's easiest to use. The pen and etchant

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resist can be obtained from Radio Shack and other sources. (You might have to thin the resist to draw fine lines with it.)

procedure

The first step is to draw your parts onto the graph paper - the "longhand method." Do this by placing a dot for each end or pin of the part. For resistors I use 13-mm (0.5-inch) spacing between the dots. Then I draw the schematic symbol for the part between the dots. This helps keep track of the parts. If I



fig. 1. Materials needed for the "longhand layout method" for making PC boards. Shown are a scored copper-clad board, etchant, etchant resist, etchant-resist pen (preferred), and the circuit laid out on graph paper.

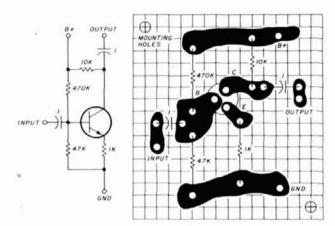
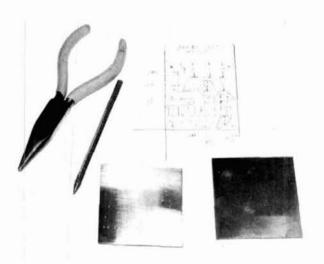


fig. 2. Typical circuit layout on graph paper, which has 10 squares per 25.4 mm (10 squares per inch). The copper-clad board should be taped to the back of the circuit layout so that the dots in the pattern can be punched onto the board.

have the room, I put the value of the part next to it, which helps me later when I insert the parts into the board.

Once the pattern is drawn onto the graph paper, the next step is to transfer it to the copper-clad board. With the copper-clad board beneath the graph paper, take a sharp-pointed tool and tap the dots on the graph paper. Important - the board must not slip while you're doing this. I prevent slippage by taping the board to the back of the graph paper.

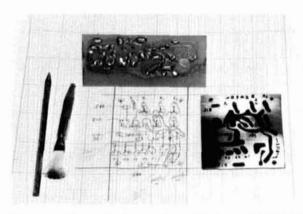


Tools and materials used in the "longhand layout method." The long nose pliers are used to tap the center punch, which transfers component-mounting holes from the graph paper to the copper-clad board. Note that one of the copper-clad boards (left) hasn't been thoroughly cleaned.

When all the dots have been punched, remove the board from the graph paper. With the etchant resist, draw the same pattern on the copper-clad board by connecting the correct dots together. Don't worry should you connect the wrong dots. If this happens, let the resist dry. It can be removed with a scribe or typewriter eraser.

The board can now be etched. When it has been etched, remove the resist by rubbing the board with steel wool or a household cleanser.

Now drill the holes, which will be easy since you



The typewriter eraser is used to remove unwanted etchant resist. At top is an example of a pattern on a copper-clad board using the "longhand layout method." A completed built-up PC board is shown at right.

have already centertapped each hole when you transferred the dots from the graph paper to the copper-clad board. I use a no. 60 drill, which is 1 mm, or about 0.04 inch.

You are now ready to insert your parts. Refer to your graph-paper pattern and put the parts into the correct holes on the board. That's all there is to it. Once you've used this method I think you'll agree it's a fast, simple way to make circuit boards.

A couple more things need mention: one is to clean the copper-clad board before applying the resist. If the board isn't free of dirt and oil, the resist will not stick to the copper. Also, the more information you can print on your pattern, the easier it will be to insert parts and wire to switches or any other part that's not on the board. So it's a good idea to plan your layout with this in mind.

I hope the photos and drawing will answer questions I may have overlooked in the text. I've been using this method for ten years and haven't found its equal.

ham radio

monolithic crystal filters

Introducing the MCF,
a new and
improved development in
crystal-filter technology —
a report on
its applications in
communications electronics

When high selectivity in electronic circuits is a must, mechanical filters, ceramic filters, or crystal filters are used. A new type of crystal filter is becoming popular: the monolithic crystal filter (MCF). It replaces the conventional crystal filter and leads to new applications for this new product, because the MCF shows comparable electrical features but is smaller and can be made more economically in large-scale production. What is a monolithic crystal filter, and what is the difference between it and conventional crystal filters?

conventional crystal filter

Conventional crystal filters consist of one to ten single quartz crystals coupled together by inductors, capacitors, and resistors in a distinct way to yield the filter characteristics required by the design specifications. As a typical example, **fig. 1**, shows the internal circuit of the well-known XF-9B bandpass filter of KVG* which is a standard filter for single-sideband applications. It's an eight-pole filter, which means it

includes eight crystals. The bandwidth is ± 1.2 kHz at -6 dB attenuation (related to the passband) at a center frequency of 9 MHz.

Each stage consists of tuned half-lattice bridges with one crystal in each branch. The stages are coupled through C4 and C5. For matching the filter impedance to 500 ohms, the input and output circuits have stacked windings. To compensate for strays, both resonate circuits are tuned to a higher frequency. With external trimming capacitors the filter can be aligned to the exact center frequency, which is also the point of minimum passband ripple.

monolithic two-pole (dual) filter

Instead of single crystals, the monolithic crystal filters use multiple resonators, which consist of several vibrating systems plated onto one common crystal blank. Typically, these vibrators are of the thickness shear type (AT cuts) with a frequency range from 5-30 MHz in the fundamental mode (plano-parallel crystal blanks). Most types of multiple resonators work in the fundamental mode, but the same principle can be applied to overtone resonators as well.

The simplest arrangement of multiple resonators is the monolithic two-pole filter, referred to as "dual." What is its difference compared with single crystals?

The equivalent electrical circuit of a single crystal is shown in **fig. 2**. It involves a high-Q series-resonant circuit consisting of the motional parameters L1, C1, R1 and the static capacitance, C_{θ} , in parallel. The Q is greater than 50,000.1,2

The series-resonant frequency is

$$f_s = \frac{1}{2\pi\sqrt{L_1C_1}} \tag{1}$$

and parallel-resonant frequency above f_s is

$$f_p = \frac{1}{2\pi\sqrt{L_1\left(\frac{C_0C_1}{C_0 + C_1}\right)}} \tag{2}$$

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^{*}Kristallverarbeitung Neckarbischofsheim, West Germany; available in the United States and Canada from Spectrum International, Post Office Box 1084, Concord, Massachusetts 01742.

The monolithic two-pole filter has two pairs of electrodes, which are plated onto one common crystal so that both resonators are mechanically coupled to each other by the crystal blank in a well-defined magnitude. This leads to the equivalent electrical circuit shown in **fig. 3**.

Both resonators consist of the motional quantities L1, C1, R1 and L1', C1', R1'. The static capacitances across each pair of electrodes are C_0 and C_0' . The four-pole in **fig. 3** (dashed lines) involving C_k and the negative capacitances $-C_k$ —a so-called impedance inverter configuration³—represents the mechanical coupling between both systems.⁴

The coupling factor, k is

$$k = \frac{C_1}{C_k} \tag{3}$$

and has an order of magnitude of 10^{-4} - 10^{-3} in fundamental-mode duals. Overtone duals have smaller coupling factors. The coupling factor of a particular dual is determined by the configuration of the electrodes, the gap between them, the crystallographic direction of coupling, the mass of the electrode plating, and the thickness of the quartz blank.⁵

The resonant frequencies of both resonator pairs are normally equal or very close to each other:

$$f_1' \approx f_1 = \frac{1}{2\pi\sqrt{L_1C_1}}$$
 (4)

Because of the coupling effect these frequencies can't be measured directly. For example, between pin A and pin B, with pin C open (fig. 3), you can measure a frequency f_I^* , that's lower than that obtained with eq. 4.*

Two new resonant frequencies occur, which are the characteristic vibration modes of monolithic dual resonators. These frequencies are of great significance for the application of duals as filter components:

1. When both resonator systems are connected in parallel, as shown in **fig. 4A**, series resonance appears at the so-called "symmetric frequency":

$$f_{sym} = f_1 \sqrt{1-k} \tag{5}$$

In this circuit both systems vibrate in equal phase. This means that the mechanical displacement (of the

*More precisely, two frequencies exist:

$$f_{1,2}^* = f_1 \sqrt{1 + \frac{C_1}{2C_0} \pm \sqrt{k^2 + \frac{C_1^2}{2C_0^2}}}$$

The lower frequency is between f_{sym} and f_{asym} , where f_{sym} and f_{asym} are the "symmetric frequency" and "antisymmetric frequency" as described in the following text.

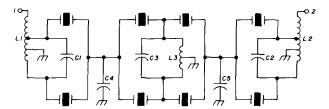


fig. 1. Internal circuit of KVG filter XF-9B.

thickness-shear motion of the crystal) takes place in the same direction in both systems.

2. Connecting both resonators as shown in **fig. 4B**, series resonance appears at the "antisymmetric frequency," f_{asym} , which is above f_{sym} :

$$f_{asym} = f_1 \sqrt{1+k} \approx f_{sym} (1+k)$$
 (6)

In this configuration both systems vibrate in phase opposition.

Additionally, both frequencies, f_{sym} and f_{asym} , can be measured between pins A and B with short circuited output pins (B and C). In this case, symmetric and antisymmetric frequencies are the frequencies of maximum input admittance of the four pole.⁶

The frequency difference between both characteristic frequencies — often called "mode spacing" — increases with higher coupling factor, k, as you can see in eq. 6.

monolithic multipole resonators

The principle of the monolithic dual resonator can be expanded and leads to monolithic multipole resonators with up to eight or ten resonator systems on the same crystal disc.

The mathematical synthesis of such vibrators is complex. Also measuring and production techniques are difficult. Each type of filter needs a certain configuration of the resonators. This restricts the feasibility of economically producing a large number of filter specifications in smaller quantities. Furthermore, with a larger number of resonators, the problem of spurious responses increases. This is why the multipole monolithic crystal filter hasn't become popular except for special applications such as channel filters. The tendency is to obtain multipole crystal filters by stacking several dual resonators, as explained below.

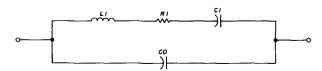


fig. 2. Equivalent electrical circuit of a crystal.

monolithic crystal filters

We now consider the MCF in its role as a practical addition to electronic circuits. Examples are given to show how the dual can be used in noise blankers. noise filters, and fm discriminators. We then examine the MCF in another practical application in which more than two poles can be synthesized by connecting several duals in series. First, some background on two-pole filters.

Two-pole filter characteristics. As shown by the theory of network synthesis and by the theorem of Bartlett,8 the equivalent electrical circuit of a dual (fig. 3) can be transformed into an equivalent halflattice bridge with a differential transformer, which has the same response of amplitude and phase versus frequency.9

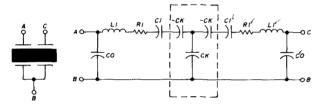


fig. 3. Configuration and equivalent electrical circuit of a dual resonator.

When a dual is terminated at its input and output with an impedance value, $Z_{in} = Z_{out}$, the elementary form of a two-pole crystal filter results. The identical selectivity curve is shown as the equivalent half-lattice, two-pole filter with single crystals terminated with the same impedances (fig. 5).

Comparing the number of components in both filters shows clearly that a monolithic filter is much easier to construct: there is no need for a differential transformer and, instead of two crystals, only one crystal component is required.

The bandpass response of such a filter depends on the magnitude of the termination resistors. The characteristic impedance is

$$R_0 = 2\pi L_1 \Delta f \tag{7}$$

where L_1 is the motional inductance of one resonator (see fig. 3), and Δf is one-half the filter bandwidth. Typical duals have characteristic impedances of several kilohms at, for example, 10.7 MHz.

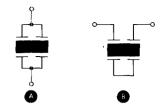


fig. 4. Pin configurations for measuring symmetric (A) and antisymmetric (B) frequencies.

Reasonable bandpass filter responses (with curves having a near-rectangular passband characteristic) can be achieved with terminating impedances of $R = Z_{in} = Z_{out}$ smaller than R_0 . The smaller the R value, the higher the passband ripple, but the skirts of the filter characteristic will be steeper.*

By proper selection of the termination impedances and characteristic frequencies of the dual, every filter response known from the filter theory of effective parameters (i.e., Chebyshev, Gaussian, Bessel, Legendre)9 can be synthesized.

The skirts of this filter can be made steeper by introducing a coupling capacitor, C_A between both resonators. This capacitance produces an attenuation peak at both sides of the passband. But at the same time the stopband attenuation decreases, as shown in fig. 6, for several values of C_A (as multiples of C_0), 10

To fulfill the total selectivity demands of a special device (e.g., a receiver), such a two-pole filter will surely not be sufficient. Despite this disadvantage. there are some interesting application examples for this simple filter component.

Dual as an i-f preselector in noise blankers. Noise blankers are designed to blank out short-duration noise pulses with high amplitude, especially in shortwave or mobile receivers. The best point at which to insert a noise blanker in a receiver is ahead of the i-f stages before the crystal filter, which provides the main selectivity. This is because narrowbandwidth filters produce ringing, which distorts the signal. 11

The block diagram of a typical noise blanker is shown in fig. 7. It includes two alternatives to obtain the noise information. Version A derives it from the i-f signal; version B obtains it from a separate noise receiver tuned to a "silent" frequency. 12

Following the mixer a broadband i-f filter must be

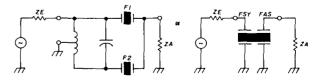


fig. 5. Two-pole crystal filter with differential transformer and equivalent monolithic crystal filter.

^{*}The classical wave-parameter theory yields the following example: For $\frac{R}{R_{\theta}} = 0.8$, the passband ripple is about 0.1 dB, while the bandwidth at attenuation is at the three-fold bandwidth a -3 dB. With $\frac{R}{R_0} = 0.5$, the ripple increases to 0.22 dB, while the -20-dB bandwidth is only 2.8 times the - 3-dB bandwidth.

inserted, one narrow enough to cut off strong out-ofband signals but broad enough to avoid ringing. A monolithic dual is a unique device for this i-f preselector, as it can be inserted directly with pure ohmic termination and without any alignment effort — if the dual spurious responses are negligible.

For the well-known 9-MHz crystal filter line, KVG offers its dual type XF-912 for a bandwidth of ± 7.5 kHz (at -3 dB). It is housed in a 3-pin HC-18 case and needs terminating resistances of 4.0 kilohms for a Chebyshev response. Different types with other bandwidths are available on request.

Dual as a noise filter. Normally, the main selectivity of commercial receiver i-f strips is produced by a high-performance (e.g., eight-pole-type crystal) filter ahead of the i-f amplifier stages. But usually the following broadband high-gain amplifier stages generate broadband noise, which reaches the second detector in full magnitude. This noise, which can be very inconvenient, especially at extremely narrow bandwidths (as with the CW-filter type XF-9 NB from KVG), can be reduced drastically by a simple filter directly ahead of the demodulator stage. This is one more typical application for a monolithic dual, which can be easily inserted without any adjustments. For example, the type XF-912 can be used again. Similar duals exist also for other i-fs such as 10.7 MHz or 21.4 MHz.

Dual as an fm discriminator. IC quadrature detectors are frequently used to demodulate fm signals. The principle on which these demodulators work is as follows.

The frequency-modulated i-f signal is applied to one port of an AND gate; the other port is connected to a phase-delayed portion of the same signal. The

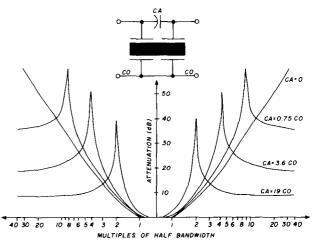


fig. 6. Frequency response of a dual with attenuation poles achieved by an additional capacitor, C_Δ .

phase delay depends on the frequency. The gate output yields the demodulated audio-frequency signal if followed by an integrating RC lowpass filter.

The phase-delay circuit is a parallel-resonant circuit, which is coupled through a choke or a small capacitor, as shown in **fig. 8A**. This circuit provides a phase delay that increases or decreases linearly with frequency changes in the vicinity of the resonant frequency. ¹³ Higher slopes of the phase vs frequency curves, and thereby higher recovered audio, can be realized by using a bandpass filter with a coupling coefficient of about $Q \cdot k = 0.7$. This is shown in **fig. 8B**. ¹⁴

The bandpass-filter can be substituted with a conventional dual resonator, which is designed as a Bessel-function filter with linear phase characteristics. The result is excellent demodulator linearity as

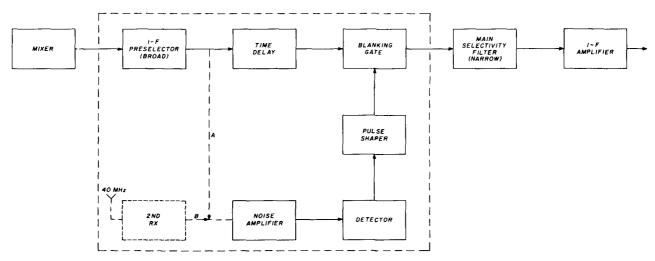


fig. 7. Block diagram for a typical noise blanker. Version A: noise information derived from i-f. Version B: noise information derived from separate noise receiver.

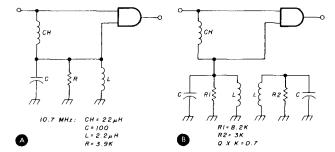


fig. 8. Schematics of f-m quadrature detectors - (A) with single resonant circuit; (B) with bandpass filter.

well as high audio yield. The demodulator response is the typical S-shaped characteristic. The two peaks are approximately at the symmetric and the antisymmetric frequencies.

Fig. 9 shows a circuit working on this principle. It consists of the RCA integrated circuit CA3089E. Both terminating resistors, R1 and R2, are chosen for best phase linearity (*i.e.*, constant group delay) in the passband between the two peaks. Their values depend on the motional parameters, L_1C_1 , and the mode spacing of the dual.

This principle can be generalized for other quadrature detector ICs such as ULN2113A (Sprague), TAA661 (Signetics), or TBA120S (Siemens).

multipole filters with n > 2

Monolithic crystal filters with more than two poles can be synthesized by connecting several dual resonators in series whereby they are coupled to each other by capacitors to ground (*i.e.*, the common electrode). As an example, **fig. 10** shows the internal

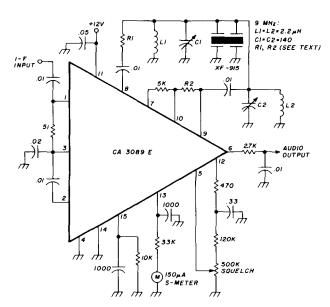


fig. 9. F-m demodulator with dual as discriminator.

circuit of the KVG monolithic filter XFM-107B (bandwidth \pm 7.5 kHz at 10.7 MHz). Input and output are terminated with tuned circuits, which transform the filter impedance to a standard value of 910 ohms (with $C_{ext} = 25 \ pF$ in parallel).

As with single duals, such composite filter structures can be terminated directly with a pure ohmic resistance given by the filter synthesis. In this case both tuned circuits can be omitted.

Furthermore, with increasing bandwidth up to about 1 per cent (i.e., 1 part in 100) of the center frequency, the coupling capacitors become so small that they are realized by the static input and output capacitances of the coupled duals plus stray capacitances alone. Then the simplest structure of a monolithic crystal filter can be achieved. It consists only of

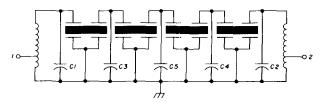


fig. 10. Internal circuit of KVG-Monolithic filter XFM-107B.

a chain of directly coupled duals, as you can see in **fig. 11** for the KVG monolithic filter XFM-107S03 (10.7 MHz bandwidth \pm 10.6 kHz).

Comparing this circuit with that of a discrete crystal filter as in **fig. 1**, the simplification is evident. Further, the half-lattice filters with more than five crystals usually need a third tuned circuit (see L_3 , C_3 in **fig. 1**), which gives additional insertion losses that can't be compensated for by the termination. This isn't necessary with monolithic filters because they present much smaller amounts of insertion loss than conventional crystal filters.

The electrical properties of monolithic crystal filters are equivalent to those of crystal filters with discrete components. That's why you can realize all known filter responses in monolithic structures, but with an upper limit for the bandwidth. Generally, filters with monolithic crystals vibrating in the nth overtone mode can be obtained with relative bandwidths smaller by the factor $\frac{1}{n^2}$.

The theoretical filter curves for ideal (lossless) filters — that means, the responses of attenuation and phase *vs.* frequency — are cataloged in normalized representation in the literature on the subject.^{9,16}

summary

Monolithic crystal filters (MCFs) stand for the sim-

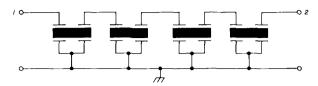


fig. 11. Internal circuit of KVG-Monolithic filter XFM-107S03.

plification of the structure of multipole crystal filters. Instead of discrete crystals, MCFs contain multiple crystal systems on a common quartz blank, which are mechanically coupled between each other by the quartz blank.

The typical application is in the frequency range of AT-cut crystals, most in the fundamental mode, but with increasing importance also as third or higher overtone monoliths. However, bandwidth is somewhat reduced in overtone applications.

The most usual forms of MCFs are dual resonators and series configurations of duals-to-high-pole filters. Beyond this, there are new interesting applications for duals as i-f preselectors, simple noise filters. or f-m demodulators.

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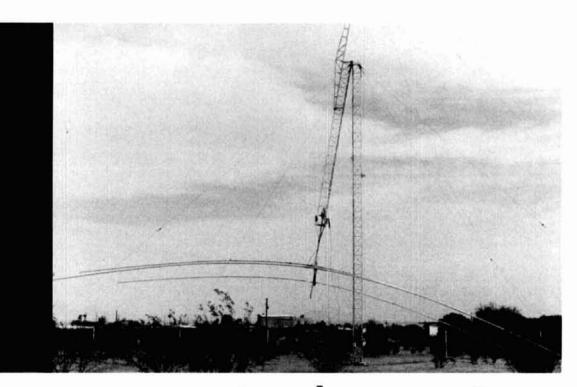
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rotary beam antenna for 40 meters

Recipe for a really big signal on 7 MHz: a 3-element Yagi beam on a foldover tower

A full-sized, three-element 40-meter rotary beam has been a dream of mine for many years. They are scarce in the Midwest, where I operated as W9ERU for many years. I personally, at one time, got as close as the Hy-Gain DB24, a shortened two elements on 40, and three elements on 20. When I moved to the Phoenix area, I initially tried a conical monopole. However, not being satisfied with the results, I finally put up the 40 meter only version of the DB24. This antenna had two folded elements, each about 13.5 meters (44 feet) long, on a 4.9 meter (16 feet) boom.

tower design considerations

I mounted the two-element short beam, and later the full-sized antenna, on a 22-meter (72-feet) Rohn model 45 foldover tower. Legs on this tower are about 46 cm (18 inches) apart. I decided to mount the hinge between the fourth and fifth sections above the ground, giving three sections above the hinge. When the tower was folded over, this would allow the antenna to hang about 4 meters (13.3 feet) above the ground. In addition, it appeared that there would be enough clearance to allow the ends of the elements to pass the ground in such an operation. And, the portion above the hinge, would only be about 9 meters (30 feet) long, reducing the strain on the hoisting mechanism. Guy wires, two sets of four, were attached to anchors about 15 meters (50 feet) out from the base, and about 21 meters (70 feet) apart in a square configuration. See fig. 1. These dimensions were chosen to allow the use of a 12-meter (40-foot) boom at the top and still get it to pass through two of the lower guy wires.

construction

There are many parts to a beam antenna installation such as mine. I've provided photos showing

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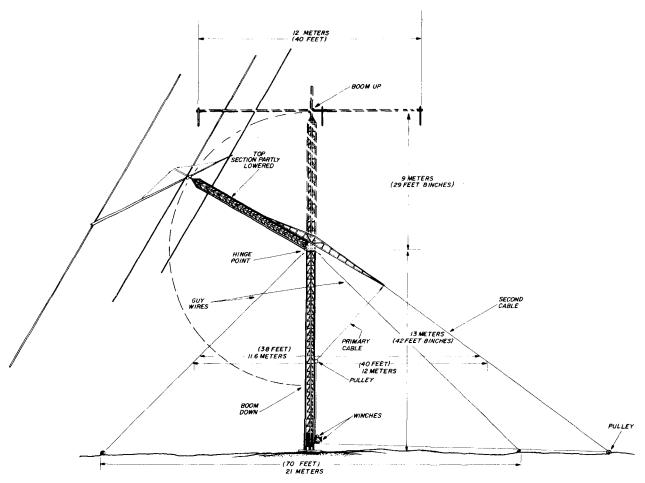


fig. 1. Overall dimensions of the 40-meter, three-element beam. Antenna is mounted on a Rohn model 45 foldover tower 22 meters (72 feet) high. Two sets of guys are used and two winches assure safe operation.

essential hardware and some views showing the antenna during and after installation.

Tower and rotator hardware. The tower lowering and raising mechanism is extremely important. A runaway when raising or lowering the antenna would result in a disaster. So I used two winches and cables. One, supplied by Rohn, was hand operated and located at the tower base. Its cable is passed through a pulley part way up the tower and to the

This article is sprinkled with many metric conversions, which tend to disrupt continuity and ease of reading. But there's a good reason: we must face the fact that the outmoded and cumbersome English system of measurements is rapidly becoming extinct in the technical literature. So to help in the transition, author Hubbell's article has been edited to show first the metric dimensions, followed by English equivalents in parentheses. We apologize for this slight inconvenience. Sometime in the not-too-distant future we'll discard all references in our articles to the English system of measurements. We've graduated from tubes to solid-state devices without too much trouble. Let's progress further with the metric system, a totally logical and convenient method of defining measurements. Editor.

tailpiece or tower boom, which went up as the top end of the tower came down. A second cable from a motor-driven winch at the tower base went out to a pulley some 18 meters (60 feet) from the base and up to the same tailpiece. Thus, pull could be exerted from two directions (see fig. 1).

The prop-pitch rotator and selsyn indicator chaindrive system were conventional. I welded a bell housing to the prop-pitch motor top gear and connected the housing to a piece of 41-mm (1.61-inch) pipe*, which passes through the top of the tower and is held to the boom saddle on top. This pipe is actually in two pieces, joined by a larger diameter sleeve and bolted through the drive pipe and the sleeve. It is not a one-piece shaft. A bicycle chain-and-sprocket system drives a 115-volt ac selsyn, which was offset from the main rotator pipe drive. It also drives another similar selsyn in the tower base. A limitswitch system allows the rotating pipe to turn about 450 degrees.

^{*}The pipe used was called "Inch and a half" water pipe. In actuality, the pipe has a standard inside diameter of 4 cm (1.61 inches) and an outside diameter of 4.8 cm (1.9 inches).

Boom construction. The boom was made of two 6meter (20-foot) lengths of 102-mm (4-inch) diameter irrigation tubing. To couple these tubes required a sleeve. I found only 0.4 mm (26 gauge) galvanized iron readily available, so two pieces 0.6 meter (2 feet) long and about 0.3 meter (12.5 inches) wide were rolled into tubes, one inside the other, and filed down until this double-walled sleeve just fit inside the 102mm (4-inch) tubing, 0.3 meter (12 inches) each way. All three pieces were drilled for several rows of blind rivets, which made a strong joint. Now I had a boom 12 meters (40 feet) long (see fig. 1).

At the junction of the two boom halves, a hole about 33.5 mm (1-5/16 inches) in diameter was drilled through the boom to pass a piece of "one inch" pipe.

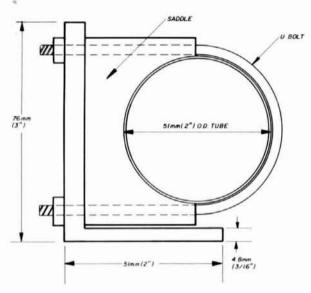
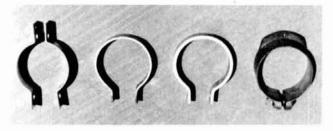


fig. 2. Details of boom-to-element clamps. The clamps are known as "muffler clamps" and were obtained from an automotive supply store.

This pipe is used as the vertical support for a truss system to take up the vertical strain on the boom.

Boom mounting. I made a U-shaped saddle of 6mm (0.25-inch) thick steel with two sides 76 by 508 mm (3 by 20 inches) and a bottom 102 by 356 mm (4 by 14 inches). A 41-mm (1.61-inch) pin, 152 mm (6 inches) long, was centered in this piece (see photo). All items were welded together and holes were drilled through the sides of this saddle 25.5 mm (1 inch) from each end and 51-mm (2 inches) above the inside bottom piece. This allowed for 12.5-mm (0.5-inch) bolts to be passed through the saddle and boom. The bolts were 45.7 cm (18 inches) apart. Enough clearance was made to allow the boom to be tilted on a single bolt if desired.

A 16-mm (0.625-inch) hole was drilled axially into the 41-mm (1.61-inch) pin. The hole was 76 mm (3 inches) deep. This hole was for another pin large



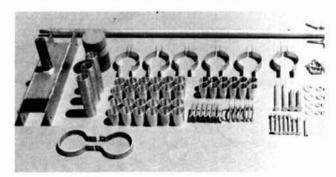
The boom clamps. The two-piece steel clamps, left, were used in the final version.

enough to fit inside a 27-mm (1.049-inch) diameter aluminum pipe turned down to 16 mm (0.625 inch), for 76 mm (3 inches) projecting from the pipe, which was about 2.5 meters (8 feet) overall. When saddle, boom, and this guying pipe were assembled, a guying system took the strain 3 meters (10 feet) out from the center of the boom on each side, up to the top of the 2.5-meter (8-foot) vertical guy support. Turnbuckles allowed adjustment.

Boom-to-element mounts. A horizontal aluminum angle holds the center of the element. This angle is 81 cm (32 inches) long, made of 51 by 76 by 4.8 mm (2 by 3 by 0.187 inch) material. Two bolts near the center hold the element to the 51-mm (2-inch) face, and two muffler clamps near the outer ends of the angle hold the element to the 76-mm (3-inch) face (fig. 2). These muffler clamps were marked "1-7/8 inch" but fit 51-mm (2-inch) tubing perfectly. They were bought at an automotive parts supply store.

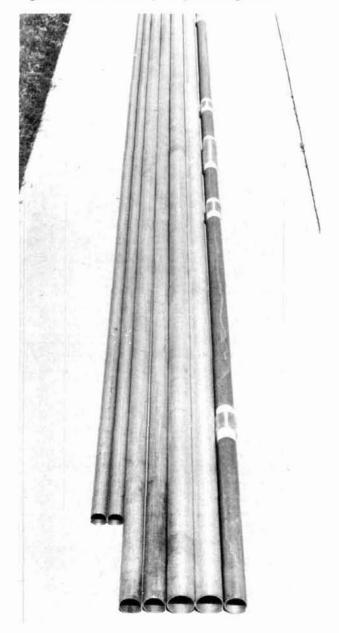
These angles are mounted crosswise with respect to the boom. They are bolted to two 30-cm (12-inch) lengths of 38 by 38 by 4.8 mm (1.5 by 1.5 by 0.188 inch) aluminum angle, positioned 41 mm (1.63 inches) apart (see photo). These 30-cm (12-inch) lengths were in turn bolted to two clamps, which passed around the boom. Two bolts hold the 38-mm (1.5-inch) angles to the clamps. All three angles are secured by a triangular gusset made of aluminum 3mm (0.125-inch) thick, 45 by 45 by 63 cm (18 by 18 by 25 inches) with its corners cut off about 25 mm (1 inch) from the ends on the long dimension.

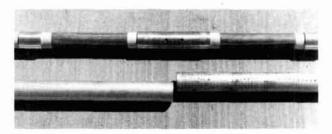
Hardware including boom saddle, boom guy, sleeves, clamps, and other items described in the text.



The boom-to-element clamps changed over the years. In Bill Orr's Beam Antenna Handbook (1955) edition), clamps were shown made of cast aluminum intended to hold irrigation tubing together. When some of these broke I replaced them with clamps made from 25.5 by 6.5 mm (1 by 0.25 inch) aluminum bar bought at a hardware store. When these broke I had others made at a blacksmith shop from 25.5 × 3 mm (1 by 0.125 inch) steel. All had a flaw. The degree of tightening needed on the clamp made a difference in the spacing of the aluminum angles mentioned above. The two-piece clamp shown at the left-hand end of the four in the photo solved the problem by allowing the adjustments to be made independently of each other.

All parts for the driven element (seven pieces of tubing). Light-colored bands are tape for positioning the sleeves.





Element center showing reinforcing tubing.

Antenna elements. Now I could mount elements to the boom, but I needed the elements. On the first try I used two 9-meter (30-foot) lengths of 51-mm (2inch) irrigation tubing joined at the center by a 152mm (6-inch) sleeve of aluminum pipe turned to fit inside the irrigation tubing. Short extensions on the ends brought the element lengths to the design figures of 19.8 meters, 20.3 meters, and 21.3 meters (64.8, 66.6, and 69.7 feet) for the director, driven element, and reflector respectively. I arrived at these numbers for the design frequency by the formula:

140.2/7.1 meters or 460/7.1 feet (director length) 144.2/7.1 meters 473/7.1 feet (driven-element length) 150.9/7.1 meters 495/7.1 feet (reflector length)

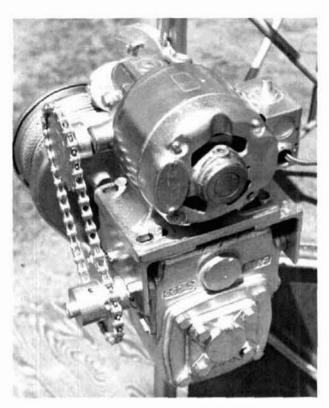
Results were unsatisfactory. Besides being plagued by an intermittent short circuit, which I eventually found in my gamma match, a windstorm soon bent the soft irrigation tubing. Back to the drawing board.

Through a local office of a tubing-supply firm, I bought six lengths of 6061T6 duraluminum 3.7 meters (12 feet) long, 51 mm (2 inch) diameter, with 1.2 mm (0.049 inch) wall. I also got six lengths of 3.7 meters by 41.3 mm by 1.2 mm (12 feet by 1.63 inches by 0.049 inch) tubing, and, from my stock of used tubing, six 3.7-meter (12-foot) lengths of 32-mm (1.25-inch) material with 1.7-mm (0.065-inch) wall thicker than needed, but I had it on hand.

By overlapping the tubing about 30 cm (1 foot) at the joints between the 51-mm, 41.3-mm, and 32-mm (2, 1.63, and 1.25-inch) tubes, I had 10.4-meter (34foot) lengths, and two of these gave me elements 21 meters (68 feet) long. The butt joint in the center (two 51-mm, or 2-inch, tubes) and the overlapping joints of different size tubing were all joined by sleeves turned from short lengths of aluminum pipe nominally 48 mm (1.9 inches) OD and from aluminum pipe couplings that fit inside the 41.3-mm (1.63-inch) tubing over the 32-mm (1.25-inch) tubing.

Additional sleeves were inserted inside the 51-mm (2-inch) tubes for support where the muffler clamps were placed. All overlapping joints and sleeves were held in place by steel 10-32 (M5) bolts held with lockwashers and nuts.

I did some more reading on element lengths for



Closeup of the homebrew slow-speed winch.

close-spaced, three-element yagis, and changed element lengths according to:

138.7/7.080 meters or 455/7.080 feet (director) 144.8/7.080 meters 475/7.080 feet (driven-element) 152.4/7.080 meters 500/7.080 feet (reflector)

This gave me new element lengths of 19.6, 20.5, and 21.5 meters (64.3, 67.1, and 70.6 feet).

I had written Bill Orr, W6SAI, for advice on the element lengths and he suggested that tapered elements might give a higher resonant frequency than the design figures indicated. The change was slight: from 7080 kHz to about 7120 kHz.

Matching section. Bill also strongly recommended that I use a small-diameter gamma rod, "as large diameter gamma rods detune the driven element." So I used triple-stranded 2-mm (12-AWG) copperweld guywire. The gamma rod terminated in a 127by 152- by 229-mm (5- by 6- by 9-inch) aluminum box mounted on the gusset plate of the driven elementto-boom mounting. Inside was an air-variable capacitor for the gamma match, about 35 to 200 pF, driven by a reversible motor with a 4-rpm gearing. An added fixed capacitance of 50, 100, or 150 pF could also be switched in parallel with the variable capacitor.

The other end of the gamma rod was grounded to the driven element by an aluminum arm clamped to the gamma rod and the driven element. To get a good contact to the three strands of 2-mm (12-AWG) wire, the strands were soldered together, and a 19mm (0.75-inch) OD aluminum cylinder, with a hole to fit snugly over the three strands, was placed over the soldered area and fastened with three set screws.

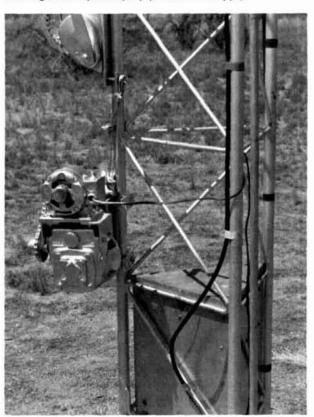
The aluminum grounding arm from gamma rod to driven element was hinged in the middle to allow for motion of the driven element, and a piece of heavy shield braid paralleled this joint. The ends were formed to fit around the aluminum cylinder and the driven element. Mating clamps held this assembly. The remaining outer end of the gamma rod wires was fastened to a long, coiled spring. The other end of this spring was clamped to the driven element with a stainless-steel hose clamp. The length of the gamma rod is about 188 cm (70 inches).

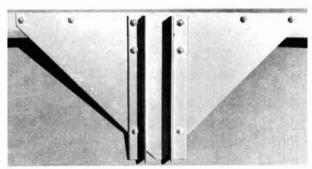
The gamma-capacitor box was arranged so that all items were supported on the cover, which could be removed by removing six thumb nuts. The gamma rod and ground connections were made by mating banana plugs and jacks.

adjustment and testing

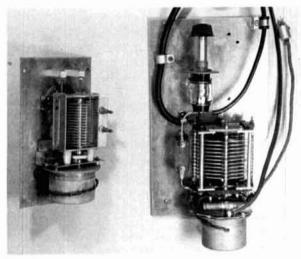
These procedures were performed with the beam in the down position, hanging about 4 meters (13 feet) from the ground. The motor-driven capacitor gamma match was operated from the ground at the tower base. The only adjustment made from a ladder was the length of the gamma rod, and the added tapswitched capacitors were reached from the tower

Tower base showing the two winches and metal box housing the selsyn and prop-pitch motor supply.

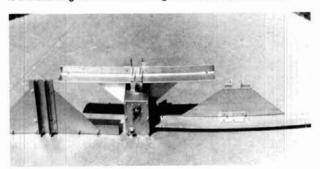




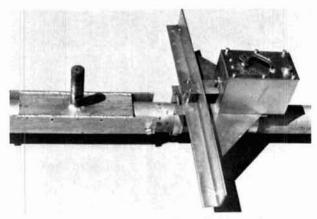
One side of an element-to-boom mount.



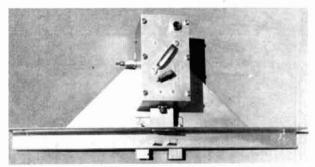
Gamma capacitor and motor-drive assembly, left. At right is a matching network for using the tower on 80 meters.



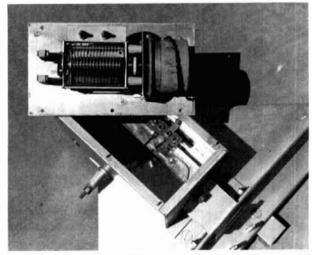
Element-to-boom mounts. Center assembly is for driven element with gamma box.



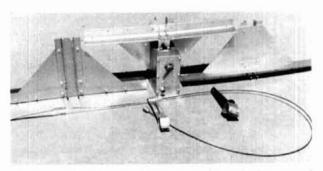
Boom center with boom saddle and driven element-toboom support in place.



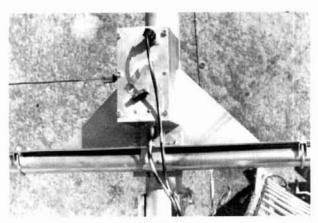
Driven-element-to-boom mount with gamma match box.



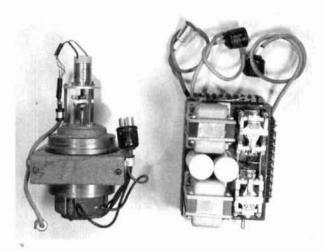
Driven-element mount with gamma capacitor and motor drive.



Element-to-boom mounts; gamma rod and clamp; and insulating bar, which clamps to driven element and reduces gamma-rod vibration.



View looking down from the driven-element mount.



Selsyn repeater system, left, and dc supply for the prop-pitch motor.

itself. A small vfo-controlled transmitter on the ground, with an swr bridge, provided the readings for adjustment. Unfortunately, the test readings in the down position didn't hold precisely for the beam up position. And the readings made with a 10-watt scale on the Bendix Micro-Match didn't directly compare with those taken with the output of a linear. Also, 60 plus meters (200 feet) of coaxial cable gave better readings in the shack than were obtained at the tower.

However, my 4-1000A linear1 was quite tolerant, as was my Johnson "desk kilowatt." So I managed quite well with an swr of 1.3 at 7000 kHz, 1.0 at 7050 kHz, and 1.25 at 7100 kHz, all with the same gamma capacitor setting. I also had 1.5 at 7150 kHz, 2.0 at 7200 kHz, 2.25 at 7250 kHz, and 1.9 at 7300 kHz, each with the gamma capacitor retuned for lowest swr. Many commercially built transmitters wouldn't stand up with the higher swr readings, though a simple matching system would take care of the problem.

precautions

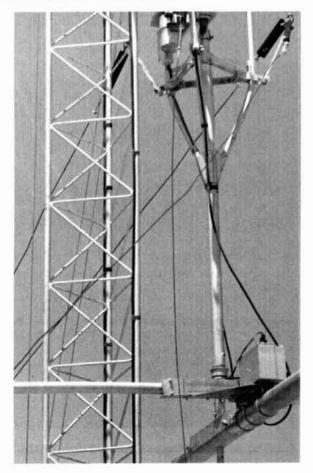
Many details haven't been covered so far; one is important. The weak point in the element-to-boom support was where the muffler clamps hold the elements. When the elements failed at this point in a 1974 wind storm, I decided that added strength was needed here and the stress should be distributed over a considerable section of the center portion of each element. Heavier wall in the 51-mm (2-inch) tubing or the addition of an internal tube to strengthen the center would do it. I used the latter method and put 3.7-meter (12-foot) lengths of 41-mm (1.63-inch) diameter tubing with 1.5-mm (0.058-inch) wall inside the elements, overlapping the 51-mm (2-inch) tubing 1.8 meters (6 feet) on each side. I used sleeves between the reinforcing tube and the inside of the element tube at the center, with the muffler clamps at the outer ends. They were all held in place by masking tape.

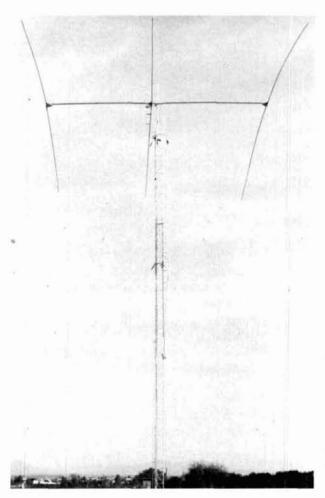
Internal support must be provided for the boom where the clamps are applied, as the tube wall is only 1.3 mm (0.05 inch) thick. Disks of 51-mm (2-inch) wood were turned to fit tightly inside the boom and driven into place where the clamp bands are installed. Friction at these points was all that kept the elements level and in place - except that the elements do hang under the boom rather than being placed on top, a difference of about 152 mm (6 inches).

Vibration dampers made of 13-mm (0.5-inch) rope were installed inside the elements, just a bit short of the full length, and fastened at the center with a bit of wire, making a loop where bolts would pass through during assembly to keep the rope from falling out when the beam was raised.

I built this antenna by myself and assembled it to the tower, which I also assembled. All the help I had was from W7EH, who helped me pour the guy wire anchors, and from the crane and operator who set the tower upright. I raise and lower the beam alone, not because I couldn't use help, but I like to go slowly and check and recheck for safety.

Tower inverted showing rotator, driven element and mount, and the gamma matching system.





The 40-meter beam on fully erect tower ready for action.

A word of warning; the Rohn tower is an excellent piece of equipment but is not rated to carry the top load I use when raising and lowering the tower. I have no idea what the margin of safety is. Mine has stood up very well for nearly ten years. But be warned - the foldover system is overloaded! Once up and in place, the two sets of four guy wires, well out from the base, make the tower safe even if the antenna is torn to pieces. The guy wires are insulated from the tower and ground. They're broken up with insulators so that the tower can be used for a vertical radiator on 80 and 160 meters.

Does it work? It sure does. It's very nice to enter a pileup of East-Coast hams working Europe or the Near East and make contacts right along with them. I'm getting too old for 48-hour contest stretches, but I thoroughly enjoy having a big sig on 40 meters.

reference

1. Gene Hubbell, W7DI, "Ecology Linear," ham radio, March, 1972, pages 6-15.

ham radio

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the Micoder:

some improvements

Modifications to the Micoder tone encoder, eliminating the 9-volt battery and adding a crystal-controlled encoder

Amateur Radio magazines have published articles on improving the early production models of the Heath Micoder. The faults found were in the tone outputs. Imbalance between high and low tones and poor frequency stability were prime complaints. My Micoder seemed to work well in these respects, but the *Touch-Tone** feature is not used that much and almost never at temperature extremes. My main complaint is that the 9-volt battery poops out about twice a year and disables the microphone as well as the tone encoder.

The solution is to power the device from the 12-volt supply in the rig. The Micoder cable contains an extra wire that's used as a ground in parallel with the shield. With the encoder components available to-

circuit description

The circuit is shown in **fig. 1**. Note that the microphone- and tone-circuit voltages are obtained from separate 9-volt and 5-volt zeners. The filtering provided in the rig (an HW-2036 in my case) and the 9-volt zener provide adequate rejection of hum (from the ac supply) and hash (from the mobile supply).

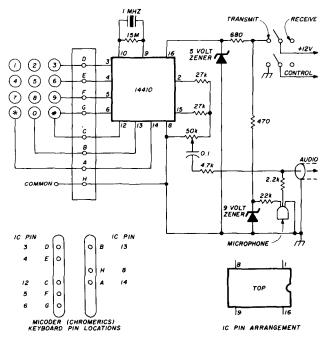


fig. 1. Modifications for improving the Heath Micoder. Circuit eliminates the 9-volt battery and adds a crystal-controlled encoder. Parts are available from Data Signal, Inc., and Radio Shack.

By George A. Wilson, W10LP, 318 Fisher Street, Walpole, Massachusetts 02081

day, this modification makes the improvement simple and inexpensive.

^{*}Touch-Tone is a trademark of American Telephone and Telegraph.

construction

The original circuit board was stripped of parts and reused to mount the keyboard and the output-level control. Make sure no solder bridges occur and that solder splashes don't cause shorts between circuits. This makes it possible to use the PC-board paths outboard of the keyboard pin receptacles as connection points for the wires to the encoder circuit. The battery and its connector and cable are removed and discarded.

Lift the black wire in the cable at each end. Solder a small tie point to the switch on the end opposite the existing tie points. The black wire (+12 volts) goes to the switch lug where the +9 volts from the battery was connected. The encoder circuit and 9-volt zener circuit are fed from the switch lug that previously accepted the red lead from the original circuit board.

The encoder is easily built using a PC board such as that sold by Data Signal, Inc.* They can also supply the chip and the crystal. A minor board modification is required to accommodate an extra connection for the output level control — a simple no. 60 (1 mm) hole drilled in a blank spot does the trick. When wrapped with plastic foam, the completed encoder

*Data Signal, Inc., 2403 Commerce Lane, Albany, Georgia 31707.

circuit occupies the space that was previously occupied by the battery.

The level-adjustment pot can be a miniature 6.4-mm (0.25-inch) shaft-type with a screwdriver slot. If this type is available, mount it where the LED was previously. Its shaft will protrude into the hole in the outer case previously occupied by the LED. A PC-type control can be cemented to the circuit board if the miniature unit previously described isn't available.

Connect the HW-2036-end of the black wire in the cable to the + 12-volt leads of the LEDs on the front panel. Lengthen the lead by about 25 mm (1 inch). The joint should be insulated with a short piece of sleeving.

This mod costs less than \$20, requires no new circuit board, and solves both the battery and encoder problems.

reference

1. Fallenbeck, "Micoder," QST, April 1978.

bibliography

- 1. DeLaune, "Digital Touch-Tone Encoder," ham radio, April 1975.
- 2. Lowenstein, "Hand-Held Touch-Tone," ham radio, September 1975.

ham radio

vhf/uhf preamplifier burnout

Soon after arriving in New England, I started to experience random burnout of my vhf/uhf preamps, both bipolar transistors and fets. Initially I blamed the burnout on electrical storms, but the problem increased drastically in the winter, a time when electrical storms are usually at a minimum. Occasionally I even lost second-stage preamps and multipliers in my local oscillator chain.

How could this be? All normal methods of burnout protection failed to reduce the failure rate. The plot thickened when I left one of the preamps terminated in a 50-ohm load and it still blew out!

Then I connected a high-impedance, battery-operated digital VTVM to the B + supply to the preamps. Everything looked fine until I keyed my kilowatt linear on 80-meter CW (where I spend most of my operat-

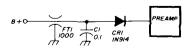


fig. 1. Circuit for preventing burnout of sensitive vhf/uhf preamplifiers in strong rf fields. C1 is a 0.1-µF mylar or ceramic disc capacitor installed between the feedthrough capacitor FT1 and the idiot diode CR1.

ing time during the winter) while using my east/west dipole; the dc voltage on the preamp went wild. When the west sloper was connected to the linear, the problem disappeared.

It developed that the problem was twofold: rf pick-up on the power-supply lines to the preamps, and rectification in the reverse-voltage protection diode (also known as the *idiot* diode). Eliminating the idiot diode or shortening the power-supply leads are not good solutions; placing a 0.01- μ F mylar or ceramic disc capacitor just ahead of the idiot diode as shown in **fig. 1**, however, prevents rf from reaching the diode and, hence, from being rectified.

No burnouts have been experienced with zener diode biasing as described in *ham radio*.* In the zener bias circuit the zener diode clamps and prevents the transistor voltages from soaring.

The circuit in **fig. 1** won't solve all your preamplifier burnout problems, but it should give longer life to those expensive low-noise semiconductors where large rf fields and high-frequency operation are prevalent. **Joe Reisert, W1JR**

^{*}J. Reisert, W1JAA, "Ultra Low-Noise UHF Preamplifier," ham radio, March 1975, page 8.

multiple quarter-wave matching transformers

There are many ways of matching the impedance of an antenna to the output impedance of a transmitter. These methods often have limited bandwidths over which they provide a match. For example, the popular quarter-wave transformer shown in fig. 1 has a bandwidth of approximately \pm 10 per cent for a vswr of less than 1.5:1. Fortunately, the bandwidth of this method can be easily increased by cascading several transformers (see fig. 2).

To speed up the design of multiple quarter-wave matching transformers, 1 several nomographs have been developed. To make these nomographs more useful to amateurs, only the characteristic impedance of standard coaxial transmission lines is shown. Fig. 3 shows a nomograph for a one-section, or standard quarter-wave transformer. The impedance values shown are values of characteristic impedance for coaxial type transmission lines. Fig. 4 shows the design nomograph for a two-section transformer and fig. 5 shows the nomograph for a three-section transformer. To decide whether to use the one-, two-, or three-section transformer, another nomograph is provided in fig. 6.

The nomographs will give the impedances of the individual transformer sections. The lengths of the sections can be found by the following formula:

$$L(meters) = \frac{75}{f_{MHz}} \times vf \text{ or } L(feet) = \frac{246}{f_{MHz}} \times vf$$
 (1)

where

L =length of the section

f = center frequency

vf = velocity factor

The velocity of propagation can be obtained from table 1.

design examples

The use of these nomographs can best be explained by the use of several examples.

Example 1. Assume you are trying to match an antenna impedance of 100 ohms to a 50-ohm transmitter. One solution to this problem is to use a one-section transformer. Use a straight edge to connect

100 ohms on the R_L line and 50 ohms on the R_O line. Read the characteristic impedance of the transformer as slightly under 72 ohms. A 73- or 75-ohm coax from table 1 may be used with little difference.

Example 2. You want to match a 250-ohm antenna to your 50-ohm transmitter, with the vswr not to exceed 1.5:1. The antenna is a log periodic that covers the 80- and 40-meter bands.

The first step is to compute the bandwidth of the antenna. Assume the frequency range of the antenna is 3.5 to 7.5 MHz. The bandwidth is 72.7 per cent $\left(\frac{7.5-3.5}{5.5}\times100\right)$, or about ±36 per cent. Since this exceeds the bandwidth of a one-section transformer, **fig. 6** is used to determine the correct number of sections.

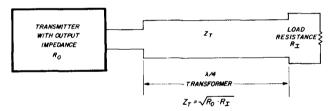


fig. 1. Schematic diagram of a standard one-section transformer. When terminated in a resistance R_L , the impedance seen at the transmitter end will be Z_T^2/R_L . By varying the impedance of the quarter-wave transformer, the load presented to the transmitter can be made to more closely match the transmitter's output impedance.

Lay a straight edge on **fig. 6** so that it lines up with 1.5 on the S scale on the left and with 5 on the R_L/R_O scale on the right. Draw a line between these two points and indicate where this line crosses the unmarked vertical line between the S and N scales. Now, move the straight edge so that it lines up with the dot on the unmarked line and the BW scale. Draw a line between these two points.

By Samuel Guccione, K3BY, 110 Chalet Court, Camden, Delaware 19934

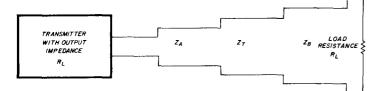


fig. 2. A multiple quarter-wave matching transformer can be used to match two impedances with a large difference in value, or to lower the vswr over the frequency range on which the transformer is to be used.

Now, note where this second line crosses the N scale. It should cross at 2, meaning that a two-section transformer will just satisfy the initial requirements. If desired, a three-section transformer could be used to achieve a lower vswr. Using **fig. 4**, for the two-section transformers, the impedances needed are 74 ohms and 170 ohms. The closest standard values that will work are 73- and 185-ohm cables. Using these values will give a small error, which will increase the vswr above 1.5:1. A three-section transformer (see **fig. 5**) may be more appropriate for lower vswr.

In this example, I've assumed that the antenna presented a constant impedance with frequency, and that the antenna could accept an unbalanced feed without disturbing its radiation-pattern characteristics.

table 1. Characteristic impedances of coaxial cable identified by RG type.

nominal characteristic		velocity
impedance in ohms	RG type	factor
50	RG-9B	0.659
	RG-58A, 58C	0.659
	RG-142, 142A, 142B	0.695
	RG-174	0.659
	RG-178B	0.695
	RG-196A	0.695
	RG-213	0.659
51	RG-9, 9A	0.659
52	RG-8, 8A	0.659
	RG-14A	0.659
	RG-17, 17A	0.659
	RG-18A	0.659
53.5	RG-55B	0.659
	RG-58	0.659
58	RG-54A	0.659
73	RG-59	0.659
75	RG-11, 11A	0.659
	RG-59B	0.659
	RG-140	0.695
	RG-179B	0.695
	RG-187A	0.695
	RG-216	0.659
83	RG-212	0.659
93	RG-62, 62A, 62B	0.84
	RG-71A, 71B	0.84
95	RG-180B	0.695
	RG-195A	0.695
125	RG-63, 63B	0.84
	RG-79B	0.84
185	RG-114, 114A	0.88

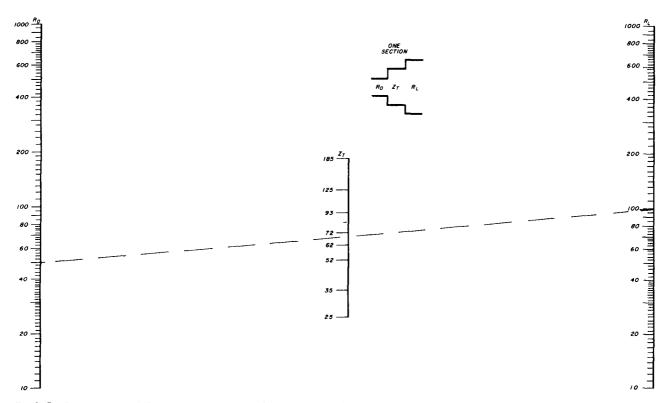


fig. 3. Design nomograph for a one-section transformer. The dashed line illustrates example one in the text. By connecting the points representing the two known values ($R_O = 50$ ohms and $R_L = 100$ ohms) the impedance of the matching section can be determined.

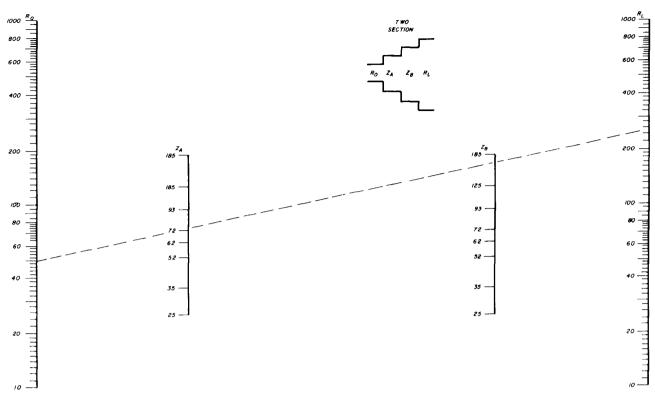


fig. 4. Design nomograph for a two-section transformer. In this illustration, the dashed line represents example two in the text. As in fig. 3, a line is connected between the two values to be matched. The impedance of the two matching sections is read from the nomograph.

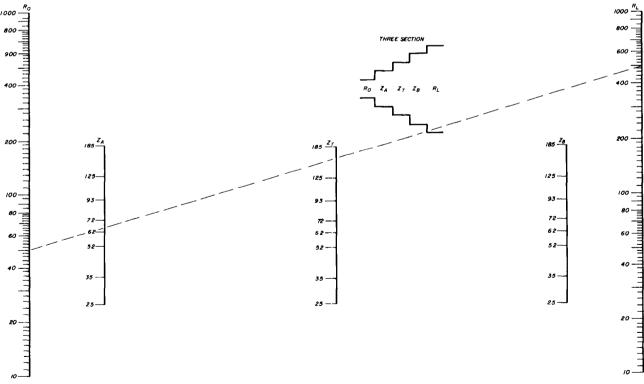


fig. 5. Nomograph for a three-section matching transformer. The impedances necessary to fulfill the bandwidth requirement in example three are illustrated by the dotted line.

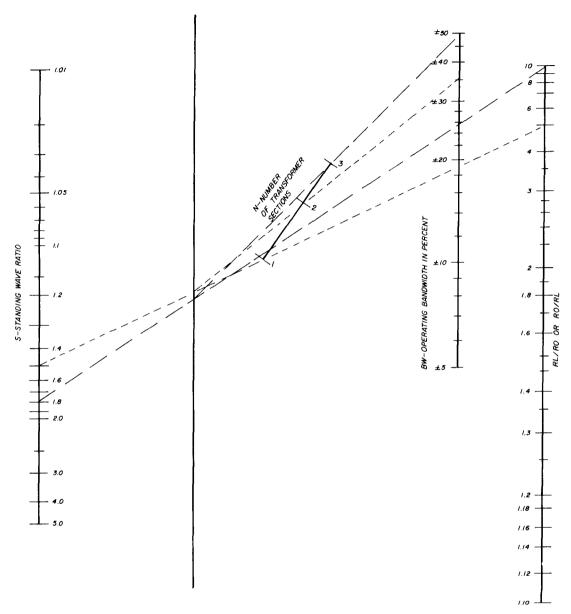


fig. 6. Nomograph for transformer selection. In example two, as represented by the small dashed lines, this figure has been used to determine the number of sections needed for matching two impedances within specific vswr limitations. The large, dashed lines represent example three.

Example 3. What will be the maximum vswr of a three-section transformer used to match a 500-ohm load to 50 ohms over a \pm 50 per cent bandwidth?

Using **fig. 6**, lay a straight edge so that the 50 on the BW scale and 3 on the N scale line up. Draw a line through these two points and extend this line to the unmarked scale. Place a dot where the line crosses the unmarked scale. Now lay the straight edge so that it lines up with the 10 on the R_L/R_O scale and the dot on the unmarked scale. Read off the vswr of 1.8:1 on the S scale. Thus, a three-section transformer has a vswr of 1.8:1 over a \pm 50 per

cent bandwidth with a load resistance ten times the input impedance.

If you have a problem where a quarter-wave transformer can be used, these nomographs will greatly speed up the process of determining the optimum number of transformer sections and impedances of the sections.

reference

1. Samuel Guccione, "Nomograms Speed Design of λ/4 Transformers," *Microwaves*, August, 1975, page 48.

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AM-380

phase-locked 9-MHz BFO

Construction of a
9-MHz BFO system
which can be
phase-locked to a
1-MHz reference standard

The frequency synthesizer system for an ultrastable receiver or transceiver must include a BFO which is phase locked to the external frequency standard. Fig. 1 shows the diagram for a crystal beat-frequency oscillator which can be phase locked to a 1-MHz reference. The BFO delivers a signal exactly 1 kHz below 9 MHz. When driving the detector of a receiver using a narrowband 9-MHz i-f filter, the receiver will deliver an audio pitch of exactly 1000 Hz for an input carrier producing an i-f signal of exactly 9 MHz. Variations on this basic design permit operation at any desired BFO frequency near 9 MHz.

In this circuit Q1 is the voltage-tunable crystal oscillator; Q2 is a power-amplifier/driver stage which produces output suitable for the local-oscillator port

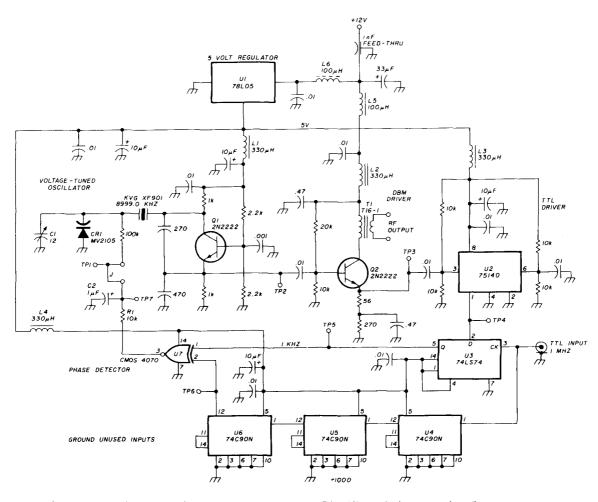
of a double-balanced mixer. Output from the Q2 emitter feeds a 75140 TTL line receiver which supplies the appropriate levels for the D input of the 74LS74 "stripper" circuit. The type-D flip-flop is clocked by the 1-MHz reference signal; its output is a TTL-level squarewave at a frequency equal to the crystal oscillator frequency minus the nearest harmonic of the reference input. Thus, this circuit "strips" off the 9-MHz portion of the oscillator frequency, leaving only the difference frequency at approximately 1 kHz.

A chain of three 74C90 decade dividers delivers a 1-kHz reference to pin 2 of the CMOS 4070 phase detector; the stripper output goes to the other input, pin 1. The lowpass filter R1-C2 reduces the ripple output from the phase detector and delivers the control voltage to the tuning diode CR1.

design approach for low noise

Because this system uses a 1-MHz reference signal, radiation from the reference oscillator must be prevented from reaching the receiver's i-f system. I have found that, with reasonable care, the 9-MHz reference harmonic can be eliminated. I used power-supply filters consisting of L1 through L6 with low value ceramic and $10-\mu F$ tantalum electrolytic capacitors to clean up the power supply system. By specifying CMOS and LS-TTL devices where possible, I keep the noise sources themselves as quiet as possible. All power-supply lines are bypassed to a massive circuit-board ground plane with capacitors having

By Raymond C. Petit, W7GHM, Post Office Box 51, Oak Harbor, Washington 98277



C1 3-12 pF miniature trimmer capacitors
CR1 MV2105 varactor diode (Motorola)

R1 10k nominal; may require adjustment

T1 16:1 broadband rf transformer (Mini-Circuits T16-1)

fig. 1. Circuit for the phase-locked 8.999 MHz BFO with output suitable for driving a double-balanced mixer. Output is +7 dBm into 50 ohms (5 mW). L1 through L4 are rf chokes.

the shortest possible leads. The entire circuit is enclosed in an rf-tight aluminum enclosure with connectors for the reference input, BFO output, and power. The shielded case itself does not carry any rf currents.

assembly and checkout

First build the Q1 and Q2 circuits, including the 5-volt regulator. Apply power and check the regulator output. Then check for about 2 volts at TP2 and about 4 volts at TP3. Place a 50-ohm load on the rf output and check for about 1.5 volts peak-to-peak. Connect a source of 2.5 volts to TP1 and adjust C1 to obtain exactly 8999.000 kHz at the rf output. If you cannot obtain this frequency, check the crystal and increase or decrease the net value of capacitor C1.

Wire the 75140 TTL driver and check TP4 for the desired TTL-level output. Wire the 74LS74 and connect the 1-MHz TTL-level reference signal. TP5

should show about a 1-kHz TTL squarewave which varies in frequency when the voltage at TP1 is varied. Wire the decade divider chain and look for a 1-kHz squarewave at TP6. With the jumper J removed from between TP1 and TP7, observe the waveform at TP7 while varying the voltage at TP1. With the TP1 voltage at 2.5 volts, TP7 should show a very low frequency triangle wave going from zero to 5 volts. As the TP1 voltage is raised or lowered, this waveform should pass through zero beat, rise in frequency, and diminish in amplitude.

Wire in the jumper J and remove the voltage source from TP1. The junction TP1-TP7 should now show a 2.5-volt steady dc level. When C1 is varied slightly, this dc level should rise or fall slightly. If C1 is moved too far, TP1 will suddenly show a low-frequency triangle wave. By moving C1 back, this should suddenly "capture" and again show a steady level. Adjust C1 until this level is again at 2.5 volts.

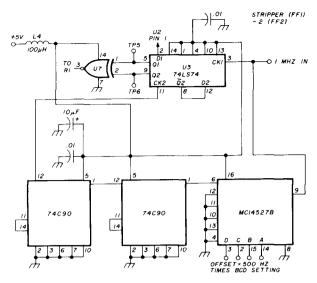


fig. 2. Schematic of a rate multiplier offset synthesizer for the phase-locked BFO.

modifications for other frequencies

If the desired BFO frequency is 9001.000 kHz, substitute a crystal of this frequency. No other modifications are required. For single-sideband work, where the BFO frequency will be plus or minus 1500 Hz from

9 MHz, use the appropriate crystal and substitute the rate-multiplier circuit of fig. 2. With this circuit you can select any frequency offset from 500 Hz to 4500 Hz. The MC14527B rate multiplier accepts a 1-MHz input from the reference oscillator and generates unevenly spaced output pulses, the average rate of which can be programmed with the four rate input settings D, C, B, and A. The rate input is in BCD, positive logic. If the setting is 3 (BCD 0011), for every 10 input pulses from the reference there will be three output pulses. The two 74C90 decade counters and the second half of the 74LS74 serve as a divide-by-200. With this arrangement a rate setting of three produces an average frequency output of the rate multiplier of 300 kHz, and the divider brings this to 1500 Hz.

The next article will describe the phase-locked upconverter which translates the 1100 to 1600 kHz output of the first loop¹ to the required local-oscillator frequencies for each of the amateur bands from 160 meters through 30 MHz.

reference

1. Raymond C. Petit, W7GHM, "Synthesized High-Frequency Local-Oscillator System," ham radio, October 1978, page 60.

ham radio



magnetic mount

for mobile antennas

A novel antenna mount made from a cast-off surgical device called a *Bio-Pump*

It's not often you run across the perfect piece of junk. In this case, it's a magnet with all the qualities you'd want in a magnetic mount for an antenna:

- 1. Holding power
- 2. Shape
- 3. Plastic case
- 4. Coax-fitting-sized hole

The device that houses this magnet isn't piled up at your local surplus store; in fact it won't be there at all. Fortunately, however, it's available to those wishing to seek one out. More on this later.

My son Brian had a strange looking contraption among his collection of junk, and I paid little attention to it for months. The thing finally aroused my curiosity one day when it started collecting nuts and bolts on its own, so I attempted to figure out its use and purpose. The marks on its side said *Bio-Pump*, whatever that meant.

It looked expensively made and had a magnet buried deep inside. Since it was headed for the scrap pile anyway, I struck it with a hammer and ended up with a magnet encased in plastic. And that's the heart of this story.

development of an idea

I didn't immediately visualize the gadget as an antenna mount. It took a while since I didn't need one at the time. Holes for an antenna mount had been recently put into my International Scout, justified by the fact that it was the only mount I had without buying one. I also needed strong support for the colinear antenna that would be used.

When I finally got around to the *Bio-Pump*'s mounting potential I checked out its holding power. Using a metal file cabinet, I wasn't prepared for how it was grabbed out of my hands by the magnetic attraction.

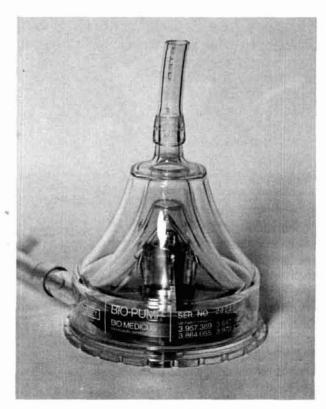
Since my location is near metropolitan Minneapolis-St. Paul, and normal driving takes me near the repeater I use often, I decided to build a quarter-wave whip for the mobile rig. Result: I put the colinear into permanent storage. Now I don't whack tree limbs and can drive into my garage.

how to get one

It turns out that this pump is a disposable item after being used during open-heart surgery. In the Twin Cities alone, 1500 to 2000 open-heart operations are performed annually. Not all of them use this particular pump; however, many do, and its use is increasing.

I can't supply direct contacts for you to obtain

By W. H. Kelley, WØHK, Route 1, Box 295A, Maple Plain, Minnesota 55359



The Bio-Pump, which contains a healthy magnet that can be used to mount a mobile antenna. These devices are expendable after heart-surgery operations and can be obtained free from many hospitals.

one, but knowing someone who works in a hospital is how I've obtained three of them for this article. Since they're disposable, I believe a few phone calls will put you in touch with the right people. If they're reluctant to give you one and prefer instead to throw it away or destroy it, offer to destroy it for them in their presence just for the encased magnet. In any case, be open about who you are and what you want it for.

I'm not usually in favor of articles in which an exotic hard-to-obtain component is required, but obtaining this pump is the heart (pun intended) of this design. Finding one will be worth your effort.

description and use

The plastic covering is perfect for not marring the car surface. Construction is simple, as you will see.

Amateurs have a proclivity for making do. The name of the game is innovation; if you don't have it, you make it. This article shows what one amateur operator did with a cast-off piece of junk. The result is a mag mount for a mobile antenna that sticks to metal like a burn to a ham sandwich. Does your mobile whip hit the dirt when a big semi tractor-trailer rig goes by? Try this mount and forget your worries. Editor.

It's shaped for minimum wind resistance and the holding power is unbelievable.

To accent that last point, I went on a fishing trip where four-wheel drive was periodically required. You'd probably be impressed if I told you that the antenna stayed on top without moving. Better than that, I moved it inside the vehicle on a vertical panel and it did not move in the slightest for several days.

When you get one, don't use a hammer as indiscriminately as I did, especially if you only have one plastic pump. When breaking the outer plastic case, start at the opposite end of the magnet. The plastic that houses the magnet is the fourth wall in as you look inside. The second and third walls are actually one piece and do a good job of protecting the magnet as you break through the outer wall.

The plastic chips very easily, so protect your eyes and don't go barefoot until you clean up. Once you remove enough plastic to expose what you're after, use an allen wrench to remove the screw, washer, and seal from where it's held on the bottom. The photo shows how.

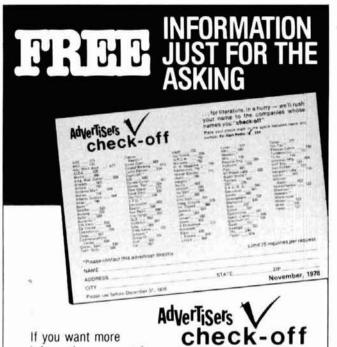
You'll see a couple of strips of something wandering around inside and the inclination is to figure out a way to remove them. Don't! I tried and it's impossible without doing damage.

A hacksaw can be used to cut the pointed plastic end of the inner container. Cut just above the inside metal cylinder top surface. You'll find that the plastic is molded around this corner, leaving about 0.8 mm (1/32 inch) lipped over the top tends to seal at that

Both ends of the rotating works in the center will now be exposed and can be disassembled. Don't



The Bio-Pump after applying a hammer to break the plastic outer shell, left. A complete instrument before destruction is shown at center. At right is the inner plastic container that houses the magnet.



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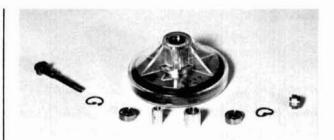
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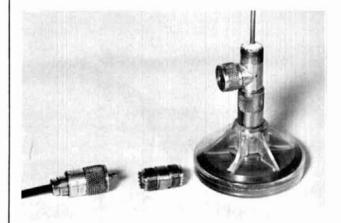
Inner part of the Bio-Pump with pointed end sawed off.

drop the pump on a hard surface, as it will be damaged.

Disassemble the parts in the same order as shown in the photo. The two snap rings can be bent and pulled out with a sharp-pointed object. Now is the time to set the magnet onto a metal surface and be impressed.

The following description shows how I used this magnet. It was chosen more because of parts on hand than anything else. (See photo.) The rear of a PL-259 coax fitting will be a loose fit. Some solder on the fitting, filed down to a press fit, will do the job.

If you have some no. 277 *Loctite* as I did, it will also work. A tee coax fitting allows a point to feed the antenna.



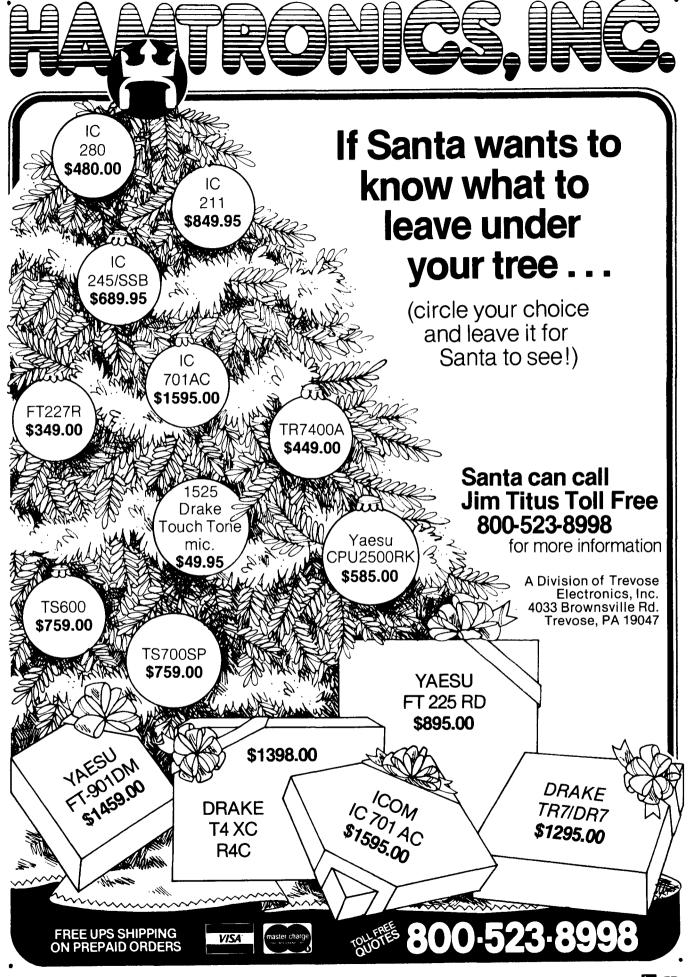
Complete mag mount showing coax Tee fitting and coaxcable accessories. Antenna is a quarter-wavelength whip made from 3-mm (1/8-inch) diameter brass welding rod.

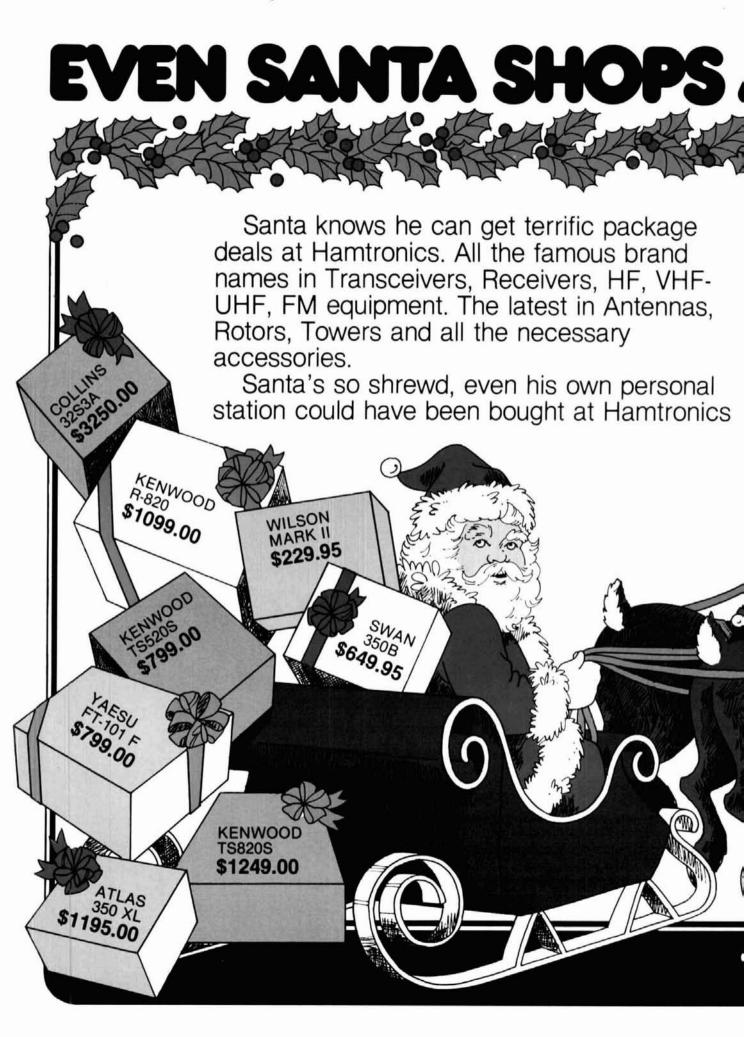
For a whip, I used 3 mm (1/8 inch) diameter brass welding rod. Put some sealant at the whip base for a weather seal after soldering the rod into the fitting.

acknowledgment

Special thanks to Neil Gravatt for taking the pictures.

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Digiratt 1K —

a digital reperf/TD

Continuing in the Digiratt series, KB9AT has developed a 1024-character digital RTTY reperf/TD Most ideas become reality because someone defines a need and develops an approach to meet that need. The original Digiratt RTTY AFSK Generator and PLL Demodulator1 was designed so I could simply and rapidly place a vhf RTTY station on the air with few frills. The Digiratt RY Test Generator2 came into being because of a badly misadjusted model 15 and the resulting need for a stable test signal. When a model 28 KSR arrived. I found that a model 28 ASR would be more desirable than the standard KSR. Therefore, of necessity, a digital replacement for the usual mechanical device was designed.

design concepts

I initially decided to design a unit which was the equivalent of a Reperf/TD, which is to say it could copy RTTY off the air as well as from the local keyboard. Also, since all my RTTY equipment was TTL based, it followed that the new unit would also be TTL. It had to have the ability to edit mistakes without re-entering the entire message and it also had to have a memory capability of a minimum of 16 lines of 64 characters. I decided to include a recirculator so that all, or a portion, of the memory could automatically be repeated over and over for calling CQ, printing R/Y's, and doing "Quick Brown Fox" type testing.

By John Loughmiller, KB9AT, RR1, Box 480C, Borden, Indiana 47106

The resulting Digiratt 1K allows the user to write (enter) up to 1024 RTTY characters into the memory, either from a local keyboard or off the air (see **fig. 1**). Entry speed can be as slow as desired so long as the Baud rate is 45.45 Hz. Full 60 wpm entry speed is allowed, with other speeds accommodated by changing the system clock frequency.

In the read mode, the data is clocked out at a constant rate approaching 60 wpm. Either 128 characters or all 1024 characters can be recirculated. (This could be 2 lines of 64 characters or 16 lines of 64 characters.) To change or delete a character, the memory is allowed to read (print) up to and including the character immediately preceding the one to be changed or deleted. The unit is then switched to the write mode and the correction is made. Read is re-established and the next correction made in a like manner.

theory of operation

Clock generator. The clock generator board* is composed of a 1-MHz oscillator followed by a divide-by-1375 circuit which furnishes an output of 727.2 Hz (fig. 2). This frequency is 16 times the 45.45 Hz Baud rate used for 60 wpm RTTY and is also the frequency required by the UAR/T for input data processing.

Data entry. The 727.2-Hz clock signal, in addition to being applied to the UAR/T, is also divided by 16 (U1 and U2), becoming the shift register clock. It is further divided by 8 (U3) and used as the shift register load/read pulse (see **fig. 3**).

Serial RTTY data enters the board and is applied to the UAR/T. When the character has been processed and is ready to be transferred to memory, a 5-bit parallel data word appears on pins 9 through 12 of the UAR/T, and the received data available flag changes

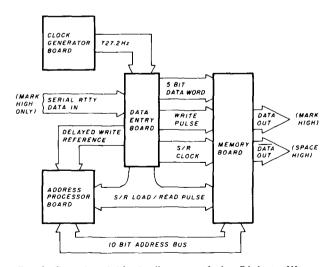


fig. 1. Functional block diagram of the Digiratt 1K reperf/TD.

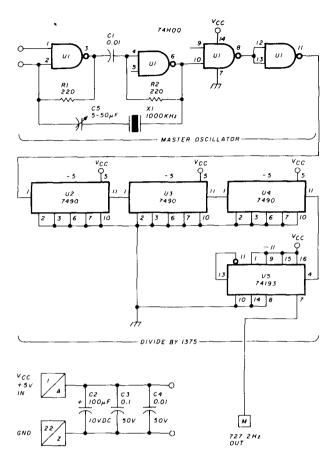


fig. 2. Schematic diagram of the clock and divider board. A basic 1-MHz oscillator is divided by 1375 to produce the 727.2-Hz waveform.

state. This flag is then used to fire U4. The output from U4 triggers U5, and also becomes the write pulse to load data into memory. U5's output is called delayed write reference and is used to decrement address processor one number and simultaneously clear the UAR/T for reception of the next character.

Address Processor. The three binary counters (U1 through U3 in fig. 4) receive either the delayed write reference or the shift register load/read pulses as a clock input, depending on whether the write or read mode is selected. When the load button is pressed, a predetermined address is loaded into these counters, either 1024₂ or 128₂ depending upon the position of the full/partial switch. This allows the user to instantly return to a predetermined point. If the *Recirc* switch (recirculator) is on, the predetermined

^{*}A copy of the printed circuit board layout is available by sending a self-addressed, stamped envelope to *ham radio*, Greenville, New Hampshire 03048. In addition, a complete set of etched, drilled, and plated circuit boards is available for \$17.50 from Circuit Board Specialists, P.O. Box 969, Pueblo, Colorado 81002. A complete parts kit is also available for \$55.00.

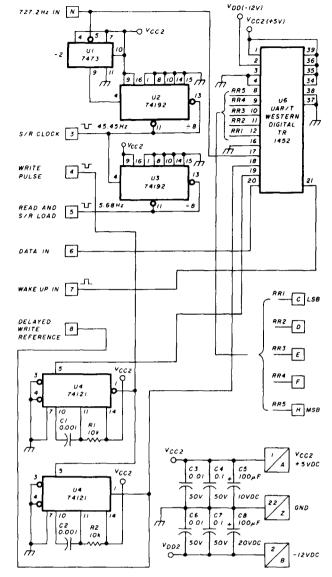


fig. 3. The UAR/T on the data entry board is used to convert the serial RTTY data to a 5-bit parallel form. U1 through U3 provide additional division of the 727.2-Hz signal to provide the correct frequency for the output shift register.

address will constantly be reloaded at the end of the countdown cycle, resulting in recirculation of either 128 or 1024 characters, indefinitely. The outputs from the three 74193s is the binary address used to control the 2102 RAMs during both read and write modes.

Memory. In all modes, the binary address is applied to all five RAMs, one bit per RAM (**fig. 5**). The write pulse enters the data into memory. The delayed write reference pulse, which follows, increments the address processor to the next lower address. This procedure then repeats for subsequent characters.

In read mode, the shift register-load/read pulse both decrements the address processor and loads the data from the RAMs into the 74165 shift register (U6). Finally, the clock pulse into pin 2 of U6 shifts out the serial data at a 45.45 Hz Baud rate.

Power supply. The power supply, as shown in **fig. 6**, is quite straightforward, using the popular LM309K regulators. I do suggest that the reader not deviate from the schematic with respect to the use of separate regulators. Also, bypassing on the individual boards should not be omitted.

With the power supply and on-board decoupling shown, I've placed the unit on top of an operating linear power amplifier, with no problems resulting from the rf field. In extreme cases, however, you may have to bypass the RAMs with 0.01-µF capacitors at their V_{cc} pins. For that matter, any of the ICs may need bypassing if erratic operation is experienced in a high-intensity rf field.

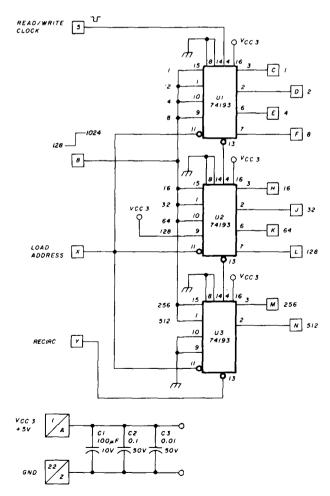


fig. 4. The address processor generates the 10-bit binary address for the RAMs. The 74193s count down from the number loaded into their pre-set inputs.

initial start-up

Apply power and press the clear button to reset the UAR/T registers. This must be done each time the system is turned on, or you can build in a power-up system using a resistor, capacitor, and inverter. In any event, the UAR/T registers must be reset at turn on.

Select full address, write, run, no recirc, and press the load button. Enter 1024 Blanks and select run and read. Press load address once and observe the playback. If you get anything other than blanks on playback you probably have a defective RAM. Consult a baudot code chart and determine which RAM would have to be defective to cause the resultant printed letter; the RAMs correspond directly to the baudot chart. Next, load *Ltrs* into all 1024 spaces and repeat the procedure outlined above. If you can store 1024

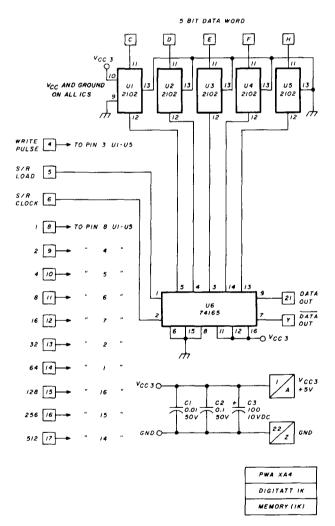


fig. 5. The memory board contains five 2102 RAMs and the output shift register. The 10-bit address lines are simultaneously applied to all 2102s. The hardwired inputs of the shift register are used to generate the stop and start pulses.

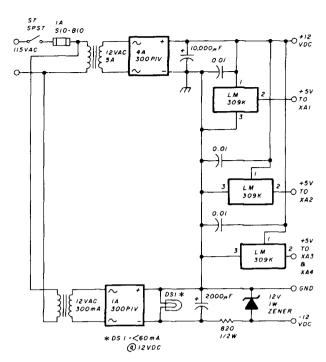


fig. 6. Schematic of the power supply for the Digiratt 1K. The small indicator lamp should draw less than 60 mA at 12 volts. The three LM309K regulators are mounted on the back of the case.

blanks and 1024 *Ltrs* and recover them without error, the RAMs are probably acceptable.*

Obviously, you'll need a baudot code chart (found in most RTTY books) to trouble-shoot RAM problems. I bought eight "prime" 2102 ICs, and three were bad at certain addresses. It was a blessing in disguise, however, since this trouble-shooting procedure was the result of that problem.

system interface

The UAR/T requires the input mark signal to be a TTL high. Therefore, any system devised to enter data will have to furnish a high logic level when the loop is in the mark condition. With respect to outputs from the Digiratt 1K, both mark and space signals are available with TTL high levels as their true conditions.

One suggested input circuit is the system devised by John Alford, WA4VOS, 2 using a diode bridge and optical isolator. Another method would be to key a reed relay from the local loop. Connect the relay contacts in such a manner that the center arm is connected to 5 Vdc when the loop is in the mark state and ground when the loop is in space condition. You may find it necessary to debounce the contacts of the reed relay.

*It should be noted, however, that special memory testing routines are required to disclose some failures. This procedure is *not* a complete test but should suffice in most cases.

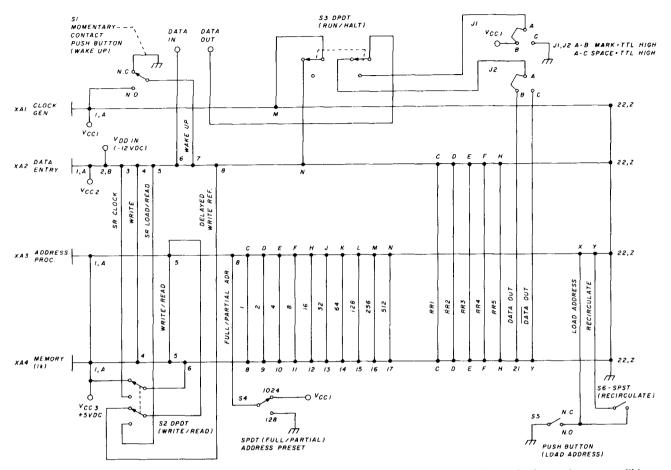


fig. 7. Interconnection wiring diagram for the Digiratt 1K. With jumpers J1 and J2 connected to A-B, the mark output will be a TTL high level; connected to A-C the mark level will be a low level.

operation

- Write. 1. Select either full or partial memory.
 - 2. Press load address.
 - **3.** Turn the recirculator off (even if it will be used during read).
 - 4. Set the write/read switch to write.
 - 5. Set run/halt to run.
 - 6. Enter the data.

If the recirculator is to be used, fill in all unused memory slots with blanks, or random data will be printed until reset occurs.

- Read. 1. Set full/part switch as required.
 - 2. Press load address.
 - 3. Set write/read to read.
 - 4. Set recirc as desired.
 - 5. Set run/halt to run when ready to read out the data.

future expansion

I've designed a couple of options which are not on the prototype unit. The first is a one-step forward/backward text editor which allows correction of mistakes. Second, an address preset using BCD thumbwheels and 7-segment displays, making the memory completely random access. You merely dial up the place where a particular message is stored, press the load address button, and read out the contents. Finally, a 2K version and an ASCII version are also designed but not yet built.

references

- 1. John Loughmiller, WB9ATW, "Digiratt AFSK Generator and Phase Locked Loop Terminal Unit," ham radio, September, 1977, page 26.
- 2. John Loughmiller, "RTTY Text Generator," ham radio, January, 1978, page 64.
- 3. John Alford, WA4VOS, "Improved Digital AFSK," ham radio, March, 1977, page 22.
- 4. Johnathan A. Titus, "The UAR/T and How It Works," ham radio, February, 1976, page 58.

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DESCRIPTION: The CT-50 is a versatile and precision frequency counter which will measure frequencies to 60 mHz and up to 600 mHz with the CT-600 option. Large Scale Integration, CMOS circuitry and solid state display technology have enabled this counter to match performance found in units selling for over three times as much. Low power consumption (typically 300-400 ma) makes the CT-50 ideal for portable battery operation. Features of the CT-50 include: large 8 digit LED display, RF shielded all metal case, easy pushbutton operation, automatic decimal point, fully socketed IC chips and input protection to 50 volts to insure against accidental burnout or overload. And, the best feature of all is the easy assembly. Clear, step by step instructions guide you to a finished unit you can rely on. Use the order blank below or call us direct and order yours today!

SPECIFICATIONS:

Frequency range: 5 Hz to 65 mHz, 600 mHz with CT-600 Resolution: 10 Hz @ 0.1 sec gate, 1 Hz @ 1 sec gate Readout: 8 digit, 0.4" high LED, direct readout in mHz

Accuracy: adjustable to 0.5 ppm

Stability: 2.0 ppm over 10° to 40° C, temperature compensated Input: BNC, 1 megohm/20 pf direct, 50 ohm with CT-600

Overload: 50VAC maximum, all modes

Sensitivity: less than 25 my to 65 mHz, 50-150 my to 600 mHz

Power: 110 VAC 5 Watts or 12 VDC @ 400 ma Size: 6" x 4" x2", high quality aluminum case, 2 lbs

ICS: 13 units, all socketed

CT-600: 600 mHz prescaler option, fits inside CT-50

CB-1; Color burst adapter, use with color TV for extreme accuracy and stability, typically 0.001 ppm

OPTIONS:

CB-1 option: The CT-50 time base may be locked to an external frequency standard. The television networks maintain extremely accurate atomic based frequency standards to maintain color tint on TV programs. These standards are typically accurate to one part in 10 to the 12. By locking the CT-50 to one of these network standards, we are able to get super accuracy. The CB-1 adapter interfaces a standard color TV receiver to the CT-50 so that one can take advantage of the TV network frequency standards. The CB-1 requires connection to a color television for operation.

CT-600 option: The CT-600 prescaler option enables the CT-50 counter to measure frequencies as high as 600 mHz with sensitivity in the 20 to 150 mv range, depending upon frequency. Typical sensitivity at 150 mHz is 25 mv. The CT-600 mounts on the same PC board as the CT-50, no extra boxes or PC boards are required. The scaler utilizes a state of the art ECL IC chip and two transistor pre-amplifier, thus eliminating the need for external pre-amp devices.

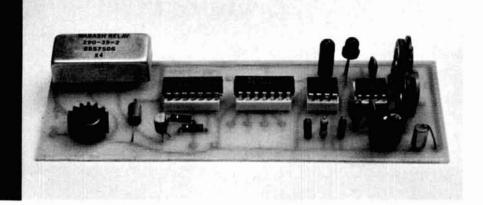
ramsey electronics

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CT-50, 60 mHz Counter Kit	\$89.95
CT-50 WT, 60 mHz counter, wired, tested CT-600, 600 mHz prescaler option	159.95
for CT-50, add	29.95
ACCESSORIES	
DC probe, direct input, general purpose type	\$12.95
High impedance probe, does not load circuit	15.95
Low pass probe, used when measuring audio	15.95
High pass probe, reduces low freq pickup	15.95
VHF flexible rubber antenna, BNC connector	12.95
Color burst adapter, for calibration, high accuracy typically 0,001 ppm accuracy, stability	14.95

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City	State	Zip



tone-alert decoder

Construction of a simple tone-alert decoder for emergency call-ups of RACES groups and the Amateur Radio **Emergency Service**

How do you monitor for a possible call-up of the local ARES or RACES group on the local repeater while you're at the office — without subjecting other office employees to the normal traffic on the repeater? How do you monitor for emergency calls throughout the night without being jolted out of bed by an amateur who works the sign-off shift at the TV station and puts out a call on his way home from work? One solution is a telephone call-up using the telephone tree system. The problem is that if one or two members are not home, it can really slow down or even stop the fan-out process.

After considering and experimenting with several alerting systems, the Wayne County (Michigan) Amateur Radio Public Service Corps (ARPSC) developed an alerting system over the local repeater that is based upon a tone-alert. The basic concept of the system is that any ARPSC member equipped with the tone-alert decoder who wishes to be notified of an alert - but does not want to listen to the repeater audio - can switch the tone-alert decoder to monitor mode. This allows the audio from the receiver to pass to the decoder but does not allow it to pass on to the speaker.

To activate the tone-alert the amateur calling the alert transmits the appropriate audio tone for the proper duration; the decoder latches up and the relay contacts switch the audio onto the speaker.

So far the tone-alert system has been used for several tornado watch "sky warn" net call-ups in Wayne county, Washtenaw, Monroe, and Genisee County groups have also started the development of tonealert systems using the circuit described in this article.

There were four primary objectives of the decoder design:

- 1. A decoder which would not give false alerts by detecting voice peaks or alternator whine.
- 2. A decoder which would cost less than \$15, because the higher the cost, the fewer that would be placed into service by volunteers.
- 3. A decoder which was built from readily available parts; this would increase the number which would be built by volunteers.
- 4. A decoder circuit which could be reset by the appropriate signal.

After the criteria were established, a search of the literature was made and no simple circuit was found that met our objectives. (After our system was developed a decoder meeting some of the design requirements was published in QST.)

A prototype was developed and evaluated for several months on a receiver monitoring the local repeater. Tests were made of various call-up systems and tone encoders. When the design was considered satisfactory a batch of circuit boards was produced and kits of parts made available to ARPSC members who wanted to participate in the program. The record of performance is good for the units now in service.

By Harold C. Nowland, W8ZXH, and Stan Briggs, W8MPD, Hal-Tronix, Post Office Box 1101, Southgate, Michigan 48195.

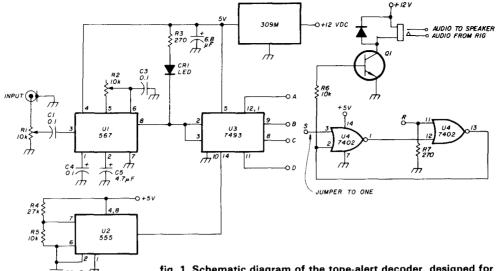


fig. 1. Schematic diagram of the tone-alert decoder, designed for the call-up of amateur radio volunteers during emergencies.

circuit description

Integrated circuit U1 is a NE567 PLL tone-decoder timed to the system alert tone frequency (see **fig. 1**). The audio from the receiver passes through the level control R1 to the 567 input. R2 and C2 control the 567's operating frequency. When the proper tone is not being received the 567 output is about +4 volts, which is applied to the reset input of U3. U2 is an oscillator operating at about 1 Hz. The output is fed to the clock input of the 4-bit counter, U3; as long as the reset is held high by the output of U1, however, the counter stops at zero.

When the proper tone frequency is received at the input of U1 the output voltage goes to zero. This turns the LED CR1 on, indicating that the PLL has locked up. This also allows the reset input to the counter to go low, allowing the counter to start to count. The outputs of the counter will go high in turn as the counter counts up: (1) high after the first pulse, (2) high after the second pulse, (3) high after four pulses, (4) high after eight pulses.

The desired delay is obtained by selecting the output to be jumpered to the SET input to the latch, U4. If a delay of four clock pulses (about four seconds) is desired, for example, a jumper would be placed between C and S. Once the count reaches four, the output of the latch will turn on the relay driver transistor Q1.

If the tone is not on for the whole four clock pulses, the counter will reset when the PLL output goes high. This protects from lock-up on voice peaks, alternator whine, or other spurious signals.

The relay contacts may be placed in series with the

external speaker circuit so that audio can be heard when the relay is operated.

To reset the latch a positive voltage must be applied briefly to the R input. The circuit can also be reset remotely by using a diode from the D output to the R input. This allows the tone of eight clock pulse duration to reset the decoders remotely if desired. The format is thus: ON, four clock pulses; OFF, eight clock pulses.

construction

The layout of the circuit is not very critical. The only critical component is the 0.1- μ F capacitor (C3). This should be a good quality Mylar capacitor to limit frequency drift.*

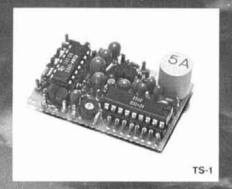
The tune-up procedure is as follows:

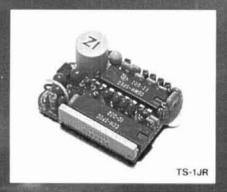
- 1. Set R1 and R2 at the center of their ranges.
- 2. Apply the desired tone frequency to the decoder input. R2 should be adjusted until the LED lights, indicating the lock-up of the PLL.
- 3. The signal level should be reduced and R2 should be adjusted to locate its "center of lock."
- **4.** On-the-air checks should be made to determine the appropriate level setting of R1.

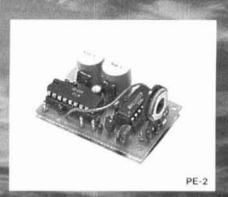
Various types of tone generators may be used. The Wayne County ARPSC group uses a quartz crystal oscillator which is counted down to the proper audio frequency by a 14-stage ripple-carry binary counter/divider integrated circuit such as the CD4020. Other groups have found that simple oscillators using 555 timers with good components and voltage regulation are adequate.

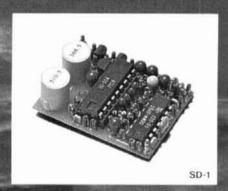
ham radio

^{*}A parts kit, less relay, and an etched and drilled circuit board are available for \$10.95 plus \$1.00 postage from Hal-Tronix, Box 1101, Southgate, Michigan 48195.





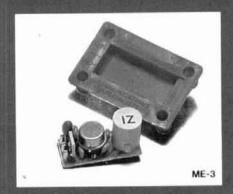


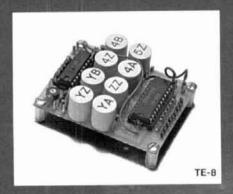


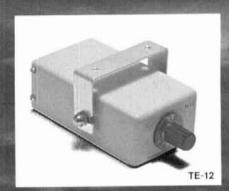
THE DAWNING

The age of tone control has come to Amateur Radio. What better way to utilize our ever diminishing resource of frequency spectrum? Sub-audible tone control allows several repeaters to share the same channel with minimal geographic separation. It allows protection from intermod and interference for repeaters, remote base stations, and autopatches. It even allows silent monitoring of our crowded simplex channels.

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TS-1 Sub-Audible Encoder-Decoder • Microminiature in size, 1.25" x 2.0" x .65" • Encodes and decodes simultaneously • \$59.95 complete with K-1 element.

TS-1JR Sub-Audible Encoder-Decoder • Microminiature version of the TS-1 measuring just 1.0" x 1.25" x .65", for handheld units • **\$79.95** complete with K-1 element.

ME-3 Sub-Audible Encoder • Microminiature in size, measures .45" x 1.1" x .6" • Instant start-up • \$29.95 complete with K-1 element.

TE-8 Eight-Tone Sub-Audible Encoder • Measures 2.6" x 2.0" x .7" • Frequency selection made by either a pull to ground or to supply • \$69.95 with 8 K-1 elements.

PE-2 Two-Tone Sequential Encoder for paging • Two call unit • Measures 1.25" x 2.0" x .65" • \$49.95 with 2K-2 elements.

SD-1 Two-Tone Sequential Decoder • Frequency range is 268.5 - 2109.4 Hz • Measures 1.2" x 1.67" x .65" • Momentary output for horn relay, latched output for call light and receiver muting built-in • \$59.95 with 2 K-2 elements.

TE-12 Twelve-Tone Sub-Audible or Burst-Tone Encoder • Frequency range is 67.0-263.0 Hz sub-audible or 1650-4200 Hz burst-tone • Measures 4.25" x 2.5" x 1.5" • \$79.95 with 12 K-1 elements.

ST-1 Burst-Tone Encoder • Measures .95" x .5" x .5" plus K-1 measurements • Frequency range is 1650 - 4200 Hz • \$29.95 with K-1 element.

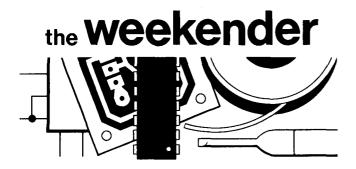


COMMUNICATIONS SPECIALISTS

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V/SA'



an antenna swr meter

In keeping with the Weekender theme, here is a useful station accessory that can be constructed in the course of a weekend and should provide many vears of service.

The antenna meter (or vswr bridge, as this type of device is usually called) provides an indication of match or degree of mismatch between the transmitter and antenna system. I don't like to refer to the unit as a bridge, because that implies a device of extreme accuracy and brings to mind instruments built by General Radio and the like. Rather, the antenna meter is a simple instrument that samples the forward (incident) and reflected voltages in the coaxial feedline between the transmitter and antenna system. This easily built device is useful for tuning your transmitter for maximum output, adjusting an antenna tuner for minimum reflected voltage, and tuning your antenna system to a favorite spot on the band.

This antenna meter is for use with 50-ohm coaxial lines, such as RG-8/U and RG-58/U. It covers the high-frequency bands and will operate with any type rig, other than QRP. The unit contains no exotic or hard-to-find parts, with construction greatly simplified through the use of a printed-circuit board. If you do not have the facilities for etching the PC board, an etched and drilled board is available as shown in the parts list. This form of construction eliminates the mechanical difficulties often associated with instruments of this type. The sensitivity of the meter varies with frequency, about 60 watts on 80 meters and approximately 3 watts on 10 meters for full-scale deflection.

circuit description

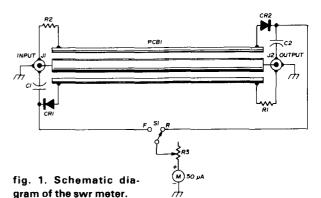
The circuit of the antenna meter shown in fig. 1 has been around a long time. It has been constructed in many forms, ranging from wires snaked inside

By Ken Powell, WB6AFT, 6949 Lenwood Way, San Jose, California 95120

coaxial cable to copper tubing spaced inside a minibox. Although construction techniques vary, the basic theory and function remain the same. In this configuration, the signal from the transmitter is applied to the input jack, J1, passes along the center conductor of the etched board, and exits via J2.

Two sense lines are etched parallel to the center conductor. The lower line along with its associated diode detector, is called the incident, or forward, line; the upper line is called the reflected, or reverse, line. The sense lines are identical, but are connected to detect voltage in opposing directions.

A forward/reflected switch connects a metering circuit which reads the voltage developed in the sense lines as the rf signal passes along the surface of the center conductor. By using the calibration control, R3, the reading from the forward sense line can be adjusted to deflect the meter to full scale. Then, through the use of the forward/reflected switch, the reflected voltage can be read to develop forward-to-reverse voltage ratio.



S1	rocker switch, dpdt	RS 275-030
J1, J2	single-hole SO-239 connector	RS 278-195
CR1, CR2	germanium diodes 1N34A	RS 276-1123
C1, C2	.005/500 V dc disc caps	RS 272-130
R1, R2	51-ohm 5 per cent resistors	RS 271-1308
R3	100k-ohm pot	RS 271-092
M1	50 μA meter	RS 22-051

The reflected reading represents the portion of the voltage that has been applied to the feedline but has been reflected back from the antenna system due to mismatch. The primary objective is to maintain the reflected reading as low as possible, with a reading near zero indicating a purely resistive 50-ohm termination at the far end of the coaxial feedline; that is, a good match between the transmitter and the antenna system.

construction

The heart of the antenna meter is the printed-circuit board shown in fig. 2. The board is double-sided glass epoxy with a thickness of 2.5 mm (3/32 inch).

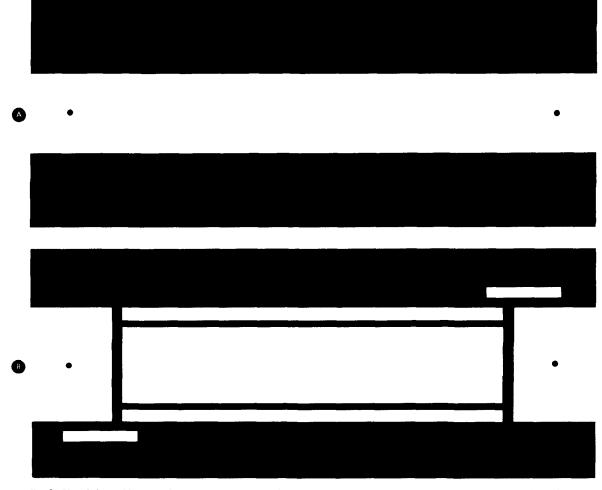


fig. 2. Circuit board layouts for the board used in the swr meter. The top side is shown at (A); the back of the board at (B). Contrary to normal ham radio style, the foil is represented by the white areas.

The only holes in the board are the 16-mm (5/8-inch) holes for the coax connectors. After etching and drilling are completed, place a piece of masking tape on the reverse side of the board (fig. 3). Lay the PC board down, component side up, and tin the areas shown in fig. 3 with a small iron and rosin-core solder. This pre-tin operation will aid in component mounting.

Form the resistors, diodes, and capacitors as depicted in **fig. 4**, and, using needle nose pliers to grasp the lead being soldered, mount the diodes, resistors, and capacitors as shown in **fig. 5**. Solder the output leads from both sense lines and leave them about 30 cm (12 inches) long. Next, mount the single-hole SO-239 UHF connectors to the circuit board, but *do not* over-tighten the hardware. Using no. 14 (1.6 mm) AWG solid copper wire, form two pieces of wire, as shown in **fig. 4**, to connect the center leads of the UHF connectors to the printed circuit board. Solder the wires to the UHF connectors first, then align the wires with the center of the board and solder in place, being sure the wires do not touch

the grounded frame of the connectors. This completes the circuit board wiring.*

Drill the required holes in the case, being careful not to damage the paint in the process. After making sure that all components fit correctly, wash the case assembly with a mild detergent to remove any wax or oils which may be present. Prior to mounting the components, apply the lettering to the case using rub-on letters (I used the Datak K61 Letter Set). Then apply a couple of light coats of Krylon Clear or Datak Matte Finish to protect the lettering and finish. This will give a professional look to your project.

Remove the lock nuts and washers from the UHF connectors and mount the circuit board and other components in the case. Wire the unit as shown in the schematic diagram, using the meter's negative terminal as the common ground point for the ground wires from both sense lines. Parts placement is not

^{*}An etched printed circuit board is available from J. Oswald, 1436 Gerhardt Avenue, San Jose, California 95125 for \$4.00 postpaid. Order board number 1004J.

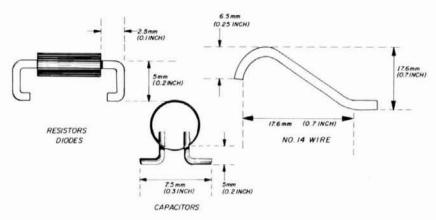


fig. 4. Component preparation prior to mounting on the circuit board.

critical, but try to keep the wires from the sense lines perpendicular to the board for a distance of a few centimeters. After assembling the cabinet, remove the plastic snap-on cover from the meter and remove the meter face. Using a common eraser, remove all the numerals from the meter face, exercising caution not to damage the scale graduations. Now, apply the new digits (see **fig. 6**) using the lettering set and

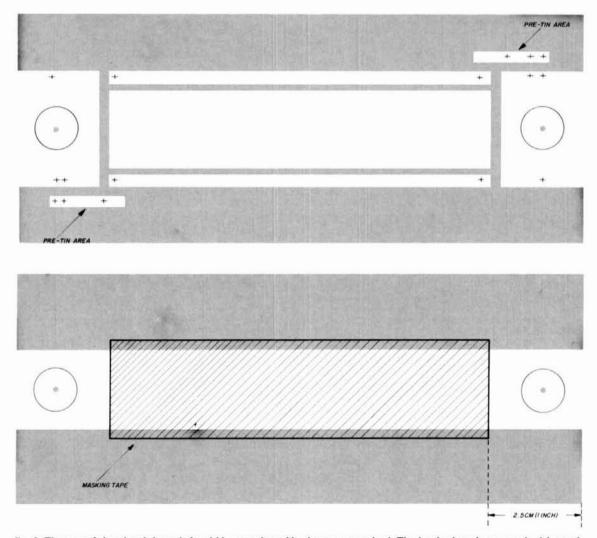


fig. 3. The top of the circuit board should be pre-tinned in the areas marked. The back plane is covered with masking tape as indicated.

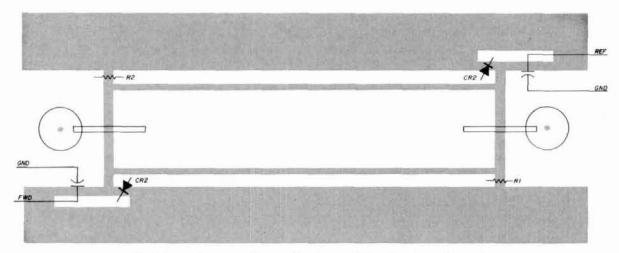


fig. 5. Parts placement diagram for the top of the printed circuit board.

finish the meter face with a light coat of clear finish. This will protect the face and hide the erasure marks. After the finish has dried, assemble the meter and mount the calibration knob.

testing and use

To test the completed unit, connect the transmitter to J1 and the antenna feedline to J2. Set the forward/reflected switch to the forward position, key the transmitter, and adjust the calibration control for a full-scale meter reading. Now, flip the switch to the reflected position and note the reading on the meter. Next, transpose the antenna and transmitter leads, key the rig, and adjust the calibration control to full scale again. Flip the switch to the forward position, and a reading close to that noted in the first test should be observed. This is due to the fact that the sense lines are the same. Transpose the antenna and transmitter leads again and the unit will be restored to normal operation.

The first antenna I put the meter to work on was my 40/80-meter dipole. I took measurements on both bands, recording the readings as shown in fig. 7. The readings indicated the 80- and 40-meter elements had the lowest reflected readings at the bottom of the bands. Shortening the 40-meter elements

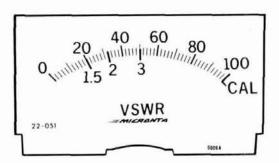
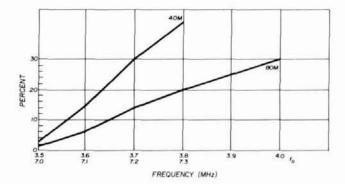


fig. 6. New calibration markings for the meter face.



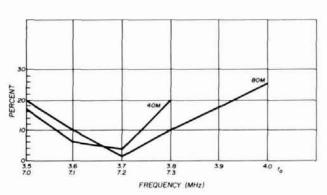


fig. 7. Readings obtained with the swr meter on a dipole for 40 and 80 meters.

first, and then the 80-meter elements, eventually yielded the reading shown in fig. 7B. This was more in keeping with the section of the band I use most. Now the antenna meter is left in the line and I use it to squeeze the last drop of output out of the little transceiver. I honestly cannot say that there has been a dramatic improvement in the overall performance of the antenna, but I am happy knowing that the vswr is more in keeping with the specifications for the transceiver.

ham radio

.. pacesetter in amateur radio



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• Select transmit offset (1 MHz/600 kHz)

The Remote Controller's display indicates frequency (even while scanning) and functions (such as autoscan, lower scan frequency limit, upper scan limit, error, and call channel).

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 Full 4-MHz coverage (144.000-147.995 MHz) on 2 meters • 800 channels • Dual concentric knobs for fast frequency change (100-kHz and 10-kHz steps) • 5-kHz offset switch • MHz selector switch... for desired band (144, 145, 146, or 147 MHz) • Mode switch for operating simplex or for switching the transmit

frequency up or down 600 kHz for repeater operation ... or for switching the transmitter to the frequency you have stored in the TR-7600's memory (while the receiver remains on the frequency you have selected with the dual knobs) • Memory channel ... with simplex or repeater (plus or minus 600 kHz transmitter offset) operation. . Digital frequency display (large, bright, orange LEDs) UNLOCK indicator ... an LED that indicates transceiver protection when the frequency selector switches are improperly positioned, or the PLL has malfunctioned • 10 watts RF



Still the same fine, time proven rig. But now with the simple addition of a plug-in crystal, the TS-700SP will be able to utilize the new repeater sub-band (144.5 to 145.5 MHz). Still features all of the fine attributes of the TS-700S: A digital frequency display, receiver pre-amp, VOX, semibreak in, and CW sidetone. Of course, it's all mode, 144-148 MHz, VFO controlled . . . and Kenwood quality throughout.

Features: 4 MHz band coverage (144 to 148 MHz) · Automatic repeater offset capability on all FCC authorized repeater subbands including 144.5-145.5 MHz • Simply dial receive frequency and radio does the rest . . . simplex, repeater, or reverse. Same features on any of 11 crystal positions . Transmit/Receive capability on 44 channels with 11 crystals . Operates all modes: SSB (upper and

lower), FM, AM and CW . Digital readout with "Kenwood Blue" digits • Receiver pre-amp • Built-in VOX • Semi break-in on CW • CW sidetone • All solid-state • AC and DC capability • 10 watts RF output on SSB, FM, CW • 3 watts on AM · 1 watt FM low-power switch · 0.25 µV for 10 dB (S+N)/N SSB/ CW sensitivity . 0.4 µV for 20 dB quieting FM sensitivity.

OR 25 WATT OUTPUT



TR-7400A

The fully-synthesized TR-7400A 2-meter FM transceiver operates on 800 channels and features repeater offset over the entire 144-148-MHz range, dual frequency readout, six-digit display, and subaudible tone encoder and decoder. RF output is at least 25 watts!

The TR-7400A 2-meter FM transceiver provides fully synthesized operation, including 600-kHz repeater offsets, over the entire 144-148-MHz range. It can operate on any of 800 channels, spaced 5 kHz apart. RF output is at least 25 W, and typically 30 W. A low power position produces 5-15 W (adjustable). Included is a dual frequency readout with large six-digit LED display plus a dial readout. The sub-

audible CTCSS signaling feature may be used on transmit and receive, or transmit only. Optional tone-burst modules are available. Receiver sensitivity is better than 0.4 μV for 20 dB quieting. Large, high Q, helical resonators minimize interference from outside the band. A two-pole 10.7-MHz monolithic crystal filter provides excellent selectivity.

Intermodulation distortion is down more than 66 dB, spurious rejection is better than -60 dB, and image rejection is better than -70 dB. See your local Authorized Kenwood Dealer today, for a demonstration of the fantastic TR-7400A.



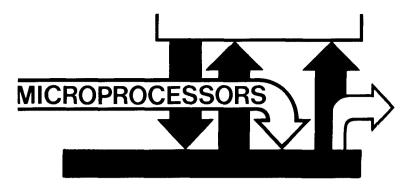
TS-600

Experience the excitement of 6 meters. The TS-600 all mode transceiver lets you experience the fun of 6 meter band openings. This 10 watt, solid state rig covers 50.0-54.0 MHz. The VFO tunes the band in 1 MHz segments. It also has provisions for

fixed frequency operation on NETS or to listen for beacons. State of the art features such as an effective noise blanker and the RIT (Receiver Incremental Tuning) circuit make the TS-600 another Kenwood "Pacesetter".

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IC tester using the KIM-1

One useful tool that's sure to be appreciated by any electronic or computer experimenter is a reliable, fast, and easy method for testing digital integrated circuits. The concept presented in this article uses the KIM-1 microprocessor as one approach to testing 7400 series ICs. While only 7400-type numbers are listed, the program is equally applicable to those devices in the 74H, 74L, 74S, 74LS, and 74C series, as well as to others whose pinout arrangement is identical to that of the 7400-type shown. Also included is an optional "search and identify" procedure to help in the identification of unmarked devices.

program basics

The program itself is basically simple. A flow chart showing the concepts involved is seen in fig. 1. Operation of the program is broken into three main parts: selection of the proper set of input combinations from a table of possible combinations; application of inputs to the device under test; and comparison of the results with the set of correct results contained in another table.

A third table of "pointers," actually the lower address byte of locations in program page 1 and 3 where the input and results tables are located, is provided in program page 0. The third table is used to identify the starting address for each variable in the program for each type of chip listed. The correct set of pointers is selected by inserting into location 0000 an appropriate code number for the chip to be tested. The ICs that can be tested are shown in table 1.

When an IC has successfully completed the testing, the device number appears in the address locations of the KIM-1 display. For type numbers above 7499, a hexadecimal digit is used to reduce the dis-

By Robert E. Babcock, W3GUL, 1706 Fawcett Avenue, McKeesport, Pennsylvania 15132

table 1. Listing of 14- and 16-pin ICs which can be tested by the KIM-1 microprocessor using the program described in

	14-pin ICs		16-pin IC:
7400	7415	7439	7442
7401	7416	7440	7445
7402	7417	7451	7446
7403	7420	7453	7447
7404	7421	7454	7448
7405	7422	7474	7449
7406	7426	7486	7485
7407	7427	74107	7490
7408	7428	74125	7492
7409	7430	74126	7493
7410	7432	74128	74145
7411	7433	74132	74148
7412	7437	74136	74151
7413	7438	74140	74153
7414			74155
			74157
			74158

play to four digits. For example, a 74107 would be shown as 74A7 and a 74157 as 74F7. Failure of any tested chip to match expected results will result in FFFF being displayed in the address locations.

To use the program, connections must be made between the KIM-1 and a suitable test socket. When testing a 14-pin IC, all pins are connected to the KIM-1 for maximum flexibility. However, when a 16-pin device is tested, all pins except ground and V_{CC} are connected to the KIM-1. The 14-pin 7400 series devices shown in table 1, with three exceptions, all use pin 14 for V_{CC} and pin 7 for ground. The three exceptions are the 7490, 7492, 7493. The 16-pin devices all use pin 16 for V_{CC} and pin 8 for ground. Note that the KIM-1 ports cannot supply enough current as a V_{CC} source, therefore, a separate connection must be made.

The following procedure is recommended before inserting the first IC into the test socket.

table 2. Connections between the test sockets and the KIM-1 applications connector.

14-pin socket	16-pin socket	KIM-1 applications connector	KIM-1 port designation
1	1	8	PA7
2	2	7	PA6
3	3	6	PA5
4	4	5	PA4
5	5	2	PA3
6	6	3	PA2
7	7	4	PA1
8	9	14	PA0
9	10	16	PB5
10	11	13	PB4
11	12	12	PB3
12	13	11	PB2
13	14	10	PB1
14	15	9	PB0

^{*}A complete program listing, IC lists, and sequence tables are available by sending a self-addressed, stamped envelope to ham radio, Greenville, New Hampshire 03048

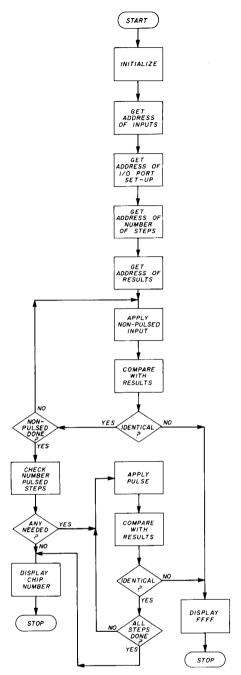


fig. 1. Flow chart for the program used with the KIM-1 microprocessor to evaluate and identify ICs.

- Put the code for the selected IC into location 0000.
- Put the starting address for the program in the ST vector (00 in 17FA, 02 in 17FB), then set address to 0200.
- 3. Press GO button display should read FFFF IC.
- 4. Attach V_{CC} and ground connections to the correct pins of the IC. If adequate, the KIM-1 power supply may be used. If an external supply is used, the grounds must be connected together.
- **5.** Insert the IC and press ST button display should read type number if good, FFFF if not.

6. Repeat step 5 for each additional IC of the same type.

This procedure ensures that the input-output ports are correctly configured for the IC to be tested, which will reduce the possibility of damage to the I/O port due to shorting of outputs and "fighting for the bus." To change the code in location 0000, leave the test socket empty and press ST. The display will again read FFFF. Pressing the + button will put the address in location 0000.

A delay subroutine is included as part of the program so that the changing values may be observed by attaching indicators (with drivers) to the various ports. To increase the delay for this purpose, put FF in 02D6 and 07 in 02D8. If observation is not desired, the delay may be removed by inserting NOP's (Op Code EA) in 0263, 0264, and 0265, and also, 0290, 0291, and 0292.

additional test capabilities

If you desire to include additional IC types, caution must be exercised. One factor that must be considered is the load imposed on the driving source by any input of a TTL chip. Some devices, such as the 7475, impose as much as four loads on certain inputs, far beyond the unbuffered capacity of the KIM-1. A second, less critical problem is the number of memory locations that are needed to provide the input and output sequences for some of the more complex ICs. Finally, in addition to the table of pointers in page 0, three other locations in that page are reserved for use by the program; these are OOEC, OOED, and OOEE. Any tables to permit testing devices not included in table 1 should avoid these locations.

IC identification

Through a small program modification, it is possible to use the KIM-1 to determine the characteristics

table 3. Example of binary values necessary to check a 7400 quad, dual NAND gate.

KIM-1 port designation	IC pin number	pin function	binary entry to DDR	hex equivalent
PA7	1	in	1	
PA6	2	in	1	D
PA5	3	out	0	
PA4	4	in	1	
PA3	5	in	1	
PA2	6	out	0	8
PA1	7	ground	0	0
PA0	8	out	0	
PB7	_	_	0	
PB6	_	_	0	3
PB5	9	in	1	3
PB4	10	in	1	
PB3	11	out	0	
PB2	12	in	1	6
PB1	13	in	1	U
PB0	14	v_{cc}	0	

of unmarked chips. To operate in this automatic search and identify mode, enter code 00 in location 0000, enter the program starting address (0200) into the ST vector, insert the unknown IC in the test socket, and press ST. The program change will cause the code in 0000 to be incremented by one if the IC fails the test for the device coded 00. If the IC coded 01 also fails, then the code will be incremented to 02 and so on until all codes for which input and output sequences have been provided have been checked. If no match is found, the display will show FFFF as before.

The *first* code which is satisfactorily matched will cause that device type to be indicated by the display. This can lead to errors in identification. As an example, suppose the IC being tested is 7420, a dual 4-input NAND gate. Investigation of the characteristics of the other devices in **table 1** will show that a 7413 is also a dual 4-input NAND Schmitt trigger. Thus, a 7420 would actually show as a 7413 in this mode. However, the fact that the device is identified as a dual 4-input NAND is helpful, and further tests can be made to completely identify the device. After the IC has been identified, remove the chip and press ST. This will make certain that the code is initially at 00 for the next chip tested.

The sequences of inputs and results, as well as the configuration for the input-output ports, can be determined by analysis of the chip and its connection to the KIM. Consideration must be given to the fact that any pin designated as an IC output must have the corresponding KIM-1 port set as an input, and vice versa. Unused IC pins, V_{CC}, and ground connections are made inputs to the KIM. Since a port location is designated as an output terminal by writing a 1 into the data direction register for that position, and similarly an input by a 0, the known pattern of a particular IC establishes the values which must be used to set up the ports for that device.

This procedure is probably best explained by an example using the 7400 two-input NAND gate. The pin arrangement for the 7400 and the truth table for each of the four identical gates contained in the chip are shown in **fig. 2**. By referring to the pinout, a table of the KIM-1 port designations, the IC pins attached

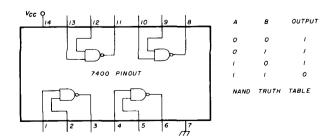


fig. 2. Pinout diagram and truth table for the 7400, a quad, dual-input NAND gate.

table 4. Binary inputs for the 00, 01, and 10 testing combinations.

KIM-1 port	IC pin	input	re	sults
designation	number	00 01 1	0 00	01 10
PA7	1	0 0 1	0	0 1
PA6	2	0 1 0	0	1 0
PA5	3	0 0 0) 1	1 1
PA4	4	0 0 1	0	0 1
PA3	5	0 1 0) 0	1 0
PA2	6	0 0 0) 1	1 1
PA1	7	0 0 0	0	0 0
PA0	8	0 0 0	1	1 1
PB7	_	0 0 0	0	0 0
PB6	_	0 0 0	0	0 0
PB5	9	0 () 1	0	0 1
PB4	10	0 1 0	0	1 0
PB3	11	0 0 0	1	1 1
PB2	12	0 0 1	0	0 1
PB1	13	0 1 0	0	1 0
PB0	14	0 0 0	1	1 1

to each, and the binary value which must be placed in each location can be made (see table 3).

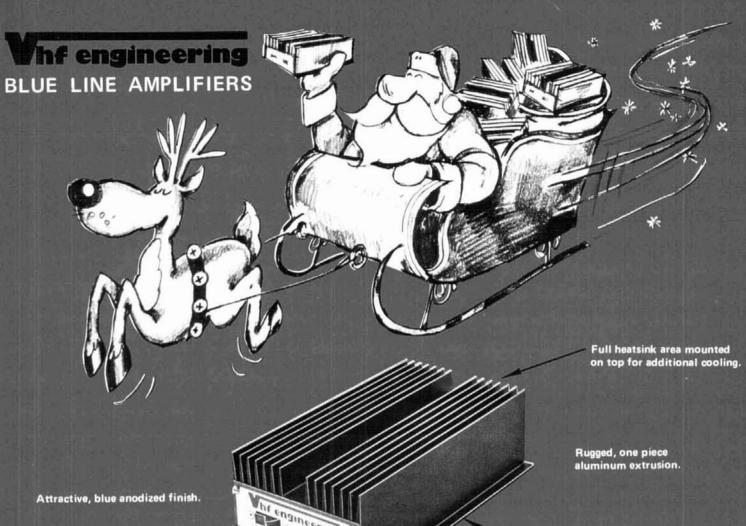
As can be seen from the hexadecimal equivalents for the binary values needed at the data direction registers, D8 must be written into PADD (1701) and 36 into PBDD (1703). The truth table also indicates that the same hex digits can be used to apply the 1-1 input combination to all four gates simultaneously. Thus, these values fulfill two functions and thereby save space in the input sequence table.

The result digits can also be determined by reference to the truth table; each 1-1 input should result in an output of 0. In fact, when the ports are sampled for correct response, the only change from the written value will occur at PB0 where V_{CC} will cause a 1 input at that port. The result digits would then be D8 at Port A and 37 at Port B. The values to be written into PAD (1700) and PBD (1702) (to apply the 1-1 combination) and the values which should read at PAD and PBD if the chip responds correctly have now been determined. The combinations for the remaining 00, 01, and 10 inputs are shown in binary form in **table 4**.

The portion of the program that applies the pulses for the 7490, 92, and 93 or other similar counters requires only two input entries for all of the pulses applied. The pulse is obtained by starting with the input at a 1 level, applying a 0 level, then returning the pin to a 1 level. After each pulse, the index that determines the input values is decremented two places; the next pulse is applied by incrementing twice and so on. This also saves program space and permits the testing of more chips with a given set of tables. The process of determining the needed tabulations is not difficult and it should be possible to find combinations that can test virtually any logic chip.

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simplified capacitance meter

A revised version of the capacitance meter uses a second 555 timer to replace the original programmable unijunction

In a previous article I described what I thought to be a very easy-to-build, direct-reading capacitance meter. 1 It would linearly read on a meter any capacitance value from 1.0 µF down to as low as 1.0 pF, just the ticket for measuring those unmarked variables and micas you can get so cheaply at the surplus and flea markets.

Well, the article proved to be guite popular, at least I got a lot of letters and phone calls about it. Only trouble was most of them asked, "Where can I get that programmable unijunction transistor (PUT), or what can I substitute for it?" Information on substitute devices was published,2 but I'm still getting phone calls nearly three years later. The circuit presented in this article is an alternative; no PUTs, just two of the popular 555 timer ICs.

theory

As it turns out, the output pulse width of a 555 timer is a linear function of the timing capacitance connected to it. If the 555 is triggered with a constant frequency, then the average dc value of the resulting pulse train is a linear function of pulse width. Putting these two facts together means that a dc meter connected to the 555's output will provide a linear indication of the timing capacitor's value. Decade capacitance ranges are implemented by switching the value of the 555 timing resistor.

Since the output waveform of the 555 doesn't quite get down to zero volts between pulses, and since there is some stray capacitance in any circuit, zero adjustments are needed to buck out that small voltage and stray capacitance. In the original circuit, I used a PUT and a transistor inverter to generate the constant-frequency trigger for the 555; here I have replaced the original circuit with another 555 operating in the free-running mode. A more detailed discussion of theory may be found in reference 1.

construction

The schematic of the capacitance meter is shown in fig. 1. It is convenient to mount most of the parts, including the four trimpots, on a piece of perforated board which is mounted to the meter terminals. The two 555 ICs will plug into a 16-pin IC socket.

Layout is not critical, but the wiring associated with the range switch and test terminals should be as short as possible to minimize stray capacitance. Use a metal box to house the instrument, and connect circuit ground to the box. Make sure the 2-pole, 5position switch is wired so that trimpot R11 is selected when the range switch is in the 100 pF position.

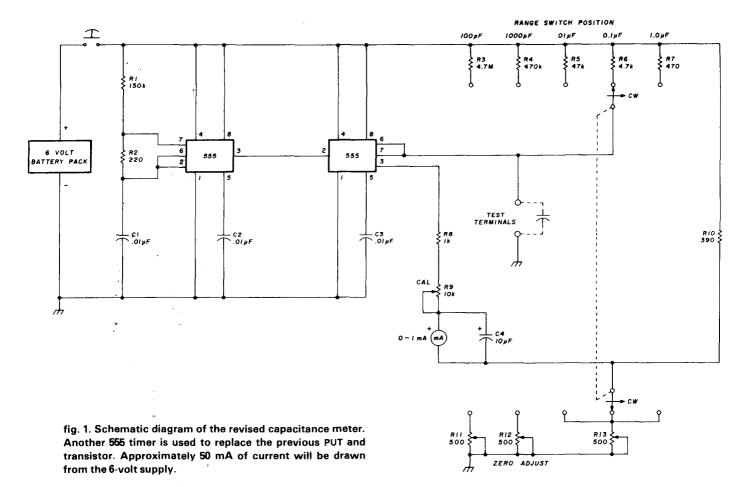
All resistors may be 1/2 watt, 5 per cent, but greater overall accuracy will result if the five resistors connected to the range switch are of one per cent tolerance. C1 should have a tolerance no greater than 10 per cent.

power supply

The circuit is designed for a six-volt power supply, to facilitate battery operation. However, the battery drain is heavy, up to about 50 mA, and varies with the range-switch setting. Since calibration stability is directly related to supply-voltage stability, batteries really shouldn't be used unless portable field-use is required. If batteries are used, they should be alkaline

My solution to the problem of infrequent use vs. cost of a built-in power supply is as follows. A simple

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75248



six-volt zener regulator, such as shown in **fig. 2**, is mounted in the capacitance meter box with two wires brought out. These wires are connected to a 12-volt bench power supply when I need to use the meter. One of the three-terminal, six-volt IC regulators should give even more stable performance than a zener.

calibration

The three zeroing trimpots, R11, R12, and R13, must be adjusted with nothing connected to the test terminals. Zeroing is accomplished by adjusting these trimpots for a zero meter reading when the push-button is depressed. Adjust R11 with the range switch in the 100-pF position. R12 is for the 1000-pF range, and R13 is for the three highest ranges. If the meter cannot be zeroed, the trimpot is probably defective, having too much "end resistance." Smaller value trimpots may be used, down to 200

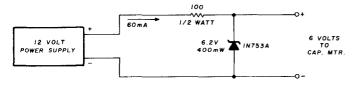


fig. 2. Zener regulator for use with the capacitance meter.

ohms, or possibly smaller, depending on the stray capacitance in a particular version of construction.

After the zeroing trimpots are adjusted, connect a 100-pF mica capacitor, with a tolerance of five per cent or better, to the test terminals. Set the range switch to 100 pF. Depress the push-button and adjust trimpot R9 for a full-scale reading on the meter. Calibration is then complete.

operating tips

When the capacitor being measured is too large for a particular range-switch setting, the circuit may be driven out of its linear range of operation. Under these conditions, the meter may read less than full-scale even though the actual capacitor value is more than the full-scale setting. To avoid such erroneous readings, test an unknown capacitor on the 1.0- μ F range first, then move the range switch to lower settings until a usable reading is obtained. Keep the original calibration capacitor handy for calibration checks.

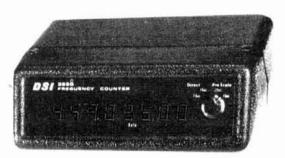
references

- 1. Courtney Hall, WA5SNZ, "Direct-Reading Capacitance Meter," ham radio. April. 1975. page 32.
- 2. Courtney Hall, WA5SNZ, Comments, ham radio, October, 1975, page 31.

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improvements to the Measurements Corporation 59 grid-dip oscillator | components. The 5Y3 rectifier was replace silicon rectifiers, and the OD3/VR150 voltage.

Used in labs for years, the model 59 grid dipper is still a great instrument.

Presented here are some ideas for updating this old workhorse

This article describes modifications to a Measurements Corporation model 59 grid-dip meter. However, it pertains to most other tube-type grid-dip meters as well. The B&W, Heath, and Millen grid-dip meters are basically similar (in the rf head) and should be adaptable to the circuitry described.

I found an rf head from a model 59 in a local surplus emporium without the meter/power-supply module and with only four of its normal complement of seven hf-vhf coils. Since the model 59 (in my opinion) is the best grid-dip meter ever built, I felt that a reconstruction project was worthwhile, so a new meter/power-supply unit was built to replace the original. Here's the story.

circuit design

Rather than copy the Measurements Corporation circuit, I designed a new supply using all solid-state

components. The 5Y3 rectifier was replaced with silicon rectifiers, and the OD3/VR150 voltage regulator tube was replaced with a 10-watt zener. As neither an O.E.M. transformer or choke was available, I used standard Triad components. I added a B + switch; this allows you to first switch on the power, which heats the 955 heater in the rf head, and then switch on the B + after a few moments wait —

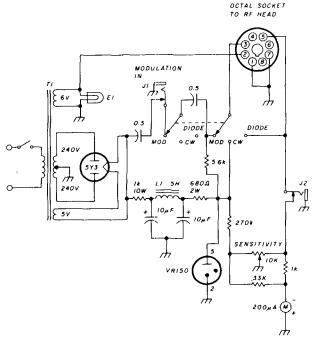


fig. 1. Original power-supply/metering portion of the Measurements Corporation model 59 grid-dip oscillator.

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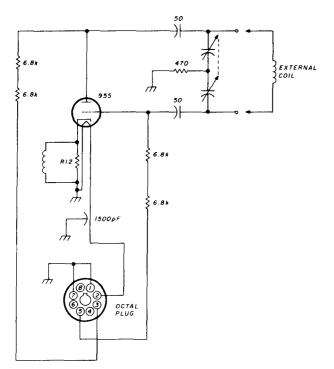


fig. 3. Schematic of the Measurements model 59 rf head. The 955 could perhaps be replaced with a modern high-performance fet.

a conservative addition because replacement 955s are now about \$10 on the surplus market.

Looking at fig. 1, which shows the original schematic diagram, we see tubes used for the rectifier and voltage regulator. Fig. 2 shows the replacement supply using solid-state devices. The rf-head is shown in fig. 3, and is the same for original and modified GDOs.

internal modulation

Note that in **fig. 1**, when the function switch is placed in the **MOD** position and no plug is inserted into closed-circuit jack J1, 120 Hz is capacitively coupled to the B + going to the rf head. The equivalent circuit is shown in **fig. 4**.

This simple method of plate-modulating the GDO leaves a lot to be desired, mostly because of the prevalence of 120 Hz as an incidental modulation on almost every oscillator. So this is the question: Is the 120-Hz modulated carrier you're hearing really from the GDO or from some other source?

To make the modulation frequency more distinctive, I added a simple, solid-state, 1-kHz modulator. This modulator uses a four-layer diode relaxation oscillator followed by a solid-state amplifier to increase level (**fig. 5**). Although this modulation isn't sinusoidal, it has a more definitive pitch (400 Hz would be another good frequency and could easily be obtained by changing the RC constants in the relaxation oscillator).

I made no attempt to modify the rf head, although it's tempting to make the GDO entirely solid state by replacing the 955 tube with an fet.

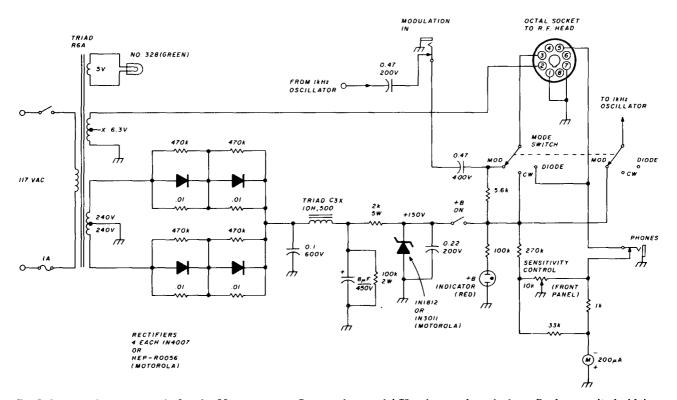


fig. 2. Improved power supply for the Measurements Corporation model 59 using modern devices. Design results in higher efficiency and longer life for this essential part of the circuit.

I built the meter/power supply into an LMB-W1A box, and screwed the wooden coil block onto the back. I then drilled a small slot into one upper edge of the W1A cabinet so that the hook on the rf head could engage it.

Measurements Corporation's successor, as well as their "standard of the industry" grid-dip meter, are both very much alive. The company is now a subsidiary of the McGraw Edison Company, Manchester, New Hampshire. The new GDO is called the model 159, but it remains pretty much the same old unit as its predecessor, the model 59. Incidental-

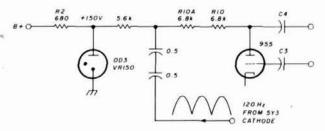


fig. 4. Equivalent circuit in the model 59 when using the unit for internal modulation. The 120-Hz modulation from the 5Y3 cathode, which appears on the 955 oscillator plate, causes ambiguous results. A better circuit is shown in fig. 5.

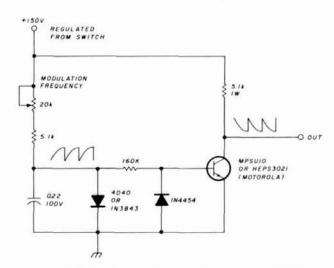


fig. 5. Simple modification to the Measurements Corporation model 59 GDO to provide 1-kHz modulation.

ly, for those (like myself) who have a couple of coils missing, it's still possible to get single replacements from McGraw Edison; they are located at Grenier Field Municipal Airport, Manchester, NH 03103.

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VHF 30-160P	128-148(5 MHz)	30W	160W	25A	70		14.9		32.5cm	1.6kg	\$289.95
1.38430-1406		25W	140W	23A	70		14,9		32.5cm	1.6kg	\$269.95
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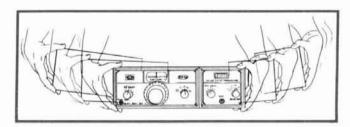
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- The TX-110-H runs 200 watts input on 20, 40, and 80 meters; 150 watts on 15 and 100 watts on 10 meters.
- Full band coverage on 15 through 80 meters; 28 to 29 MHz on 10 meters.
- Semi-break-in CW with sidetone monitoring is a standard feature.
- PTT (Press-to-Talk) operation on SSB. Lower sideband on 40 and 80 meters. Upper sideband on 10, 15, and 20 meters.
- TX-110-L 15 watt module runs on AC supply in RX-110, so it is completely self contained, including speaker. Simply connect antenna, and key or mike.
- TX-110-H requires additional AC supply to supply high current for 200 watt amplifier (Model PS-110).
- 200 watt amplifier may be added to TX-110-L at a later date, thus converting it to a TX-110-H.

- Modular design provides much easier service and maintenance. With cabinets removed everything is wide open and fully accessible. This is a piece of solid state equipment you can work on yourself, if you wish, because you can get at everything with ease.
- The RX-110, TX-110-L, and TX-110-H will all run directly from a 12 to 14 volt DC battery supply for mobile or portable operation. When the two units are mechanically joined (brackets supplied with TX-110), the transceiver slides into a plug-in mobile mount, model MM-110.

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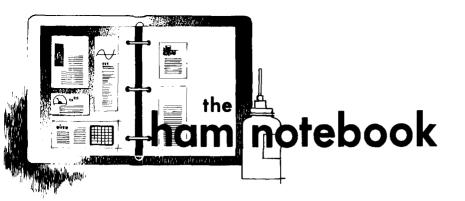
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positive lead keying for the Heath HD — 10 keyer

This is a simple and fast modification for the Heath HD-10 kever for use with the newer solid-state transmitters and transceivers that require positive keying to ground. The HD-10 is designed for negative keving. Examination of the circuit in the manual shows a PNP transistor (2N398A) with the collector to the keved line. On page 15 of the assembly manual, the lead arrangement for the keying transistor is shown with instructions to bend the base lead closer to the emitter. For solid-state units, remove Q8 from its socket, bend the base lead closer to the collector, and replace the transistor in its socket.

In my case, I used a 2N1305. Since the keying current is small (5 milliamperes), most any PNP transistor will do if one wishes to keep the original 2N398A for resale purposes. This modification takes about ten minutes because of the time required to remove the printed circuit board and replace it. No external buffer is required.

Richard Jasper, W4VAF

cleaner audio for the R-4C

The Drake R-4C offers many attractive features for the weak-signal enthusiast (excellent AGC, noise blanker, and selectable i-f filters). However, a major annoyance is the large amount of hiss in the audio. This hiss can be traced to the 50-kHz BFO feeding into the audio circuit. An effective cure is to roll off the audio

response and put a filter between the product detector and the subsequent audio stages.

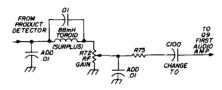


fig. 1. Schematic diagram of the changes made to the Drake R4C to eliminate the 50-kHz BFO feedthrough into the audio system.

The following simple circuit changes (see fig. 1) are an effective cure for the problem.

- 1. Change C99 to 0.1 μ F
- 2. Connect a $0.002-\mu F$ capacitor in parallel with R83
- 3. Change C100 to 0.47 μ F, improving the low-frequency audio response
- 4. Connect a $0.1-\mu F$ capacitor from the wiper of R72 to ground
- 5. Connect an 88-mH toroid, with a $0.01-\mu F$ capacitor in parallel, between the audio gain control and the product detector output
- **6.** Connect a $0.01-\mu F$ capacitor across the output of the product detector
- 7. Replace C103 with a 0.005- μ F capacitor
- 8. Change C175 to $100 \mu F$

These changes will eliminate the hiss and also clear up the low-frequency distortion.

Steve Powlishen, K1FO

crowbar circuit for the HWA-2036-3

Recently I had a problem with my HWA-2036-3 power supply which caused the driver transistor to fail. The cause of this failure was that the collector of the 2N3055 pass transistor and the mounting screw (on rear chassis) failed to make contact. thereby placing the entire load on the driver transistor. An inspection showed that I had forgotten the small 3-mm (1/8-inch) spacers in the plastic cover over the pass transistor. These spacers are very important because they force contact between the mounting screws and the pass-transistor case.

The resultant failure caused approximately 24 Vdc to be applied to the HW-2036. This increase in voltage caused the receiver's audio chip (TBA820) to burn out. I was lucky that nothing else went up in smoke. If the HW-2036 had been in the transmit mode, many hard to replace components would have been destroyed.

To solve this problem, I decided to build a simple "crowbar" circuit, within the power supply, which would blow the dc fuse if the output voltage exceeded 15 volts. The crowbar circuit is quite simple and easy to construct; it can be placed on a small printed circuit board or mounted on terminal strips. But, modification of the HWA-2036-3 requires some mechanical work. The dc fuse has to be relocated to a point ahead of the regulator instead of at the power supply output. This was done to protect the transistors and regulator chip. When the fuse is blown, it removes voltage from the entire regulator.

The following steps are required for this modification:

1. Referring to the circuit board X-ray picture on page 30 of the manual, cut the trace at a point opposite the diode D2 (it's pretty wide here). This will separate the filter section from the regulator section.

- 2. Drill two small holes (no. 55 drill [1.3 mm]) in the trace just below the one cut in step 1. This trace connects to Q1's collector and one side of R2. (These holes will be used for wire connections.)
- 3. Disconnect the red wire from point C on the circuit board and resolder it in one of the holes just drilled.
- 4. Disconnect all wires from the dc fuse socket.
- **5.** Disconnect the wires from the DC Regulated post on the printed circuit board.
- **6.** Disconnect the white wire from A on the board.
- 7. Solder a jumper wire from point A to DC Regulated on the board.
- 8. Solder the red wire which goes through the grommet to the plug on the rear panel to the DC Regulated post on the printed circuit board.
- **9.** Solder a red wire to the remaining drilled hole and solder the other end of this wire to one side of the dc fuse holder.
- 10. From the other side of the dc fuse holder, solder a red wire to point C on the board. (Just above diodes D1 and D2.)

This in effect relocates the dc fuse between the filters and the regulated section of the power supply (referring to the circuit diagram, at a point between C1 and R2/Q1's collector).

Now that the power supply has

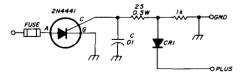


fig. 1. Schematic diagram of the simple crowbar circuit used to protect the output of the HWA-2036-3 power supply. The Zener diode is rated at 15 volts, 1 watt. The SCR is rated at 100 volts, 8 amps, but other SCRs can be used.

been modified, the crowbar circuit can be constructed and installed. Referring to fig. 1, you'll note that there are only 5 components, an SCR, 2 resistors, a Zener diode, and a bypass capacitor. Also note that there are only 3 connections, marked PLUS, FUSE and GND.

PLUS is connected to the printed circuit board at the DC Regulated terminal post.

FUSE is connected to the dc fuse holder on the same terminal as the red wire from R2/Q1.

GND is connected to chassis ground.

Once the SCR has fired and blown the dc fuse, the transceiver is protected from damage due to overvoltage. But, don't just replace the dc fuse and expect it to take off and work properly. The reason the fuse blew in the first place indicated trouble in the power supply. First, disconnect the transceiver, take the case off the power supply, and discharge the filter capacitor, C1. The SCR will blow the fuse and save your transceiver, but it will not discharge the filter capacitor. After the capacitor has been discharged, normal troubleshooting procedures can be used to locate faulty components. As a final note, use a fast blow fuse in the dc fuse holder.

James T. Conner, W3HCE

external speaker and tone pad for the HR-2B

The Regency HR-2B transceiver provides screw terminals at the rear for connecting an external speaker. I found this method to be rather awkward since I'm frequently removing the rig from the car. By carefully prying the ears up on the terminal strip, the individual screw terminals may be removed. The remaining holes are large enough to permit the

installation of miniature phone jacks. I used one closed-circuit jack for normal-through internal speaker operation and an open-circuit jack for plugging in the *Touch-Tone** pad. When the mating plug is inserted into the speaker jack, the internal speaker is muted and the external speaker is operative. Thus, connect/disconnect for mobile/base use is rapid and easy.

Paul Pagel, N1FB

setting 2-meter receivers using hf harmonics

In the absence of a counter or other suitable piece of test equipment, a surprisingly good job of setting a 2-meter fm receiver on frequency can be done using the harmonics from a 15- or 10-meter transmitter. This assumes that the transmitter or its companion receiver has reliable dial calibration. In my case, I depend on my Collins 75A4 as the frequency standard, since its dial can be read to 100 Hz increments with little difficulty, and it tracks to within 100 Hz between 100 kHz calibration points.

To set the 2-meter receiver on frequency, divide the desired channel frequency by five, tune the 10-meter receiver to that frequency, zero beat the transmitter, and zero the 2-meter receiver on the transmitter harmonic. If your transmitter doesn't cover all of 10 meters (147 MHz/5 = 29.4) MHz), the same trick can be done on 15 meters using the seventh harmonic for frequencies above 147 MHz. (147 MHz/7 = 21.0 MHz). It may be necessary to temporarily remove your low-pass filter to hear these harmonics. Make sure you are not tuning up on top of someone on 15 or 10 when you are doing this.

Using this technique, it should be possible to set a 2-meter fm receiver to within 1 kHz or less. Once the receiver frequency is set, most radios have provision for zeroing the transmitter to the receiver.

John Becker, K9MM

^{*}Touch-Tone is a registered trademark of the American Telephone and Telegraph Company.

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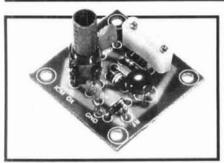
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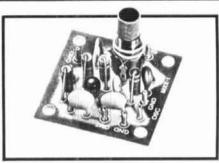


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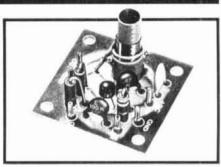
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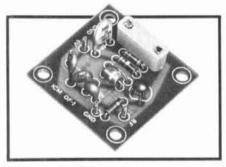


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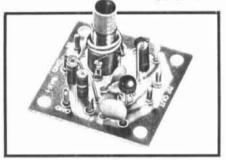
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Reflector Antennas

This is a book in the IEEE Press Selected Reprint series, prepared under the sponsorship of the IEEE Antennas and Propagation Society. While the book could have been well titled by any of a number of catchy modern phrases, its title, Reflector Antennas, sums it up nicely. The book is about reflector-type antennas, and it thoroughly covers the subject.

Editor A. W. Love has selected some 65 articles for reproduction in this book, and they are placed in nine distinct categories of interest. Some of the papers are classics well deserving of a place in your library: "Parabolic Antenna Design for Microwaves" by C. C. Cutler, from the Proceedings of the IRE, November 1947, for example; and "On the Simulation of Fraunhofer Radiation Patterns in the Fresnel Region" by D. K. Cheng, from the IRE Transactions on Antennas and Propagation, October 1957. Not all of the reprinted papers are from IRE or IEEE publications; you'll find some from the Bell Systems

Technical Journal, Electronics Letters, The Microwave Journal, and others.

Reflector Antennas is an educational volume, long on theory and the mathematics necessary to support the theory. It is also historical in that it follows - through the papers selected - the development of reflector antennas from the simple paraboloid to the more complex reflector sections with specialized feeds. In his opening paper, editor Love recounts the very beginnings of microwave energy and the use of reflector antennas as part of the exploration of it. His paper is reprinted from Radio Science, August/September 1976, and in it he reminds us of work done by an Indian physicist, J. Chunder Bose, who, at the Royal Institution in London in 1897, set up a microwave spectrometer using a horn antenna, plane and cylindrical mirrors, a dielectric prism, and a hollow waveguide radiator. Some of his experiments used a wavelength as short as 5 mm!

Reflector Antennas traces microwave industry from early experiments through World War II radar and the feed systems necessary for defenseoriented microwave radiators, then into the modern space-age requirements for very large paraboloidal and spherical reflectors with Cassegrain and dual-mode feeds used for today's globe-spanning communicationssatellite and radio-astronomy systems.

If the material presented in this book is not enough to satisfy your need for knowledge in the microwave antenna field, the extensive bibliographies at the end of each section will

provide a source of reading which will keep you thoroughly immersed in the field for as long as need be.

Reflector Antennas, 428 pages, 8-1/2 × 11 inches, is available from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048. Order JW-RA, \$14.95 plus 40 cents shipping and handling.

ssb transceiver



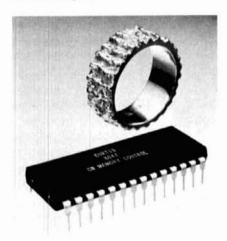
A new 100-watt minimum PEP single sideband mobile transceiver has been introduced by Swan Electronics, a subsidiary of Cubic Corporation. The 100 MX mobile transceiver is completely solid state and incorporates state-of-the-art design and styling. It features a highly reliable, extremely stable Permeability Tuned Oscillator (PTO) with 1-kHz readout resolution; built-in noise blanker and VOX; semi-break-in CW with sidetone; Receiver Incremental Tuning (RIT); 25-kHz built-in calibrator and preselector for transmit and receive.

Frequency ranges for the new unit are: 80 meters (3.5-4.0 MHz), 40 meters (7.0-7.5 MHz), 20 meters (14.0-14.5 MHz), 15 meters (21.0-21.5 MHz), 10 meters

(28.5-29.0 MHz). Modes of operation include USB, LSB, and CW. Extended frequency coverage in 500-kHz segments of the 10-meter band (28.0-28.5; 29.0-29.5; 29.5-30.0) is achieved by replacing the standard crystal with an optional crystal for the desired segment. No realignment is required.

The receiver sensitivity is better than $0.35 \,\mu\text{V}$ at 50 ohms for 10 dB signal plus noise-to-noise ratio for all bands. Audio output is four watts into a 4-ohm load. Audio bandpass response is 300 to 3000 Hertz. Provisions for an external speaker or headphones are on the rear panel, and a gimbal-type mobile mount is included as standard equipment. For additional information contact Chuck Inskeep, Director of Marketing, Swan Electronics, 305 Airport Road, Oceanside, California 92054.

memory control IC



A one-chip message memory control IC has been introduced by Curtis Electro Devices. Called the 8047, this 28-pin CMOS device requires only a 2102 (1K × 1 RAM) or equivalent memory IC plus an 8043/8044 or equivalent keyer to provide a programmable set of four 32-character CW messages.

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The SCR1000/SCAP is a complete Autopatch Repeater — fully assembled, set-up and checked-out in our lab. As with all Spec Comm products, all workmanship and components are of the very highest quality. The price? A very reasonable \$1700.00 — complete! (\$2195.00 w/WP641 Duplexer.) (For Rptr., w/o Rcvr. Preselector, deduct \$85.00.) Get your order in A.S.A.P.!

- The SCR 1000 the finest repeater available on the amateur market . . . often compared to "commercial" units selling for 3 times the price! This is a 30Wt. unit, with a very sensitive & selective receiver. Included is a built-in AC Supply, CW IDer, Rcvr. Preselector, full metering and lighted status indicators/control push buttons, crystals, local mic, etc. Also, jacks for emergency power, remote control, autopatch, etc.
- Custom options available: Duplexers, Cable, 'PL', HI/LO Power, Touch ToneTM Control, 60-70W. Transmitter, Racks, etc. Inquire.
- The Spec Comm Repeater System . . . a sound investment . . . available only by direct factory order. SCR1000 . . .\$1150.00 (\$1065.00 w/o Preselector.) Commercial price somewhat higher.

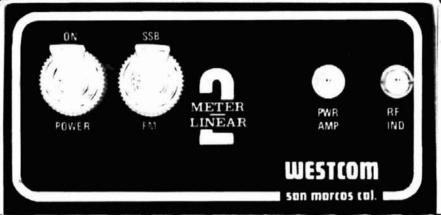
A Full Line of SCR1000 Repeater Boards & Complete Sub-Assemblies Are Also Available: Inquire.

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DIGS OUT THE WEAK SIGNAL **BOOSTS THE POWER**

Now Westcom gives you twice the advantage. ... a low noise receiver preamplifier and an output power amplifier, all in the same package! No modification of your transceiver is required since it's all in one high performance, low cost unit. The low noise U310 J.FET yields 12dB gain, 2dB NF, and the receive amp may be used independent of power amp. This unit is a natural for OSCAR uplink or long haul weak signal TROPO work. Available in 90w or 125w.

- An add-on unit, no internal connections or adjustments required to associated
- Standard Amplifiers operate FM. Linear Models operate all modes: SSB, FM, AM, RTTY, CW.
- emitter ballasting resistors achieve extreme ruggedness under severe operating conditions
 Withstands 20:1 VSWR under specified
- operating conditions
 "Microstrip" design provides high stability
 and optimum performance over wide band-
- Factory adjusted, no tuning required
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- Detachable fused power connector Conservatively rated with oversized heat
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 Red LED indicators for monitoring DC and
- One year material and workmanship limited

MODEL NO. (two meter)	INPUT POWER (watts)	MINIMUM OUTPUT W. (at max input)	MAXIMUM CURRENT 13.8 VDC	PRICE
FM Mode				
2m 15x70	5-15	70	8	\$119.95
2m 15x90	5-15	90	11	\$134.95
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All Mode-Linearized				
2m 15x70L	2-15	70	8	\$129.95
2m 15x90L	2-15	90	11	\$149.95
2m 25x125L	5-25	125	18	\$179.95
All Mode-Linearized with pre-amp				
2m 15-00B1	2.15	90	11	\$170 Q5

*Linear; AM, CW, FM, SSB, RTTY, Linear models work well with low power transmitters of 2-3 watts to yield 30-40w output. Size: 4 1/8 x 5 1/2 x 2 5/8

125

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tem allows easy and accurate message insertion with complete freedom of "pause" and "stop" placement. Additional 2102's can be added for almost unlimited memory storage (message length or quantity).

The 8047 operates from 5 Vdc and draws less than 10 mA of current. It is priced in single quantities at \$39.95. An 8047-1 kit containing the 8047, sockets, printed circuit board, 2102 memory, and manual is priced at \$69.95. For additional information, contact Curtis Electro Devices Incorporated, Box 4090, Mountain View, California 94040; or call (415) 964-3136.

TTL and CMOS logic probe

Heath Company, the world's largest manufacturer of electronic kits, has released the IT-7410/ ST-7410 Logic Probes, designed for in-circuit testing of TTL and CMOS integrated circuits. Features include switch selection of threshold levels for either TTL or CMOS circuitry and lamps that turn on when the input voltage crosses the appropriate level. A memory circuit is incorporated into the design of the unit to turn on an LED when either threshold level is crossed.

The new probes provide true logic level detection at high frequencies (not ac-coupled) and will detect pulses as short as 10 ns. Upper frequency limits are 100 MHz (TTL or CMOS with 5 Vdc squarewave) and 80 MHz (CMOS at 15 Vdc squarewave). Power for the Logic Probe is drawn from the circuit under test via two spring-loaded, insulated clips. A ground lead is provided for high-frequency operation. Probe overload protection is 50 Vdc continuous and 175 Vdc for five seconds. The IT-7410 is the kit version: the ST-7410 is the assembled version.

For more information about the new Logic Probes, which are mail order priced at \$39.95 kit and \$64.95 assembled, send for your free copy of the latest Heathkit catalog. Write

Heath Company, Department 350-690, Benton Harbor, Michigan 49022.

keyboard keyer IC

A single IC containing most of the electronics for a deluxe keyboard keyer has been introduced by Curtis Electro Devices. Called the 8045, the 40-pin CMOS device uses one or more FSC 3341s, a Curtis 8043 or 8044 keyer, and a set of keyswitches to produce the equivalent of the Curtis KB-4200 keyboard keyer, including an electronic paddle keyer. By adding the new 8047 Message Memory Control IC, 2102 RAM and a 4028 CMOS decoder, the equivalent of a KM-420 memory is added.

The 8045 creates a nominal fifty-seven position keyboard containing all the commonly used letters, figures, punctuation, space bar, and special characters (AA, KN, AS, SK, AR) all without shifting. It affords two-key rollover for "burst" typing, 32/64 character (or more) storage for smooth transmission, access to four message memories, analog output for buffer status meter, full and empty buffer indication, plus a preload function. It operates from +5 Vdc and requires less than 10 mA of supply current.

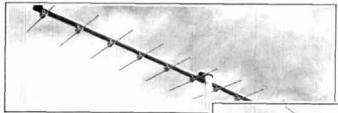
Priced at \$59.95 in single quantities, the 8054 is available from stock. A semi-kit (8045-1) containing the 8045, 3341, printed circuit board, sockets, and edge connector is priced at \$89.95. (The 8044-4 keyer semi-kit is priced at \$54.95.) For further information contact Curtis Electro Devices Incorporated, Box 4090, Mountain View, California 94040; or call (415) 964-3136.

desoldering wick

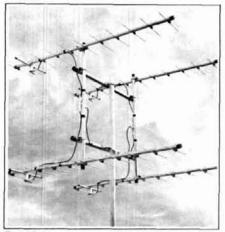
Chemtronics Incorporated has announced the latest addition to its popular line of solder and industrial chemical products, the D5 Desoldering Tool. This unique new product, which features Chemtronics' highly effective desoldering wick in a

CUSHCRAFT IS THE VHF-UHF ANTENNA COMPANY.

Cushcraft precision engineered VHF/UHF Yagi beams have become the standard of comparision the world over for SSB and CW operation on 6 meters through 432 MHz. Built by skilled craftsmen from the best available materials, these beams represent that rare combination of high electrical performance, rugged construction, and durability.

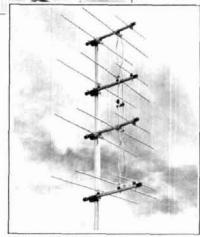


14-112-2 Meter Yagi



Quad Array

Cushcraft's Quad Arrays for 144, 220, and 432 MHz use four matched 11-element Cushcraft Yagis and are the ultimate in a high-performance Yagi array. These arrays have been carefully engineered for maximum forward gain, high front-to-back ratio, and broad frequency response. All antennas provide a low VSWR match to 50-ohm coaxial feedline.



20 Element DX Array

Cushcraft's wide variety of VHF/UHF
Beams includes an antenna for every
amateur activity above 50 MHz, whether
local ragchewing or long-haul over-thehorizon DX. All models have been
carefully optimized for maximum
forward gain with high front-to-back
ratio. The heavy wall bright hard-drawn
aluminum booms and elements are
combined with heavy formed aluminum
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hardware for long operating life and
survival in severe weather.



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SME



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.
*Touch-Tone is a registered trade name of AT&T.

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RAP-200 P. C. Card RAP-200R Rack Mount \$199.50 \$249.50



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2403 COMMERCE LANE ALBANY, GEORGIA 31707 912-883-4703 specially engineered, refillable dispenser, helps technicians desolder more efficiently while economizing on wick use. D5 may be used alone, or as an integral part of Chemtronics' new SD5 Solder/Desolder System.

The D5 Desoldering Tool consists of a 2.5-cm (1-inch) clear plastic cylinder which contains a visible supply of 1.5 meters (5 feet) of the company's specially formulated desoldering wick. Braid is fed to the work through a 6.4-cm (2-1/2-inch) Teflon probe that extends from one end of the wick supply. The heat-resistant Teflon probe allows you to desolder with pinpoint accuracy, even in highdensity circuitry. In addition, the D5 Tool's exclusive probe permits the user to shape or "web" the wick, providing maximum absorbency and further economizing on wick use. When the wick supply is exhausted, the user simply snaps the probe into the D5 Desoldering Tool Wick Refill.

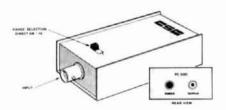
Chemtronics' pocket-sized D5 Desoldering Tool is available alone, or as part of the SD5 Solder/Desolder System. D5 wick refills are also available, allowing the D5 Tool to be economically reused for years. More information is available at Chemtronics distributors or directly from Chemtronics Incorporated, Solder Products Division, 45 Hoffman Avenue, Hauppauge, New York 11787.

DenTron's new MLA-1200 linear amplifier

The DenTron MLA-2500 now has a little brother, the new MLA-1200. Featuring super-compact design and a choice of ac or dc power supplies, the MLA-1200 is capable of 1200 watts PEP ssb service or 1000 watts on CW. A single Eimac 8875 allows long separation between the amplifier and power supply, plus minimum drive requirements of only 70 watts for the 1000-watt power level. The MLA-1200 is designed with the same state-of-the-art concepts that go

into the MLA-2500, and offers a budget-minded alternative for the amateur who doesn't need the 2000-watt continuous power level. The MLA-1200 is on your DenTron dealer's shelf right now, priced at \$399.50 for the amplifier, \$159.50 for the AC-1200 supply, and \$199.50 for the DC-1200 power supply. For more information write to the DenTron Radio Company, 2100 Enterprise Parkway, Twinsburg, Ohio, 44087.

CSC 500-MHz low-cost prescaler



Continental Specialties Corporation has just introduced an inexpensive 500-MHz prescaler capable of extending the performance of any 50-MHz frequency counter by ten times. It features a BNC input connector, diode-protected 50-ohm input, and 250-mV sensitivity from 50 to 500 MHz. Its output is a minimum 400 mV (peak-to-peak) capacitively coupled signal, available at a phone jack connector. Direct or divide-byten prescale outputs are switch selectable.

Power is supplied to the unit through a coaxial dc-type power connector. Power requirements are 7-12 Vdc at 100 μ A, maximum. An onboard voltage regulator assures trouble-free operation even from unstable power sources.

The entire PS-500 package is just $25 \times 50 \times 89$ cm (1 \times 2 \times 3-1/2 inches). Suggested resale in unit quantities is \$59.95, and CSC's full dealer margins apply. Available accessories include 110 and 220 Vac power supplies, each \$9.95; a power-connector to alligator-clip cable at \$2.95; a cigarette lighter power cord

Loop Antenna



Here is an exciting new device to improve your reception on 160, 80, the broadcast band, and on VLF.

It is well known that loops pick up far less noise than most other antennas. And they can null out interference. Now Palomar Engineers brings you these features and more in a compact, carefully engineered, attractive desktop package.

Unlike ordinary direction-finder loops, it tilts to match the incoming wave front. The result: Deep nulls up to 70 db. You have to listen to believe it!

Does the Loran on 160 give you a headache? The loop practically eliminates it. Broadcast station 2nd harmonic ruining your DX? Turn and tilt the loop and it's gone. Does your friend in the next block with his kilowatt block those weak ones? Use the loop and hear him fade out.

Loop nulls are very sharp on local and ground wave signals but usually are broad or nonexistent on distant skywave signals. This allows local interference to be eliminated while DX stations can still be heard from all directions.

The loops are Litz-wire wound on RF ferrite rods. They plug into the Loop Amplifier which boosts the loop signal 20 db and isolates and preserves the high Q of the loop. The tuning control peaks the loop and gives extra preselection to your receiver.

Plug-in loops are available for these bands:

150-550 KHz (VLF) 540-1600 KHz (Broadcast) 1600-5000 KHz (160 & 80 meters)

Send for free descriptive brochure.

Order direct. Loop Amplifier \$67.50; Plug-in Loop Antennas \$47.50 each [specify frequency band]. Add \$2 packing/shipping. Calif. residents add sales tax.

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operation - transmitting and receiving W t. LEGS THR 9 88.4 - connector - Model 777BU . . . \$54.95 80-40-20-15-10 bands 2 trap -- 102 ft. with 90 ft. RG58U - connector - Model 998BU . . \$49.95 40-20-15-10 bands 4 trap --- 54ft. with 90 ft. RG58U coax - connector - Model 1001BU . . \$48.95 20-15-10 bands 2 trap --- 26 ft. with 90 ft. RG58U coax - connector - Model 1007BU . . . \$47.95

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The TS-820S is the rig that is the talk of the Ham Bands. Too many built-in features to list here. What a rig and only \$1098.00 ppd. in U.S.A. Many accessories are also available to increase your operating pleasure and station versatility



160-10M TRANSCEIVER



TS-700SP 2M TRANSCEIVER

Super 2-meter operating capability is yours with this ultimate design. Operates all modes: SSB (upper & lower), FM, AM and CW. 4 MHz coverage (144 to 148 MHz). The combination of this unit's many exciting features with the quality & reliability that is inherent in Kenwood equipment is yours for only \$729.00 ppd, in U.S.A.

Guess which transceiver has made the Kenwood name near and dear to Amateur operators, probably more than any other piece of equipment? That's right, the TS-520S. Reliability is the name of this rig in capital letters, 80 thru 10 meters with many, many builtin features for only \$739.00 ppd. in U.S.A.



80-10M TRANSCEIVER



TR-7400A 2M MOBILE TRANSCEIVER

This brand new mobile transceiver (TR-7400A) with the astonishing price tag is causing quite a commotion. Two meters with 25W or 10W output (selectable), digital read-out, 144 through 148 MHz and 800 channels are some of the features that make this such a great buy at \$399.00 ppd. in U.S.A.

Send SASE NOW for detailed info on these systems as well as on many other fine lines, Or, better still, visit our store Monday thru Friday from 8:00 a.m. thru 5:00 p.m. The Amateurs at Klaus Radio are here to assist you in the selection of the optimum unit to fullfill your needs.

8400 N. Pioneer Parkway, Peoria, IL 61614 Jim Plack W9NWE — Phone 309-691-4840 at \$3.95: a three-foot BNC-to-BNC input cable at \$5.95; and a three-foot phono-plug to phono-plug output cable at \$3.95.

The PS-500 links directly with CSC's MAX-100 MHz frequency counter (\$134.95), and their new Mini-Max 50-MHz handheld frequency counter (\$89.95) to extend the range of either counter. In addition, it can be used with almost every counter available.

For additional information, contact Continental Specialties Corporation, 70 Fulton Terrace, New Haven, Connecticut 06509; phone (203) 624-3103.

improved 2-meter rig from Heath



Heath Company of Benton Harbor, Michigan, has made available an improved version of their HW-2036 frequency-synthesized 2-meter transceiver kit, the HW-2036A.

The HW-2036A has the same features and specifications as the HW-2026 except that the newer version allows operation on any 4-MHz segment of the transceiver's 143.5 to 148.5 MHz operating range. For those not already familiar with Heath's 2-meter rig, it features a phase-locked synthesizer/VCO loop for switch-selectable QSY operation, and a choice of simplex or standard plus or minus 600-kHz-split operation. An auxiliary switch lets the operator choose his own offset.

The synthesizer is locked to a precision 10-MHz time base. A NAND gate logic system displays locked/unlocked status and inhibits out-ofband transmissions by preventing transmitter key up. Other HW-2036A features include subaudible tone encoding, built-in 5 and 11 Vdc regulators, a hash filter/regulator, and a gimbal mount. A standard push-totalk microphone is included in the HW-2036A-2 kit at the mail-order price of \$269.95. When HW-2036A-1 (illustrated) is specified, the standard microphone is replaced by the HD-1984 Micoder II combination microphone/tone-button pad. The HW-2036A-1 sells for \$289.95. Both prices are mail order FOB Benton Harbor, Michigan.

For more information about the upgraded HW-2036A and a free catalog, write Heath Company, Dept. 350-640, Benton Harbor, Michigan 49022.

precision tuning devices

Two new series of miniature multiturn trimmer capacitors with Teflon dielectric have been added to the Tetfer line manufactured in Croydon, England. Type TPC trimmers have printed-circuit tags arranged for through-board mounting; the capacitor occupies an 8.4-mm (11/32") diameter hole cut in the circuit board, the tags being pushed into place from below. This minimizes protrusion above and below the board.

Type INS trimmers have the rotor insulated with a nylon extensionpiece to prevent the adjusting screwdriver from influencing the capacitance setting. Tetfer capacitors are cylindrical (brass-teflon-brass) with a circular ceramic base. The multi-turn screwdriver adjustment - each complete turn alters capacitance by about 1 pF — provides very fine tuning and exceptional stability. There are four models with capacitance swings of 5, 10, 15, and 20 pF. Existing versions plug into the circuit board from above, with a choice of horizontal or vertical adjustment.

New models of the Jackson Brothers TX5 vane-type transmitter-capacitors have also been announced. One

All Bands Preamplifier



- Improved reception. For receivers and transceivers.
- Continuously tuneable 1.8 54 MHz. Covers all amateur bands 160 to 6 meters; all shortwave broadcast bands.
- Low noise figure. Up to 20 db gain. Reduces image and spurious response.

Here is an exciting new device to improve your reception on all bands 1.8 to 54 MHz. It gives up to 20 db extra gain and the low noise figure of a dual gate FET to pep up your noisy receiver or transceiver. And it adds a high Q tuned circuit to give improved overload capability and image reduction. Makes that tired old rig come to life again. Even helps new models dig into weak signal territory.

Works with all transceivers up to 350 watts input power. A unique and improved circuit automatically bypasses the preamplifier when the transceiver transmits. The bypass delay is continuously variable by front panel control.

The low profile cabinet takes minimum operating table space. A heavy die cast case gives better shielding and isolation. size: 6" x 7" x 2" high. A built-in 117 volt AC supply and a connecting coaxial cable for the transceiver are included.

Order yours now! Price \$89.50 plus \$2 shipping/handling U.S. & Canada. Calif. residents add sales tax.

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ASTRON 35 AMP REGULATED POWER SUPPLY Model RS-35A

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Model	Continuous Duty (amps)	(amps)	Size (in.) H × W × D	Shipping Wt. (lbs.)	Price
RS-20A	16	20	5 × 9 × 9	19	\$94.95
RS-12A	9	12	4 × 8 × 9	13	\$72.95
RS-7A	5	7	31/4 × 6 × 71/2	71/2	\$49.95
RS-4A	3	4	31/4 × 6 × 71/2	5	\$39.95

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version has Teflon interleaves in the air spaces between the vanes to provide a higher voltage/capacitance ratio per unit volume. Another has the shaft mounted in roller bearings to reduce the drive torque. The capacitors are made with capacitance swings up to 1500 pF and in single-stator, split-stator, and differential versions. Working-voltage ratings vary from 2 kV to 6 kV.

10:1 Reduction Drive. The new type 6100 reduction drive introduced by Jackson Brothers is designed specifically for attachment to singleturn potentiometers. The 10:1 reduction ratio converts the component into a high-resolution multiturn potentiometer, and the combination costs much less than such a potentiometer. A British potentiometer manufacturer is currently supplying such combinations in quantity for use in consumer radio and TV receivers. The drive is a friction-operated epicyclic ball-drive with automatic protection against over-adjustment.

For more information, write to Swedgal Electronics, Inc., 258 Broadway, New York, New York 10007.

Cover Craft dust covers



Nothing says more about how you value your equipment than its appearance — and that's a direct result of how you treat it. Most amateurs are reasonably careful while using their equipment, but when the switches are off the matter is forgotten. Dust from cleaning, cigarette ashes, spilled coffee, bits of wire or solder — all can find their way into your gear almost any time, so why not protect it by covering it up?

In addition to their protective

value, Cover Craft dust covers enhance the resale value of equipment; cleaner rigs bring more as trade-in or flea-market items. Less maintenance is required on a clean rig, too.

Cover Craft covers are made of high-quality vinyl, with a special finish that lets the features of your equipment show through — you're not hiding your station when you protect it. Excellent construction is used; all edges and seams are machinestitched for durability and professional appearance.

All Cover Craft covers are made to precisely fit most popular makes and models of equipment. Projections and special features are allowed for when the material is hand cut and sewn. Cutouts at the appropriate places make it easy to cover your rig without removing any cables.

You can obtain Cover Craft covers for many types of gear: tape decks, hi-fi systems, microprocessors, test equipment, CB base stations, and, of course, all popular makes of amateur receivers, transmitters, transceivers, amplifiers, and accessories. They'll even custom-fit a cover for your own design if you'll tell them its size and shape.

Cover Craft dust covers — an investment in protecting your station. For prices and a list of equipment they'll fit, write to Cover Craft, Post Office Box 555, Amherst, New Hampshire 03031.

frequency counter kit



A versatile, high-quality, 600-MHz frequency counter which features accuracy, sensitivity, reliability, and low cost has been introduced by Davis Electronics. The counter is de-

PUNCH UP



- UP TO 18 TELEPHONE or control numbers retrieved with a one or two key punch.
- AUTOMATIC RE-DIAL of last number manually dialed.
- SAFE AUTOPATCH CALLS even in heavy traffic.
- 10 NUMBER RAM easily programmed in moments from the keypad.
- OPTIONAL PLUG-IN 8 NUMBER PROM custom-programmed by factory is available for \$4.95. A Prom Order Card is packed with each AD-1.
- PROGRAMMABLE TONE-LENGTH AND DURATION ensures accessing virtually any repeater having strict timing requirements. Ask any veteran autopatch user and he will tell you that this feature is an absolute must!
- EASY INTERFACING with virtually any amateur transceiver using the coil cord provided.
- CRYSTAL CONTROLLED TIMEBASE assures high stability over a wide temperature range.
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 All these features and more are possible thanks to the exclusive AEA 197701 MOS Microcomputer Chip. (OEM inquiries invited).



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- IMPROVES NOISE BLANKER OPERATION

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FR-101 SERIES KIT ... 34.50

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signed for 115 Vac or 12 Vdc operation, and is available either factoryassembled or in kit form. The Davis 7208 frequency counter incorporates the latest LSI (large-scale integration) technology in a wide-range, portable instrument measuring only $14 \times 15 \times 5 \text{ cm } (5-1/2 \times 6 \times 2)$ inches) and weighing a mere 0.79 kg (1-3/4 lbs).

Superior features of the Davis 7208 include a durable, all-metal cabinet for rf shielding, large 8-digit LED display, push-button switches, built-in prescaler, gate light, crystal timebase, and automatic decimal-point placement. Options available include a crystal oven, a rechargeable battery for total portability, and a built-in vhf-uhf preamplifier for direct measurement of low-level rf signals.

The Davis 7208 has a frequency range of from 10 Hz to 600 MHz, with 0.1 and 1.0 second gate time; resolution is 1 Hz with the 1.0 second gate, 10 Hz with the 0.1 second gate, and sensitivity is 10 mV at 60 MHz and 100 mV at 600 MHz. Input impedance is 1 megohm and 20 pF to 60 MHz; 50 ohms above 60 MHz. Time base accuracy is ±1 part-per-million for the standard package, or ±0.5 partper-million with the crystal oven.

The 600-MHz kit (model 7208K). which costs \$149.95, comes complete with all parts, drilled and plated through glass printed-circuit boards, cabinet, switches, hardware, plus a detailed assembly manual and calibrating instructions. Assembly time is about four hours. All parts are quaranteed for 90 days and factory service is available, if needed. A factory-assembled 600 MHz unit (model 7208A) costs \$199.95 (plus \$2.00 shipping); it is calibrated to specifications and guaranteed for one year. Optional (01) crystal oven is priced at \$39.95; (02) rechargeable ni-cad batteries are \$39.95; (03) carrying handle costs \$5.00; and (04) built-in vhf-uhf preamplifier is \$10.00. For further information contact Davis Electronics, 636 Sheridan Drive, Dept. 805, Tonawanda, New York 14150.

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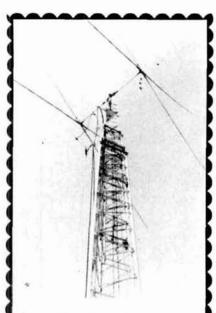
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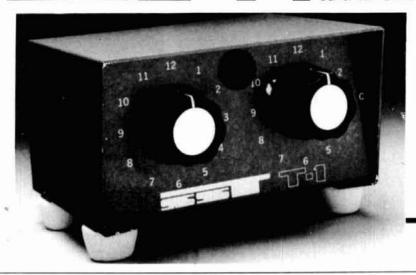
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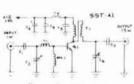


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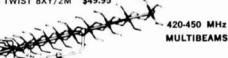


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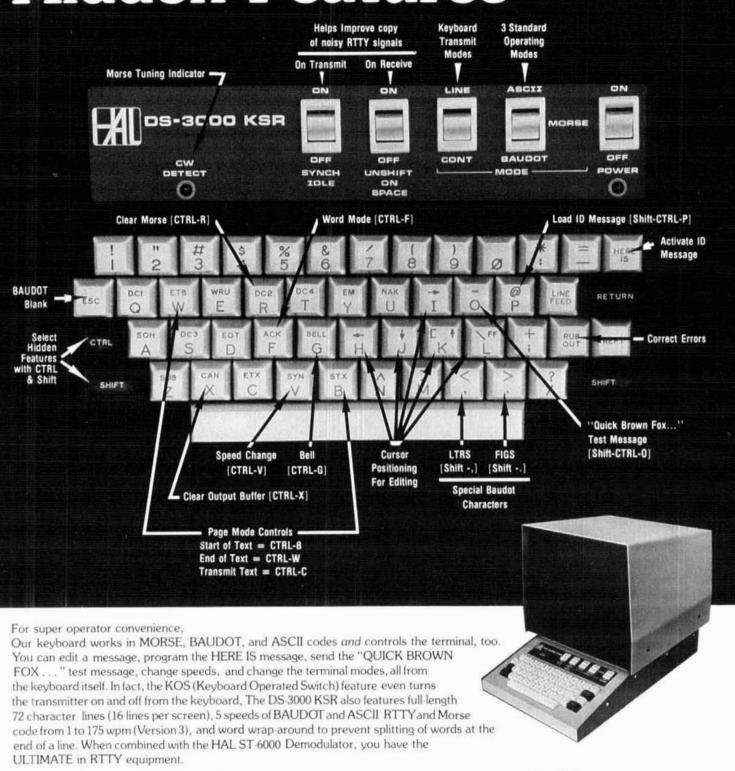


Produces upward and downward wail characteristic of police siren. 5 watts audio output, runs on 3-9 volts, uses 8-45 ohm speaker. Complete Kit, SM-3\$2.95

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YAESU AND KENWOOD FT-7, FT-101 FR-101, FT-301, TS-515, TS-520, R-599



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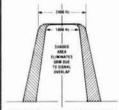
What can you do when two single sideband signals overlap? At present, not much in most older sets. If you are correctly tuned to one signal, and an adjacent one comes on, his high-or low-frequency components will be within your passband, and while you are not able to "read" him you will surely know he is there.

Many of the newer receivers seek to solve this overlap problem by providing continuously variable i-1 bandwidth or passband shifting - both difficult to adapt to existing designs. So, you can solve your problem by buying a new set! Not a pleasant prospect at today's prices

A simple, less expensive, but effective alternative is to sup-A simple, less expensive, our elective alientaive is to opportunities by the present right of the existing SSB filter in your tried-and-tested present right with our high-quality, 1800-Hz, 8-pole unit. Installation using our dual diode switch board is easy and permits the addition (now or in the tuture) of a second, sharp CW filter. Selection of the standard or sharp filters (SSB or CW) is achieved by flipping a single miniature toggle switch which can usually be mounted in an existing hole.

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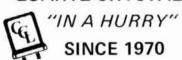
(without drilling) of up to two crystal filters of any type in addition to those for which the manufacturer provides space. These boards will accommodate any of the filters list-ed and other types planned for the future. To avoid error when ordering, specify filter number desired as well as rig with which it is to be used. Complete instructions. SPEC-IFY Set with which board is to be used. \$15 with purchase of any fil ter; \$20 without filter. Airmail Ppd US & Canada. Overseas add \$1

	Rig	Filter No. YF	Used	Center Freq. kHz.	No. of Poles	Band Width	Notes
YAESU SERIES	FT-101 FR-101	31H250 31H500 31F600 31H1.8 31H2.4 31F6.0	CW CW CW SSB SSB AM	3179.3 3179.3 3179.3 3180 3180 3180		250 Hz 500 Hz 600 Hz 1.8 kHz 2.4 kHz 6.0 kHz	Sharp unit for DX and contest work Use instead of standard 600 Hz unit Same as standard XF-30C unit, \$40 For narrow SSB Substitutefor XF-30A(6 pole) in early units Same as standard XF-30B unit, \$40
YAESI	FT-7 FT-301	89H250 89H500 90H1.8 90H2.4	CW CW SSB SSB		8 8 8	250 Hz 500 Hz 1.8 kHz 24 kHz	Sharp unit for DX and contest work Use instead of standard 600 Hz unit For narrow SSB For use in speech processor
000	TS-520 R-599	33H250 33H400 33H1.8	CW CW SSB	3395 3395 3395	8 8 8	250 Hz 400 Hz 1.8 kHz	Sharp unit for DX and contest work Use instead of standard 500 Hz unit For narrow SSB
KENWOOD	TS- 820	88H250 88H400 88H1.8	CW CW SSB	8830.7 8830.7 8830	8 8 8	250 Hz 400 Hz 1.8 kHz	Sharp unit for DX and contest work Use instead of standard 500 Hz unit For narrow SSB



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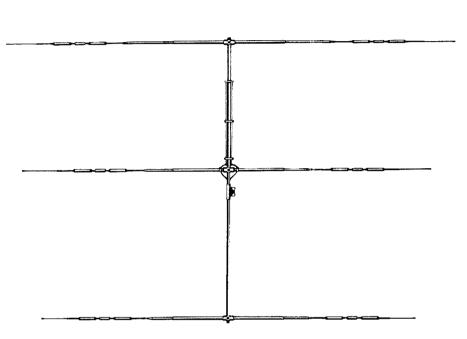




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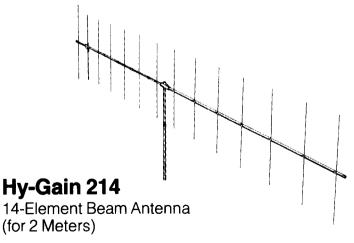
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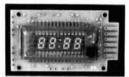
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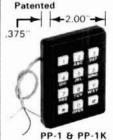
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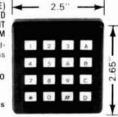
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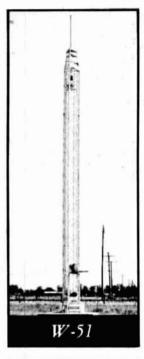
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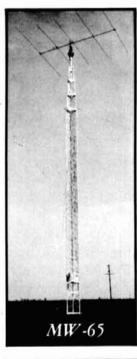
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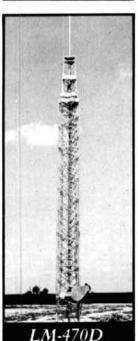
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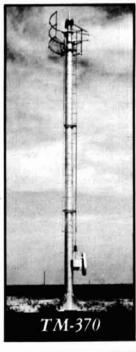
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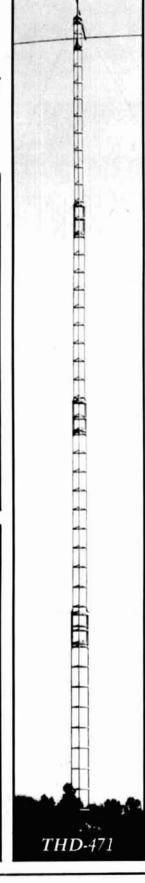
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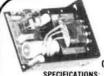
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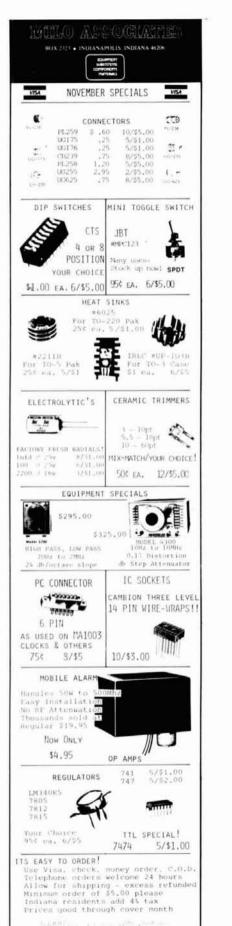
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FOR SALE: Pair of 4CX300A, one socket, all brand new, \$160.00. Pair of CTC-15 Transistors, 150W CW @ 30 MHz brand new \$80.00 Call Jack at (805) 498-6057 or write P.O. Box 964, Newbury Park, CA 91320.

TRYING TO LOCATE MARTIN PELKA who was W6 in 1970. Any information to: N7WT, 2835 E. Pershing, Phoenix, AZ 85032.

DRAKE T4XB, seldom used, mint w/manual \$400. Drake RCVR 2C, seldom used as backup, mint w/manual \$185. R. McKinney WB8SKX, (513) 562-4087 days or 3251 Hardisty, Cincinnati, 45208.

MILLER IF's, capacitors for HBR series, 3/8" ceramic forms. Johnson "U" variables, 13V/12A supply, misc. SASE list. K7PPC, 160 N. 6th, Laramie, WY, 82070.

ATV EQUIPMENT, older, with manuals; cameras, monitors, lenses, cable, etc. Send large SASE for list. Equipment Sale, Broadcasting Division, Mail Code #03, University of Cincinnati, Cincinnati, Ohio 45221.

Coming Events

NEW YORK: Binghamton Amateur Radio Association. W2OW (Fred Porterfield Memorial Station) celebrates its 50th Anniversary: 1928-1978 by announcing a CER-TIFICATE FOR CW QSO WITH W2OW, contest to end December 31, 1978. DX stations send 1 IRC and QSL to WA2IKO, Phil Horan, Box 342, RFD #2 Harpursville, NY 13787. US stations send #10 SASE with OSL to WB2GXX, Gina Ramsey, Box 334, RFD #2 Harpursville, NY 13787. Frequencies for DX stations first 50 kHz of 10, 15, and 20 meter bands; Sundays between 1000 and 1200 UTC. Frequencies for US stations 21.100 to 21.200 MHz; weekdays, 2300 UTC.

DELAWARE: QSO PARTY, November 11th and 12th between 0001 and 0600 UTC; and between 1600 and 2359 UTC both days. Stations may be worked once per band per mode for QSO points. Exchange QSO number, RS(T), and QTH (county for Delaware, and ARRL section or country for others). Frequencies: CW — 80 Hz above low end of 10, 15, 20, 40, and 80 meter bands; PHONE — 3900, 7275, 14325, 21425, 28650; NOVICE — 3710, 7120, 21120, 28160. Mailing deadline: December 15, 1978. Details from Sandy Cuccia, WB3ENF, 7 Sorrel Drive, Wilmington, Delaware 19803.

SOWP ANNUAL CHRISTMAS CW QSO PARTY, December 16th and 17th covering the full UTC period both days. No formal exchange requirements, just an op-portunity to meet on the air and exchange greetings. Call CQ SOWP (Society of Wireless Pioneers) between 50 and 60 kHz up from low end of each Amateur band. Novices, middle of each Novice band. Details from Bill Willmot, K4FT 1630 Venus Street, Merrit Island, Florida

SHERLOCK HOLMES AWARD: Sponsored by the German Section of the INTERNATIONAL POLICE ASSOCIA-TION RADIO CLUB (IPARC) open to all Amateur Radio Operators and SWLs. Period: both days — 0800-1000, 1400-1700, 1800-2000 UTC; Call CQ IPA, CW and SSB. Cross mode not allowed, crossband not allowed, Exchange: non-members: RS(T) plus serial number; members: IPA & RS(T) & serial number. CW: 3575, 7025, 14075, 21075, 28075 kHz; SSB: 3650, 7075, (foreign stations only), 14295, 21295, 28650 kHz. For complete details, write Vince Gambino, WB4QJO, 7606 Kingsbury Road, Alexandria, Virginia 22310. November 11th and 12th

MARYLAND: Mountain A.R.C. of Cumberland announces its SILVER JUBILEE CELEBRATION, starting 0000 UTC November 4th and ending 2400 November 5th. Stations may be worked only once, regardless of band or mode. No repeater contacts. Exchange RS(T) and QTH (state or country). Frequencies: 3540, 3910, 7040, 7240, 14040, 14295, 21110, 21360, 28110, 28600. Awards include special QSL for QSO with club station W3YMW, and special Certificate for QSO with five club members of M.A.R.C. Details from John P. Fanelli, Jr., WA3WSW, 609 Piedmont Avenue, Cumberland, Maryland 21502.

THE OAK PARK HIGH SCHOOL Electronics Club presents the 9th annual Swap 'N Shop on Sunday, November 26, 1978 at the Oak Park High School, 13701 Oak Park Blvd., Oak Park, Michigan 48237. Refreshments and door prizes. Donation \$1.50. Tables \$1.00 and \$2.00.

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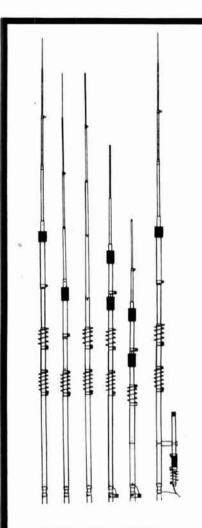
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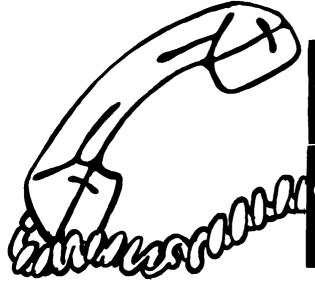




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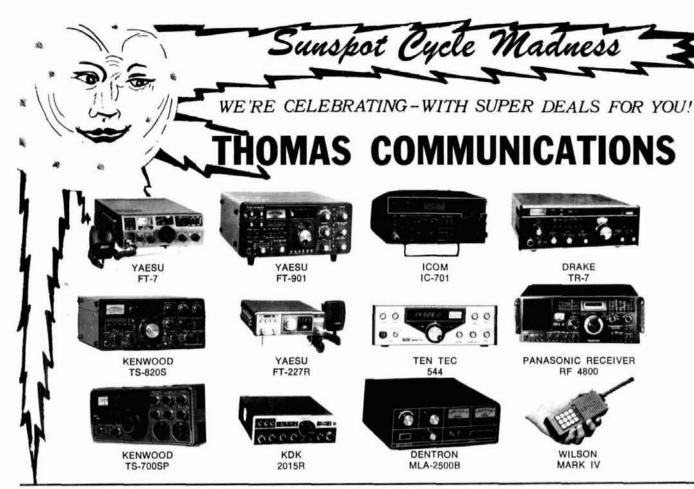
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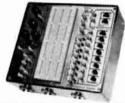
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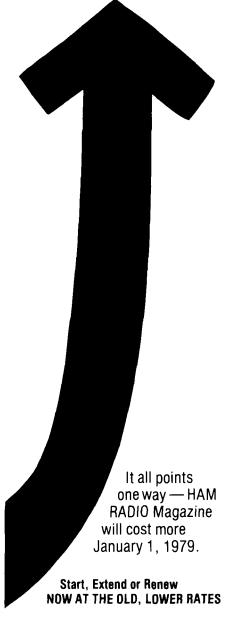
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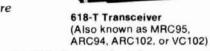
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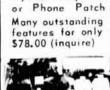
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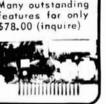
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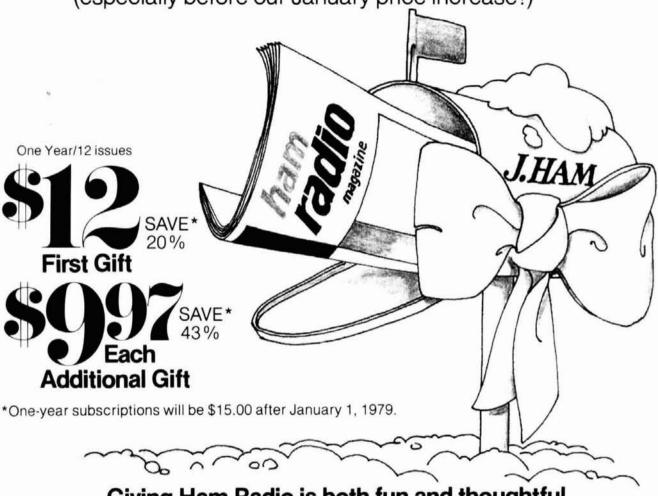
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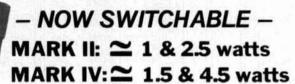
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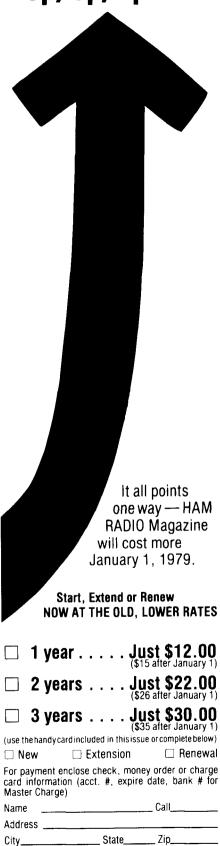
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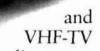
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