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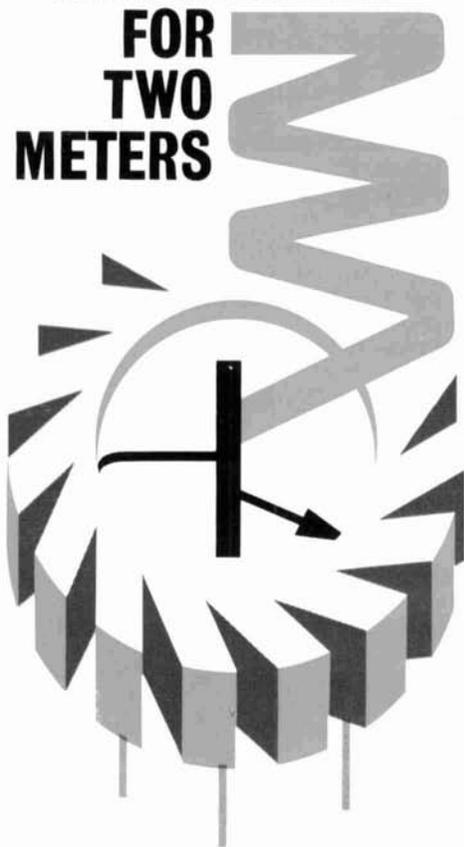
focus
on
communications
technology ...

ham radio

magazine

OCTOBER 1971

FM TRANSMITTER FOR TWO METERS



this month

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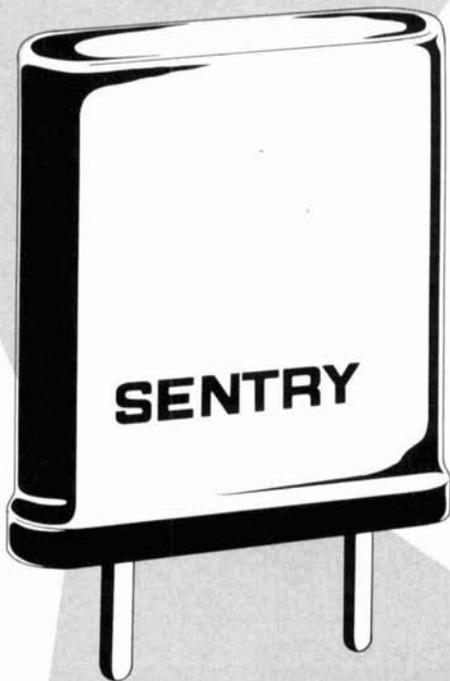
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 28.5 - 29.1 MHz
 29.1 - 29.7 MHz

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fm editor

Alfred Wilson, W6NIF
James A. Harvey, WA6IAK
associate editors

Jean Frey
art director

Wayne T. Pierce, K3SUK
cover

T. H. Tenney, Jr. W1NLB
publisher

Hilda M. Wetherbee
advertising manager

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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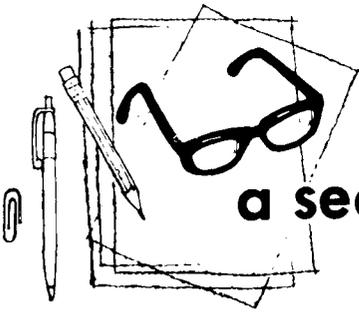
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a second look

by jim
fisk

Semiconductor microwave power devices — diodes that use drift time to generate large amounts of rf power — are finding their way into more and more commercial microwave equipment. Although these devices were predicted theoretically nearly fifteen years ago, it wasn't until 1963 that a practical solid-state power-generating device was actually built.

The Gunn diode or bulk-effect device is a simple chunk of n-type gallium arsenide that generates microwave power directly when a voltage is placed across it. When a constant voltage is applied to the semiconductor material the current through it fluctuates at an extremely rapid rate, although somewhat randomly. If the slice of semiconductor is less than about 0.005-inch thick the current no longer fluctuates randomly, but rises and falls in a cyclic way, generating microwave power.

The bulk-effect device is inherently broadband and frequency output is determined by the circuit in which it is used. Present maximum power output in the CW mode is 1 watt at 5000 MHz and pulse powers of 2000 watts at this frequency have been obtained in the lab.

Avalanche diodes are another source of microwave power. These diodes operate in two basic modes: Impatt and trapatt. Impatt (for IMPact Avalanche Transit Time) oscillators use the negative resistance that results from a combination of internal secondary emission and bunched current carriers that drift through the solid-state material and deliver rf power by causing external circuit current which is 180° out of phase with the applied voltage.

In the Trapatt mode (for TRApped Plasma Avalanche Triggered Transit) the diode operates in the Impatt mode at some high microwave frequency and the desired output is taken at a subharmonic of the Impatt frequency. Trapatt diodes can deliver 10 watts CW at 1000 MHz with about 60% efficiency. The much less efficient Impatt diodes operate at about 10% efficiency, but at much higher frequencies — up to 150 GHz.

Although the cost of off-the-shelf devices is still fairly high, some enterprising engineers have discovered that the simple rectifier diodes in your amateur gear may be uhf Trapatt oscillators in disguise. They found that ordinary Fairchild FD-300 rectifier diodes yielded 68-watt pulses at 630 MHz! Out of 100 FD-300s purchased for the experiment, 83 oscillated with greater than 35-watts *average* output. When three diodes were mounted in parallel they provided repeatable 395-watt pulses at 570 MHz; efficiency was a surprising 75%.

The Fairchild FD-333 rectifier, which has higher capacitance and greater breakdown voltage than the FD-300, has been made to generate a respectable 152-watt peak power at 630 MHz, so it's doubtful that these are isolated cases. There must be many other common silicon rectifiers that perform as well or better as microwave generators. If you are already using commercial microwave diodes in your uhf equipment, or have found other low-cost rectifiers that yield useful amounts of uhf power, I would like to hear about it.

Jim Fisk, W1DTY
editor

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FEATURING . . .

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- Digital synthesizer frequency control
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- AM, SSB, CW, RTTY, ISB
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SPECIFICATIONS • **Frequency Range:** 10 kHz to 30.0 MHz. • **Modes of Operation:** USB, LSB, CW, RTTY, AM, ISB. • **Frequency Readout:** Complete to 100 Hz on six NIXIE tubes. • **Frequency Selection:** 10 MHz, 1 MHz, 0.1 MHz switch selected; 0 to 0.1 MHz continuously variable. • **Frequency Stability:** Drift does not exceed 150 Hz in any 15 minute period with a temperature change of 7° C per hour over a range of 0° to 40° C. • **Sensitivity:** Less than 0.5 microvolt for 10 dB SINAD at 2.4 kHz SSB mode; Less than 1.0 microvolt for 10 dB SINAD at 6 kHz AM mode. • **Image Rejection:** Greater than 70 dB relative to 1 microvolt. • **Blocking:** Greater than 100 dB relative to 1 microvolt. • **Cross Modulation:** Greater than 90 dB relative to 1 microvolt. • **Intermodulation:** Greater than 80 dB relative to 1 microvolt. • **Opposite Sideband Suppression:** Greater than 60 db at 500 Hz into the opposite sideband. • **I.F. Bandwidth:** 6 kHz, 2.4 kHz, 1.2 kHz, 0.4 kHz; Selectivity @ -6 dB: 6 kHz, 2.4 kHz, 1.2 kHz, 0.4 kHz; @ -60 dB: 11.5 kHz, 4.3 kHz, 2.4 kHz, 0.8 kHz; Optional filters available for other bandwidths. • **I.F. Outputs:** 50 millivolts into 50 ohms at 1st I.F.; 5.05 MHz and 2nd I.F. 50 kHz. • **Automatic Gain Control:** Audio Output rises less than 3 dB for RF input change of 1 microvolt to 100 millivolts; Attack time 100 microseconds; Release time 750 milliseconds (Slow AGC), 25 milliseconds (Fast AGC). • **Antenna Input Impedance:** 10 kHz to 500 kHz, 1000 ohms; 500 kHz to 30 MHz, 50 ohms. • **Audio Output:** 3 watts at 5% maximum distortion into 3.2 ohm load; 1 volt into 600 ohm output line; 3.2 ohm unbalanced and two 600 ohm balanced outputs; ISB output is one of the two 600 ohm balanced outputs. • **Audio Hum and Noise:** Greater than 60 dB below rated output. • **BFO:** Derived from standard clock or variable over a ± 3 kHz range from front panel. • **Power Requirements:** 115/230 volts ± 10% single phase 50-420 Hz 15 watts; 12 or 24 VDC supply optional. • **Dimensions:** 5.25 in. H x 19 in. W x 15 in. D, (13.3 cm H x 48 cm W x 38 cm D). • **Weight:** 17 lbs (7.7 kg).

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skip tenney
W1NLB

Who won the Signal/One-Alpha 70 package and the many other fine prizes in *ham radio's* 1971 Sweepstakes? This question has been asked rather frequently of late since many readers failed to see the announcement of the top-prize winners in "A Second Look" in our August issue.

prize winners

Here are the complete results of the Sweepstakes:

grand prize winner

Signal/One CX-7
Alpha 70 Linear by ETO
Mr. John F. Longley, W2ANB

second prize winners

two-Meter solid-state transceivers
W2CNB
W4YPC
WB8DUO

The big moment finally arrives as Ham Radio's advertising manager, Hilda Wetherbee, draws the winning Sweepstakes entry.



third prize winners

W0EAD	W0EZE	K7HNV
W5PQA	WN5ZGC	W2ESH
WN1MRY	WB4OEX	K0GZW
WN1ONM	K5FRI	K0ARK
K8OSR	K9WTT	WN8HPO
W7HJR	WA4AUC	WA8VSJ
W7HJY	WN4PFR	WA8SXA
K7ZVA	WB4LOY	W8LPY
WA5PMY	WA4YTK	WA8ZJL
WA0TRF	W2WKL	WN4QYG
W1PEX	W4BRE	W6DZK
K9TTE	W4FES	WN6OBJ
WA7GUV	WA4FIG	K6BOY
WA0NEH	W4ZSK	W6JN
W4ICL	K4OG	WN6LAJ
WN5YBM	WA0QJK	WA6MQT
W9UCI	WN1KQB	

On behalf of all of us here at *ham radio*, we want to thank the many thousands of you who entered for your participation. Our only regret is that everyone could not be a winner — maybe you'll be the lucky one next time. Let's hope so.

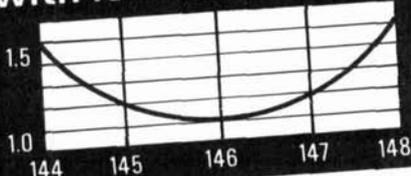
We are quite proud of this issue of *ham radio*. A check of our advertising index will show more advertisers in this issue than have appeared in any issue of any amateur magazine during 1971 (and probably for a good while before that). Although this is the result of a lot of hard work by our staff, it is also very much a credit to you, our reader. Your support of our advertisers has made *ham radio* a good investment for them. Keep up the good work!

Skip Tenney, W1NLB
Publisher

with **HUSTLER** VHF mobiles

you can be sure of **OPTIMUM GAIN** with lowest SWR . . .

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Heavy duty, precision fit, four section telescopic antenna tunable separately for combined six and two meter operation. Extends to 44". Collapses to 22".

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Model HF-62: Supplied with automotive eight ball base for fender or cowl mounting. Includes 60" RG-58-U and connectors. \$11.95

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140-500 MHz—
UHF/VHF Antenna
Low profile, roof top with factory attached cable—installs from outside without disturbing car header. 17-7 PH stainless radiating element—field trimmed for frequencies between 140 and 500 MHz. Complete with 15' RG-58-U coaxial cable.

Model UHT-1
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Easy to install on side or rear of trunk lip. No holes to drill or coax to trim. 17-7 PH stainless steel radiating element. Chart included for 140-500 MHz coverage. Complete with 17' RG-58-U and PL-259

Model THF
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2 METER POWER GAIN

143-149 MHz

3.4 db gain— $\frac{1}{2}$ wavelength
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Get the experience of solid communications, extended range and full quieting with the Hustler BBL models. Optimized gain performance with lowest SWR and the superior mechanical construction will give you the extra advantage. Both models supplied operational, ready for easy installation.

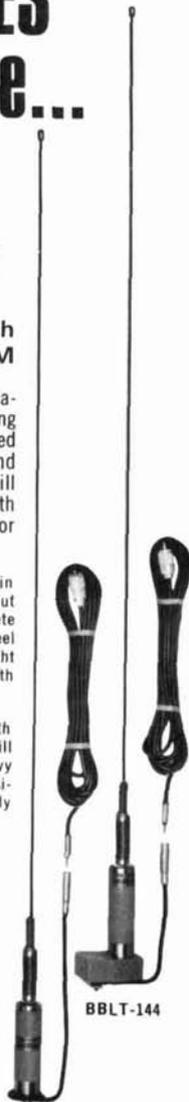
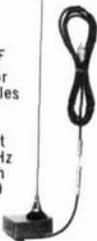
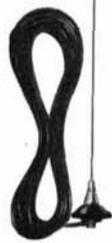
Model BBL-144: Mounts on any flat surface in $\frac{1}{4}$ " hole—easy installation on roof, (without pulling cars' header), deck or fender—complete with adjustable, taper ground stainless steel radiator, stainless steel spring. Overall height approximately 47". 17' RG-58-U coax with PL-259 connector. \$27.70

Model BBLT-144: Same as BBL-144 with Hustler trunk lip mount for no holes to drill installation on side or rear of trunk lip. Heavy duty mount assures "stay-put!" operation, positive RF ground for lowest SWR and completely hidden cable. \$34.70

Exclusive Hustler "break cable" assembly for simplest installation of all!



HF-62 HF-2



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vhf fm transmitter

A miniature,
solid-state design
that provides
low cost through
the use of
readily available
surplus Sonobouy
components

Although interest in two-meter fm has been increasing at an explosive rate, there have been few practical circuits published for high-performance vhf-fm transmitters. The solid state fm transmitter described in this article is designed for operation from 12-volt dc power supplies, provides 2-watts rf output and has excellent audio quality. Best of all, it may be duplicated easily and at low cost.

The advent of surplus Sonobouy transmitters,* and the ingenuity of some enterprising amateurs who discovered how easily they could be put on 2 meters, have resulted in a high quality 2-meter fm transmitter that is ideal for use in an fm base station, mobile unit or walkie-talkie. The 2-watt rf output has been more than adequate for working our repeater from 35 miles away using nothing but a $\frac{1}{4}$ -wave ground-plane antenna.

The original Sonobouy transmitter consists of two 4 x 12-inch printed-circuit boards (shown in photo) and is quite large and cumbersome. We decided to see what could be done about civilizing this

*Surplus Sonobouy transmitters are \$14.95 postpaid from Monks Electronics, 313 Old Farms Road, Simsbury, Connecticut 06070.

Robert Shriner, WA0UZO, Robert Heptig, K0PHF and Earl Gill, WB0EDA

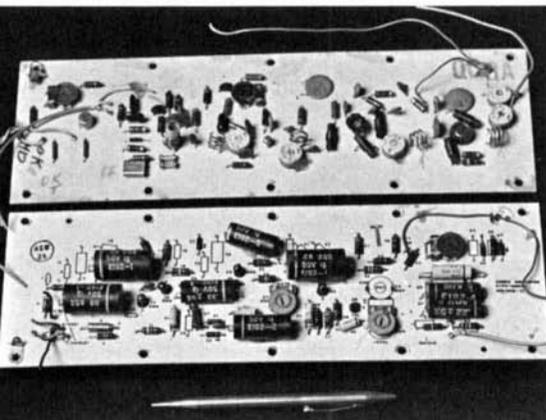
brute and reducing its size with a minimum of work and cost.

The audio section was the first to go; it simply had too many parts for the job it had to do. We decided to replace it with a simple two-stage circuit that had about the right amount of gain for 5-kHz deviation (see **fig. 1**). The new audio section uses an inexpensive field-effect input transistor (Q1) plus one of the transistors from the original Sonobaby audio board (Q2). With this circuit, deviation is controlled quite effectively by potentiometer R1. (Additional deviation may be obtained by increasing the value of C5 to 10 μ F.)

The original Sonobaby rf board appeared to use a normal amount of components and worked very well. However, the oscillator wasn't very stable after lowering the frequency from 20 to 18 MHz, so we used a new oscillator circuit that was based on proven designs. The new oscillator circuit is very stable and requires no adjustments or peaking—simply install a crystal.

The new 3 x 6-inch printed-circuit board shown in the photographs accommodates both the rf section and the audio section. This layout is the result of considerable parts shifting and several earlier boards; however, signal quality actually exceeds the quality of the orig-

Original Sonobaby transmitter board (top) and audio board (bottom).



inal. This transmitter, which we call the Sonobaby, has proven so popular that a parts list was developed that would allow amateurs not having access to a Sonobaby to build one from scratch.*

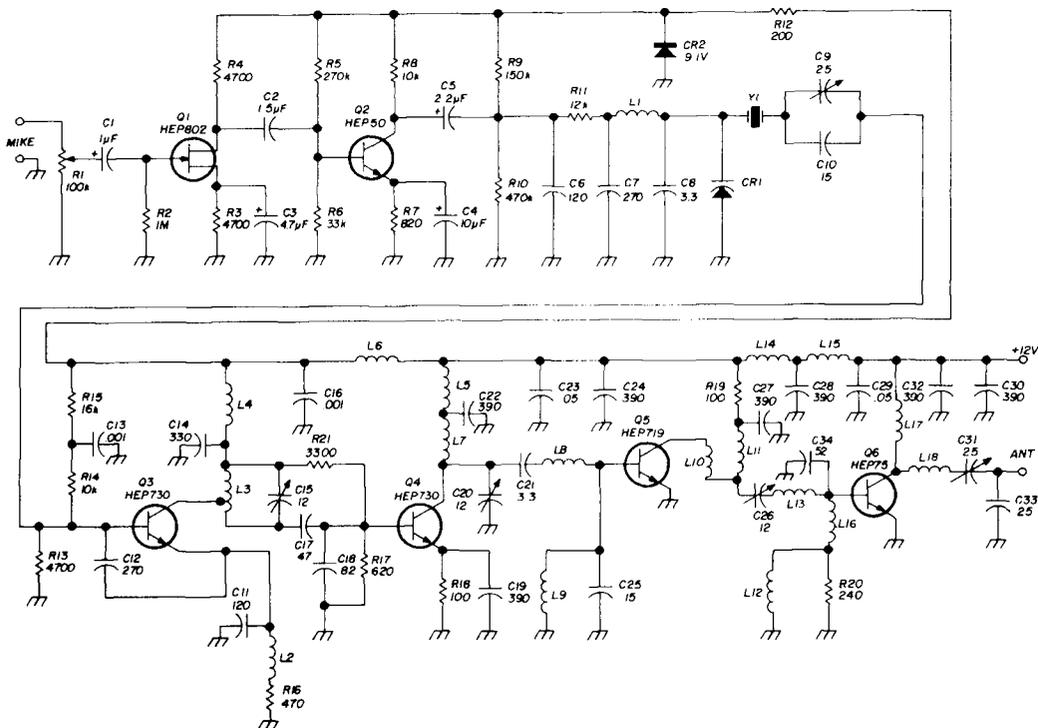


This Bearcat fm scanner monitor has been converted to a transceiver by installing a Sonobaby fm transmitter inside. Many vhf-fm receivers have been used in this manner to build complete transceivers and walkie-talkies.

*An etched and drilled epoxy circuit board can be purchased for \$7.50 from Sonobaby, Post Office Box 92, Pueblo, Colorado 81002. A complete parts list, plus changeover information from the original Sonobaby transmitter and tuneup procedure is included with each board.

A semiconductor package for the Sonobaby vhf fm transmitter is available at a special price from Circuit Specialists. The transmitter package, consisting of Q1, HEP802, Q2, HEP50, Q3 and Q4, HEP730, Q5, HEP719, Q6, HEP75, CR1, MV2101, and CR2, HEP104, is priced at \$9.05. A semiconductor package for the power supply, consisting of Q1, HEP624, CR-CR4, HEP175, and CR5, HEP605, is \$5.00 postpaid. Order from Circuit Specialists Company, Box 3047, Spotsdale, Arizona 85257.

A complete kit of parts for the Sonobaby, including circuit board and all components (less power supply) is priced at \$47.50 complete, plus shipping, from HAL Devices, Post Office Box 365H, Urbana, Illinois 61801. This parts kit includes an rf detector and 1-mA meter for use in tuneup.



- | | | | |
|--------|--|-----------------|---|
| C1 | 1 μ F electrolytic
(Mallory TAC 105M020P02) | C9,C31,C33 | 5/25 pF npo trimmer
(Erie 557-000A-5-25) |
| C2 | 1.5 μ F electrolytic
(Mallory TAC 155M020P02) | C10 | 15 pF Disc
(Sprague 5GA-Q15) |
| C3 | 4.7 μ F electrolytic
(GE 76 F-02KC4R7) | C13,C16 | .001 μ F Disc
(Sprague 5GA-D10) |
| C4 | 10 μ F electrolytic
(GE 76 F-02FC100) | C14 | 330-pF Disc
(Sprague 5GA-T33) |
| C5 | 2.2 μ F electrolytic
(Mallory TAC 225M020P02) | C15,C20,
C26 | 3-12 pF npo trimmer
(Erie 557-000A-3-12) |
| C6,C11 | 120 pF Disc
(Sprague 5GA-T12) | C17 | 47 pF Disc
(Sprague 5GA-Q47) |
| C7,C12 | 270 pF Disc
(Sprague 5GA-T27) | C18 | 82 pF Disc
(Sprague 5GA-Q82) |
| C8,C21 | 3.3 pF npo Disc
(Sprague 10TCC-V33) | | |

fig. 1. Schematic diagram of 2-watt vhf-fm transmitter.

the circuit

In the circuit in fig. 1 a field-effect transistor, Q1, is used as an audio pre-amplifier. Transistor Q2 further amplifies the audio signal and passes it on to the varactor diode, CR1. The capacitance of CR1 changes with the audio signal, varying the frequency of the crystal-controlled oscillator, Q3. The zener diode,

CR2, regulates the voltage to the audio amplifier and provides constant bias to the varactor.

The tank circuit of the 18-MHz crystal-controlled oscillator stage is tuned to 73 MHz; transistor Q4 doubles this signal to 146 MHz. Transistor Q5 drives the power amplifier, Q6, to 2 watts output on 2 meters.

C19,C22,	390 pF Disc
C24,C27,	(Sprague 5GA-T39)
C28,C30,C32	
C23,C29	0.05 μ F Disc (Centralab DD503)
C25	15 pF npo Disc (Sprague 10TCC-Q15)
C34	52 pF Disc
CR1	MV 2101 varactor (Motorola) or Eastron VC2101
CR2	9.1 V zener (Motorola HEP104)
Y1	Crystal, order from International Crystal Company for Westinghouse Air Brake. Carry Phone II 20TS-1 Transmitter; specify operating fre- quency and commercial Standard
L1	15 μ H choke (J. W. Miller 9310-40)
L2,L12	10 μ H choke (J. W. Miller 9310-36)
L3	5 turns no. 16, 15/16" ID tapped at 1 1/2 turns
L4,L5,L14	1.5 μ H
L15	(J. W. Miller 9310-16)
L6,L9	4.7 μ H (J. W. Miller 9310-28)
L7	3 turns no. 16, 15/16" ID
L8	7 turns no. 22, closewound on 3/16" plastic rod
L10	7 1/2 turns no. 22, closewound on 3/16" plastic rod
L11	5 turns no. 22 on 3/16" plastic rod, 5/16" long
L13	4 turns no. 16
L16,L17	12 turns no. 22, closewound on 3/16" plastic rod
L18	5 turns no. 16, 5/16" ID

parts; this prevents excess heat from damaging them. Tag each of the semi-conductors so you know what they are, and put them on a shelf out of the way as they should be the last parts you install on the new board.

For greater efficiency and power output rewind the coils as noted in the parts list. If you carefully examine the existing Sonobouy coils you will find that several of the original coils have the correct number of turns if they are moved to another location.

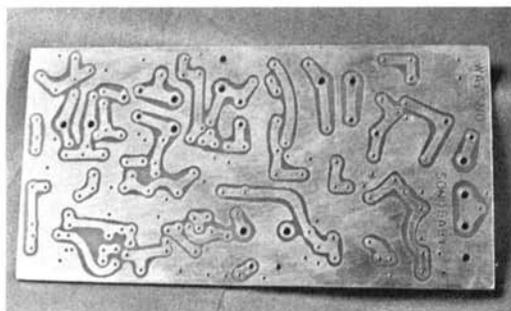
The coils wound on plastic rod stock are most easily made by first drilling two holes in the plastic rod the correct distance apart. The coils are then wound, passing the ends of the wire through the holes.

After all the components have been installed on the new printed-circuit board inspect it thoroughly for bad solder joints and short circuits before applying power. Connect some sort of indicating dummy load across the output (a wattmeter or a number-47 bulb), put a 500-mA meter in the 12-volt supply line and apply power. If you made no mistakes the unit will draw approximately 20 mA untuned; if it goes up in smoke you didn't inspect it carefully enough after installing the parts on the new circuit board.

tune up

When the transmitter has passed the first smoke test you can proceed with the tuning. First, tune C15 for maximum

Foil side of the printed-circuit board for the Sonobaby vhf-fm transmitter.



construction

To build the Sonobaby from the original Sonobouy, simply remove the components from the original circuit boards and install them on the new Sonobaby circuit board. A few small parts must be purchased; notably the field-effect transistor and the 9.1-volt zener diode. It's a good idea to unsolder the transistors and the varactor before removing the other

current through the 500-mA meter. Then, with a wavemeter (or grid-dipper in the diode position) tuned to 73 MHz and coupled to L3, adjust C15 maximum indication.

low to be measurable, is estimated at 10 milliwatts.

To put the Sonobaby on the air, connect a crystal or ceramic microphone to the input and adjust R1 for the best

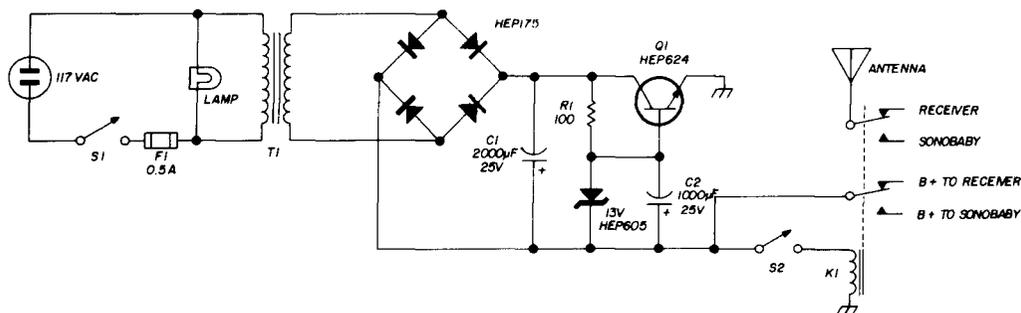


fig. 2. Power supply for the vhf-fm transmitter. Transformer T1 is a 12.6-volt filament transformer rated at 1 amp. Dpdt relay, K1, has 100-ohm coil.

The remainder of the alignment procedure is very straight forward: tune C20, C26, C31 and C33 in sequence for maximum rf output. Keep a check on the transistors during tuneup to make sure they don't get too hot. It is normal for them to be warm to the touch; if you can fry eggs on them, shut the power off then let them cool off before proceeding.

Capacitor C9 can be adjusted to put you precisely on frequency. If you have difficulty getting on frequency the value of C10 can be varied or the bias voltage set up by R9 and R19 can be changed. For best operation the bias voltage should be from +4 to +6 volts.

For the final step, couple an indicating wavemeter to the output and retune all capacitors for maximum indication. Check the second harmonic (292 MHz) to make sure it is well suppressed; if it is not touch up the tuning capacitors.

In the interest of good amateur practice (and conserving electricity) the value of R18 can be increased to as much as 1000 ohms for reduced power output. WA0UZO, who happens to live close to the repeater, has operated a Sonobaby with the final transistor removed and the antenna connected to L16. The circuit only draws 50mA, and output power, too

sounding audio. To observe the audio signal connect an oscilloscope to the collector of Q2.

transistor substitution

The bias resistors and capacitors shown in the schematic are correct for transistors salvaged from a Sonobaby transmitter. However, any substitution of transistors will likely require bias adjustments on that particular circuit.

Bias adjustment is very easy with a resistor substitution box. The procedure is simple: start at the crystal oscillator stage and adjust the bias resistors until the oscillator is putting out the desired signal; then move one stage at a time toward the final, adjusting each for maximum output.

After each change in capacitor value check for spurious and harmonic outputs. In some cases you may notice that a capacitor can be completely removed from the circuit for increased power output. Use extreme care — all capacitors in the circuit have a job to do; removing one is asking for trouble.

For example, if the 52-pF capacitor from the base of the final transistor to ground is removed, a wattmeter will indicate a decided increase in power

output. However, a careful analysis of the output signal will show that the power is all in spurious signals and harmonics. In this case I would recommend that this capacitor be changed to 30 or 40 pF and the output analyzed again.

power supply

The regulated power supply in fig. 2 has proven to be very satisfactory for operation of the Sonobaby and a solid-state whf-fm receiver. The parts are not at all critical — use whatever you have on hand. The transformer should be rated at about 1 ampere; the 13-volt zener diode can be taken from the audio board of the original Sonobaby transmitter. The transistor, Q1, should be rated at 5 or 10 watts. A dpdt relay with a coil resistance of about 100 ohms completes the system.

conclusion

The Sonobaby can be used as a miniature base station or it can be built into a walkie-talkie. There are many compact vhf-fm receivers on the market which are sold as police monitors; they can be coupled with the Sonobaby fm transmitter to provide a complete vhf-fm transmitter, and you can be on the air enjoying the benefits of 2-meter fm.

ham radio



"You say you're snowbound in a blizzard about twenty miles north of Notrees, Montana? Well, listen, I hate to interrupt, but . . ."



DIODES

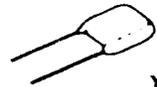


PIV	TOP-HAT		STUD-MOUNT	
	1.5 AMP	1.5 AMP	3 AMP	6 AMP
50	.04	.06	.12	.15
100	.06	.08	.16	.20
200	.08	.10	.20	.25
400	.12	.14	.28	.50
600	.14	.16	.32	.58
800		.20	.40	.65
1000		.24	.48	.75

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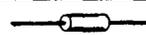


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direct-reading capacitance meter for electrolytics

This simple adapter
turns your vtm
or vom
into a
direct-reading
capacitance meter
for electrolytic
capacitors
up to 1000 μF

Some years ago I converted some transistor testers to give me direct readout of *beta* on the resistance scale of a vtm or vom. This was a real time saver at home and on the job and provided accurate readings over two decades without any switching.¹ Since then I have been looking for a way to read voltages on the resistance scale. Imagine being able to read from 1 to 100 volts on the same scale with reasonable accuracy! Although I haven't found a way to do that it became apparent that capacitance could be read this way.

The required capacitance scale is the same as the standard resistance scale. My other capacitance meters (I've had two) cover values between 10 and 1000 μF and could not be modified to do so. The adapter I built for measuring large values of capacitance is quite simple (see fig. 1).

Fig. 2 shows a version for volt-ohmmeters which works well and is for those of you who don't own a vtm.

construction

If you have a vtm use the circuit of fig. 1. Volt-ohmmeters won't be as accurate since they load the circuit more and are not quite linear on ac. A 5000-ohm-per-volt ac meter is all right; a 1000-ohm-per-volt meter is usable but not accurate. In any case, the most critical part is the capacitor you use for a standard (C_S in the diagrams). Use a good quality 5% or better tantalum capacitor, even if you have to buy it.

The value of C_S must be equal to, or a decade multiple of, the exact midscale ohms reading. This corresponds to half-scale reading on any dc scale. The value you select will determine what center scale reading represents in microfarads. There's nothing wrong with making it a two-range unit if you wish. I used a 100 μF standard since my vtm reads 10 at midscale; with this setup I can read capacitor values from 10 to 1000 μF very well.

The transformer can be a small filament type. Use as low a voltage (between 2 and 6 volts) as you can to match a 2-, 3-, or 5-volt ac range on your meter.

The resistance of the calibration pot should be selected experimentally for your particular meter. You should easily be able to set full scale with it. The calibrate switch is used only on the vtm version since the vom version will read full scale whenever the test (C_X) terminals are open.

The remaining parts may or may not be necessary; let me explain their purpose and you can decide. C_1 is a blocking capacitor. Most vtms have one built in, in which case you don't need it. Voms might read correctly without it; one of mine does. C_2 is an rf bypass; it is needed mostly on sensitive meters or if you are near a broadcast station or have your transmitter on.

The diode CR1 can be any small silicon diode or rectifier. It insures that a dc voltage will build up across the test

Herb Schoenbach, W9DJZ, 8 S. 275 Winwood Way, Apt. 6, Downers Grove, Illinois 60515

capacitor. It is not absolutely necessary because the normal rectifier action of electrolytic capacitors will do this to some extent. Readings are almost identical with or without the diode, but including it might make you feel better. At low voltages most electrolytics start to act like non-polarized devices; if you intentionally reverse polarity you get nearly the same reading.

I won't tell you how to put these parts in a box. There are enough good tips on construction practices in other articles — just go to it.

operation

Here are some tips on using this gadget. Remember to set and connect your meter as an ac voltmeter on about a 3-volt range. *Do not* set to a resistance or ohmmeter position. But when you make your capacitance measurements read the capacitance value on the resistance (ohms) scale.

Check your full-scale calibration often if your line voltage fluctuates. If you use a 6-volt transformer tap do not test

capacitors with less than a 10-volt rating. If you use a 3-volt tap you can safely test capacitors rated at 6 volts.

You will find that many electrolytics read almost double their marked value

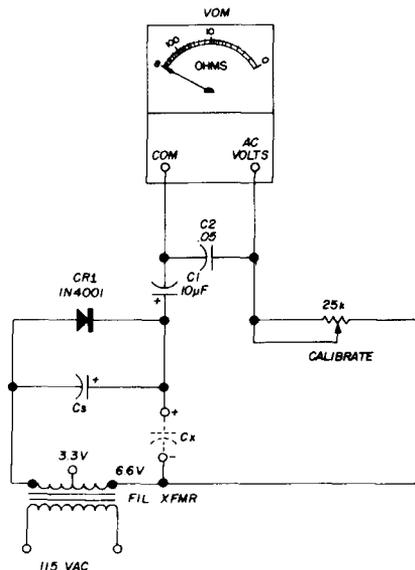


fig. 2. Capacitance meter adapter for voltmeter which has ohms increasing down-scale. To calibrate the adapter set the calibrate pot for 5 Vac full-scale with capacitor Cx out of the circuit. Cs is a 5% tantalum capacitor, 10 working volts or more (see text).

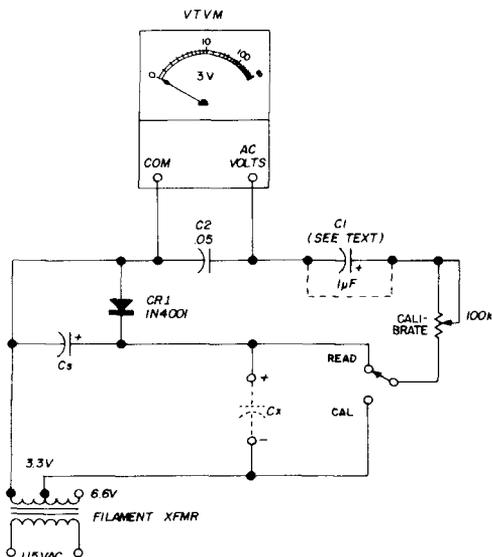


fig. 1. Capacitance meter adapter for a vtvm which has an ohms scale which increases up-scale. To calibrate the adapter, push the calibrate switch and set the calibration pot for 3 Vac full-scale. Cs is a 5% tantalum capacitor, 10 volts or more (see text).

when new. This is normal as capacitor tolerances are commonly plus 90%, minus 10%. High-voltage units may not read correctly unless they have been *formed* by operating them at their normal voltage for some time prior to testing.

Check unknown capacitors for shorts and leakage before trying to determine their capacitance. You will find leakage as low as 10,000 ohms does not affect accuracy much. You will want to check various known-value capacitors to be sure your adapter is working properly and to build up confidence in it. Even I was skeptical at first.

reference

1. Herb Schoenbach, W9DJZ, "Two More Transistor Testers," 73, September, 1969, page 133.

ham radio



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cw processor

for communications receivers

Frequency modulating
the telegraphy signals
in your receiver
provides an interesting
and profitable
addition to
conventional
receiver design

There's no doubt that a properly designed cw receiver — of the current basic design — leaves little to be desired in terms of plucking out weak signals, providing the *feel* of the band, and allowing the character of the other man's transmitter to come through. But what about those times when impulse noise is heavy and you would like to have compression without distortion? The technique described here is something a brass-pounder can *add* to his present receiver to become a discriminating man! The system operates on received signals to give them the same character they would have if they had been tone frequency modulated. Detection is set up to demodulate an fm signal. In this way the advantages of an fm system are obtained without the necessity of transmitting fm.

Don E. Hildreth, W6NRW, 936 Azalea Drive, Sunnyvale, California 94086

When impulse noise is present the results of fm are well known. And when tuning across the band looking for small signals — or just general tuning — it's a relief to be able to tune across a strong signal without having it take off the top of your head. Finally, it is nice to have a choice, and the unit shown here makes it possible to quickly choose between three modes of operation:

1. Conventional bfo
2. Fm with bfo
3. Fm with fixed tone

To simplify the description a unique detection system will be shown; then the idea will be integrated into a conventional communications receiver.

basic theory

Consider a conventional receiver (bfo turned off) tuned to a cw signal blasting out a series of di-di-di-dahs. The signal can be heard going on and off, but there is no audio beat note until the bfo is turned on. Now, instead of turning on a bfo, suppose that a different method is used to provide the audio tone.

A block diagram of this different approach is shown in **fig. 1**. In this system the received signal is frequency modulated just like an rf carrier would be in a transmitter. **Fig. 2** expands the frequency modulator block to give you an idea of how fm is accomplished. This is the same basic technique used by Armstrong in the 1930s, and is often used to produce narrow band fm.

The incoming signal is split into two channels. One of the channels is fed through a double-sideband suppressed-

carrier device, phase shifted 90 degrees, and added to the other channel. The relative signal strength in each channel determines the degree of modulation.

The 90-degree phase shift could be

tend to block small-signal inputs. To avoid that problem sufficient gain is used ahead of the multiplier to get the noise background up to at least 10 millivolts at the input to the multiplier. A fullwave

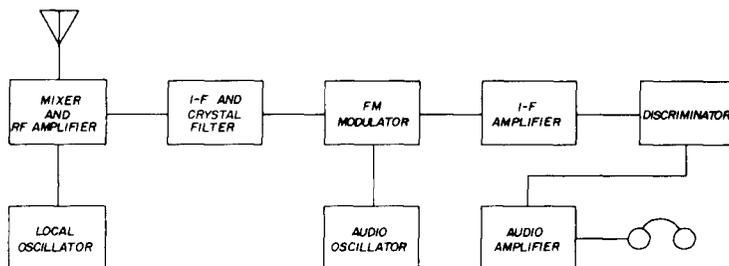


fig. 1. Basic system for frequency modulating telegraphy signals in a communications receiver.

placed in either channel — but not both. If no phase shift were present the system would simply supply amplitude modulation. In either case the detected tone is identical with what is fed into the modulating port of the double-sideband suppressed-carrier circuit.

the circuit

To a radio amateur the sound of a receiver like **fig. 1** is almost as sterile as a code-practice oscillator, so let's bear with a little more complication to get the *feel* back. The arrangement shown in **fig. 3** provides the three modes of operation and may be easily added to an existing receiver.

When the modulator is excited by the receiver's normal bfo, the *feel* is back! Then, for maximum quiet and/or very weak signals, the modulator can be switched to the input to the audio oscillator input. **Fig. 4** shows a schematic of the complete system.

At this point you may note that a considerable amount of energy is left in the carrier with sideband information some 6 dB down. To reduce that problem a final system should multiply the frequency of the signal before running it through the discriminator.

frequency multiplier

Conventional frequency multipliers

rectifier circuit with forward bias on the diodes reduces diode dead space to enable frequency multiplication to start with very small signals. (If you have worked with nbfm you are well aware that frequency multiplication will increase the modulation index, thus improving threshold signal detection.) Although more multiplication than is shown here would be beneficial for an optimum system you could obtain an improvement of only 2 or 3 dB at best.

discriminator

I chose a Travis discriminator circuit because it does not require a special transformer — a problem in this case because of the odd frequency. (A Travis discriminator could be thought of as a "push-pull staggered" slope detector.) With careful construction and adjustment of the balance potentiometer a-m rejection is reasonably good.

The easiest discriminator tuning procedure is as follows (refer to **fig. 4**): Energize the system with a strong, steady carrier at the center i-f frequency. Peak L1 while looking at the rectified dc voltage at diode CR1. Note the peak voltage; then tune L1 to raise the circuit frequency until the voltage is one-half of peak. Repeat the same procedure (opposite polarity) with L2 while looking at the voltage at diode CR2 but offset this

resonant circuit to the low side of peak voltage.

Assuming an i-f selectivity such as that recommended under *receiver requirements* is used the discriminator will be

primarily to hold shielding requirements to a reasonable level. It should be pointed out that the second harmonic of the i-f is in the broadcast band, so care should be taken to prevent pick-up from the local

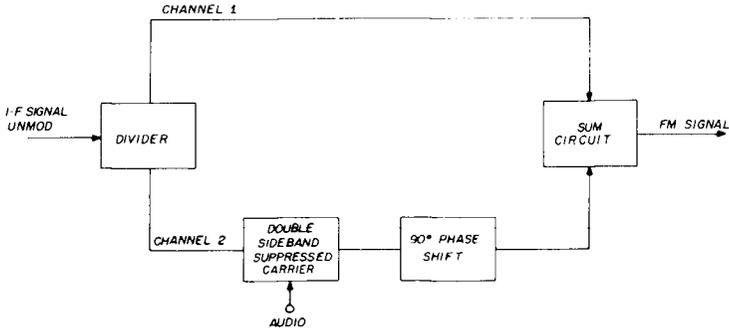


fig. 2. Basic phase modulation system. The rf input signals to the summing circuit should be approximately equal when full audio modulation is present.

relatively wide. This avoids the spurious detection on each side of a discriminator's S-curve that is sometimes annoying in tuning fm signals.

An fet was used at the input to avoid discriminator loading. This stage is followed by just enough transistor gain to get reasonable headset volume. The cir-

broadcast station. Normal point-to-point wiring is completely adequate for this circuit.

receiver requirements

While this system will function when used with a receiver without good i-f skirt selectivity the performance would be

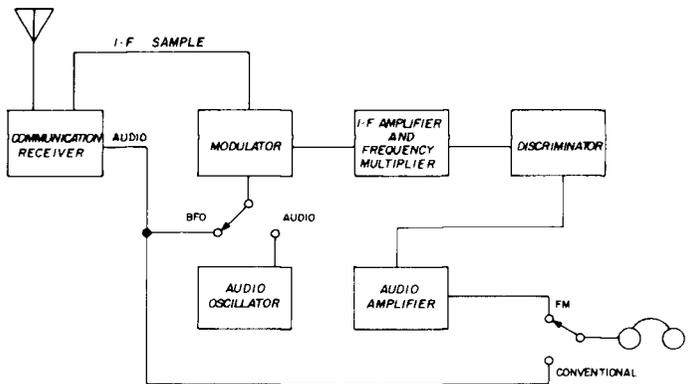


fig. 3. Signal processor is easily added to an existing receiver to provide new as well as conventional modes of cw reception.

cuitry shown in fig. 4 will easily drive an audio power stage if speaker operation is desired.

All of the circuits used in this system are rather conservative, with minimum use of LC elements. This was done

disappointing. To work well this system should be coupled to a receiver with a good crystal or mechanical filter. Present indications are that an i-f bandwidth of 200 to 400 Hz, with skirts only twice that wide 60-dB down, will give excellent

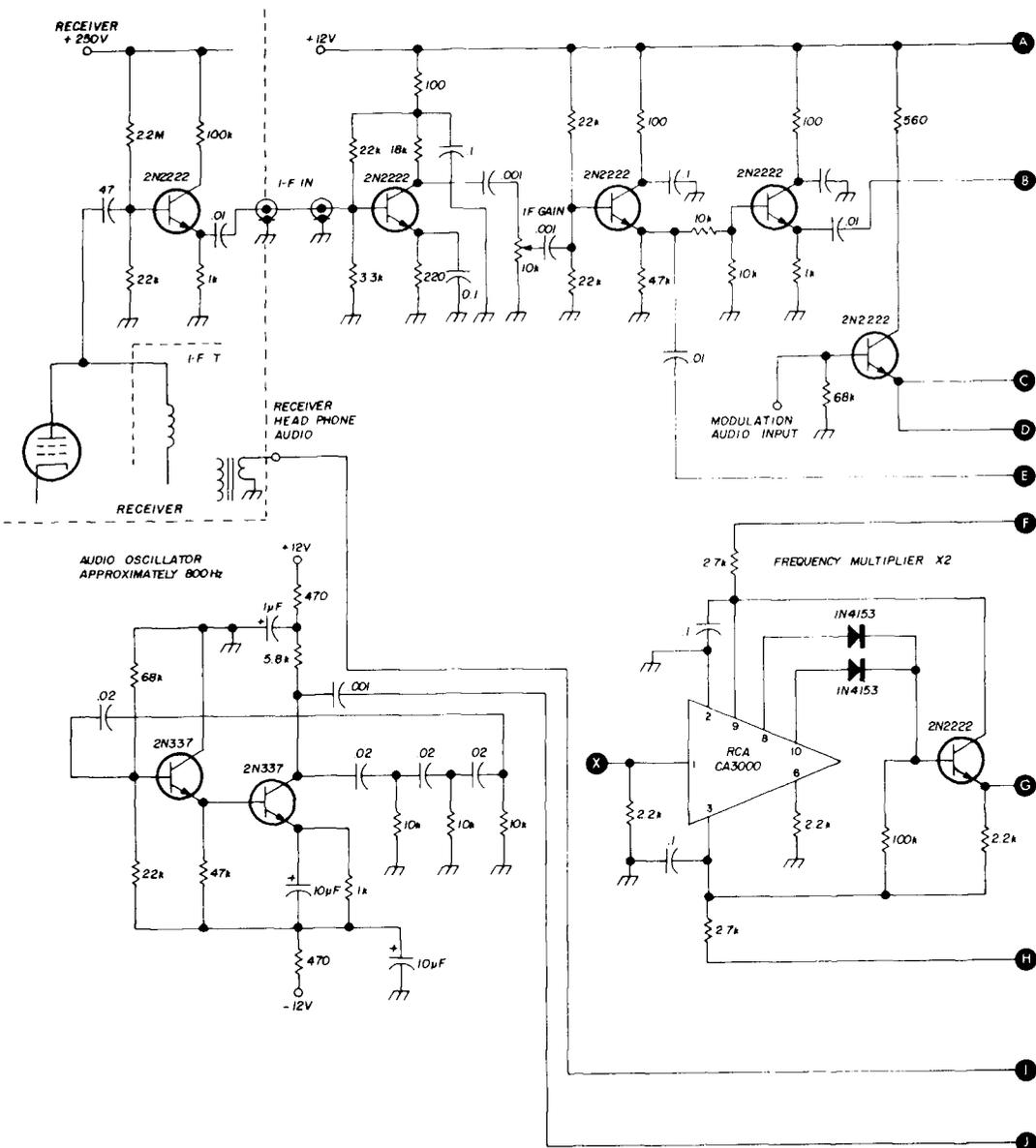


fig. 4. Schematic diagram of the cw signal processor.

results.

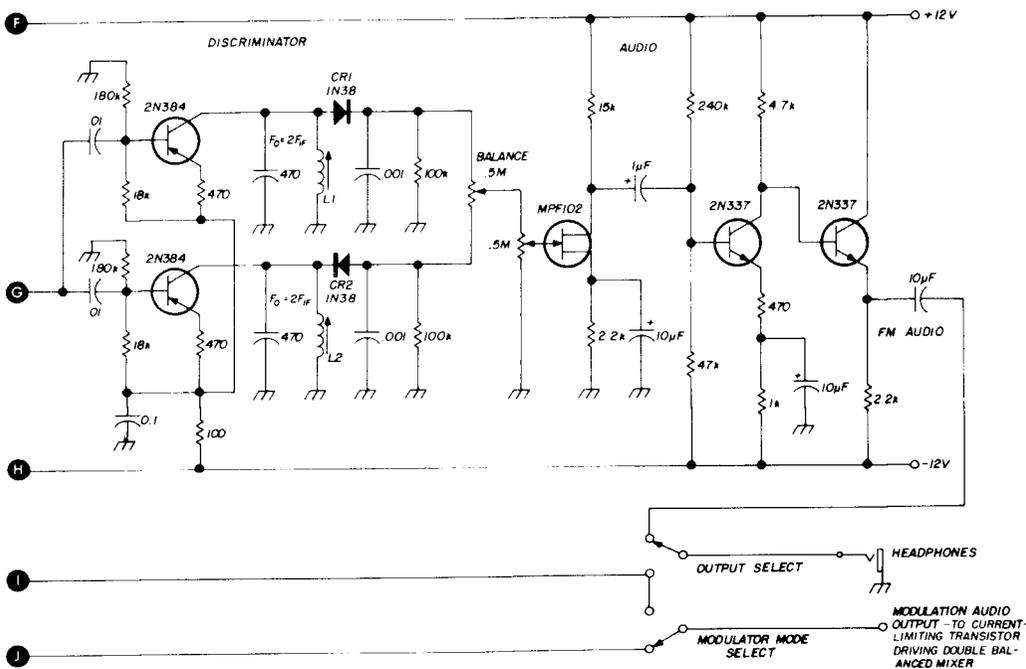
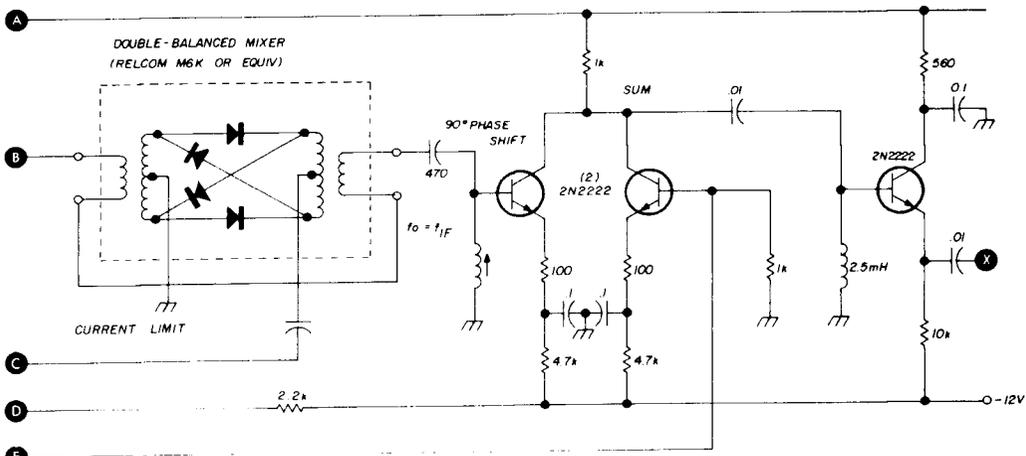
It is important to note that the coupling point should be right after the crystal filter (which is usually right after the mixer). This prevents the bfo oscillator signal from getting into the input of the frequency modulator.

summary

Aside from the improved results ob-

tained with this system in the presence of impulse noise, it is quite a treat to my tired old ears to be able to tune across the band, hearing weak signals and at the same time, not having my ear drums drilled out when that kilowatt down the street comes on.

Although the system shown in fig. 4 will not outperform a conventional bfo/product detector when low background



noise is present there is hope with future extensions of the basic idea.

The next step, of course, is to replace the discriminator with a phase-locked loop. Once this is done, a narrow-band audio filter may be used with more advantage than is gained when filters are placed after a conventional a-m or fm detector.

This frequency-modulated cw reception has proven interesting to me, but it is relatively complex, and the system with

the phase-locked loop will be even more so. (I am aware that there are easier ways to phase modulate a signal but I want two separate channels - I have plans there too!)

This article has stopped with the basic idea. With enough interest there will be more to come. The phone operators have had interesting new modulation methods to play with; isn't it about time the brass pounders had a new toy?

ham radio

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the application of stress analysis

to antenna systems

Understanding the
mechanical design
of antennas
and
supporting structures

Most amateurs at one time or another have wished to extend a mast a few feet higher or make a beam with longer elements. The question is, "How high or how long can one safely go?" The calculations involved in arriving at the answer to this question are relatively simple, and a good feeling for the problem may be had with little "stress or strain" on the part of the interested ham.

basic data

The problem simply involves comparison and matching of material strength

with maximum anticipated load. Two equations are used to develop the comparison:

$$M_L = F \times D \text{ and } M_R = f \times S$$

where M_L = bending moment developed by loading force (lb)

M_R = restraining moment developed by loaded structure (lb)

F = force of the load (lb)

D = distance from the effective point of application of F to the fulcrum (in.)

f = strength of the material being used (lb/in.²)

I = moment of inertia of the cross section about its neutral axis (in.⁴)

C = distance from the neutral axis to the extreme fiber (in.)

S = section modulus (in.³) = I/C

Some of these terms may not be too familiar and their calculation even less so. However, the table and chart makers of America have provided numerous graphic aids that do almost everything short of coming up with the final answer.

Eugene B. Fuller, W2FZJ, 1183 Wall Road, Webster, New York 14580

bending moment

Since the problem revolves about the "bending moment" concept, let's consider first what bending moment means and how it's computed. A bending moment is the product of a force and its distance to the point under consideration, where the force is the component perpendicular to a line drawn from the point of application to the point under consideration.

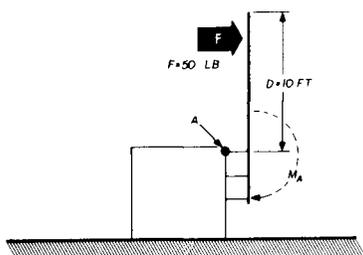


fig. 1. The bending-moment concept. Force F is a wind load exerted on a mast extending 10 feet above its last support.

example 1: In fig. 1 the bending moment at point A is:

$$\begin{aligned} M_A &= F \times D \\ &= 50 \text{ lb} \times 10 \text{ ft.} \\ &= 500 \text{ ft-lb.} \end{aligned}$$

This might represent the bending moment caused by a small antenna, with a 50-lb wind load, mounted on a mast that extends 10 feet above its last support, or the tension in a wire antenna supported by the mast.

The question now arises, "Fine, but where did the wind-load value come from?" This again is relatively easy to calculate; or if you're lucky, the job may have been done already by the manufacturer of your antenna or tower. The calculation goes like this:

$$F_{WL} = A_e \times P_W$$

where F_{WL} = wind load force

A_e = effective area on which the wind is acting

P_W = wind pressure

table 1. Summary of parameters used in bending-moment analysis examples.

wind velocity (mph)	wind pressure pressure on flat surface (lb/ft ²)*
10	.42
20	1.7
30	3.8
40	6.7
50	10.5
60	15.1
70	20.6
80	26.8
90	34.0
100	42.0

type of cross section	area factors	multiplication factor
open face (latticed) — square	— on face	2.20
	— on corner	2.40
	— triangle — on face	2.00
	— parallel to face	1.50
closed face (solid) — square or rectangular		1.00
	— hexagonal or octagonal	0.80
	— round or elliptical	0.60

*In most instances 20 lb/ft² is considered a safe value

Further explanation of A_e and P_W are as follows. The effective area is that of a flat plate which, if substituted for the object of interest, would accumulate the same total force. It is recommended that anyone working on this problem consult a handbook such as Reference 1 to obtain multiplication factors for different cross sections and wind-load values for different wind velocities. These values are summarized in table 1, and an example of their use follows.

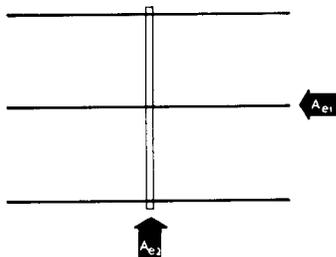


fig. 2. Typical beam antenna used in the wind-loading example. Boom is 2 in. dia. x 26 ft; elements are 1 in. dia. x 36 ft.

example 2 Find the wind load on a 3-element 20-meter beam (see **fig. 2**).

$$A_{e2} = 2 \text{ in.} \times \frac{1 \text{ ft.}}{12 \text{ in.}} \times 26 \text{ ft} \times .6 = 2.6 \text{ ft}^2$$

$$A_{e1} = 3 \times 1 \text{ in.} \times \frac{1 \text{ ft.}}{12 \text{ in.}} \times 36 \text{ ft} \times .6 = 5.4 \text{ ft}^2$$

$$A_{e_{\max}} = \sqrt{A_{e1}^2 + A_{e2}^2} = \sqrt{36} = 6 \text{ ft}^2$$

Using 20 lb/ft² wind load,

$$F_{WL} = 6 \text{ ft}^2 \times 20 \text{ lb/ft}^2 = 120 \text{ lb}$$

restraining moment

Now we'll examine the restraining moment, which is the force with which the support structure resists being bent by the loading forces. The equation for this moment, as given above, is:

$$M_R = \frac{fxl}{C} = fxS$$

The determination of the values to use in this equation has again been made easier by charts and tables available in many structural and mechanical handbooks. Some of the most common values are shown in **tables 2** and **3** and **fig. 3**.

table 2. Strength of materials.

material	tensile strength (lb/in. ²)	compressive strength elastic limit (lb/in. ²)
wrought iron	30,000	30,000
structural steel	40,000	40,000
copper (hard)	42,000	————
brass (hard)	60,000	————
aluminum-extruded tubing less than 1/4" wall;		
2024-T4	40,000	36,000
6061-T6	35,000	35,000
6063-T6	25,000	25,000
7178-T6	76,000	76,000
white oak	————	5,000
yellow pine	————	4,000

example 3 Now we can take a look at the problem of extending our mast above the tower to get the psychological advantage of a few extra feet of height. Assume the three-element 20-meter beam of **example**

table 3. Elements of section.

cross section		lb/ft	I	S	C
1/2 diameter	.031 wall	0.054	0.001	0.005	0.25
1/2	.062	0.101	0.002	0.008	0.25
1/2	.125	0.173	0.003	0.012	0.25
3/4	.062	0.159	0.008	0.021	0.375
3/4	.125	0.288	0.012	0.033	0.375
1	.062	0.216	0.020	0.041	0.500
1	.125	0.405	0.034	0.067	0.500
1 1/4	.062	0.274	0.041	0.071	0.625
1 1/4	.125	0.520	0.071	0.113	0.625
1 1/2	.062	0.332	0.073	0.097	0.750
1 1/2	.125	0.635	0.129	0.172	0.750
2	.062	0.447	0.179	0.179	1.000
2	.125	0.865	0.325	0.325	1.000
2	.250	1.616	0.537	0.537	1.000
3	.125	1.328	1.169	0.779	1.500
3	.250	2.540	2.059	1.373	1.500
4	.125	1.790	2.859	1.429	2.000
4	.250	3.463	5.200	2.600	2.000

I = moment of inertia (in.⁴)

S = section modulus (in.³)

C = distance from the neutral axis to the extreme fiber (in.)

Reference: *Alcoa Structural Handbook*, Aluminum Company of America, Pittsburg, Pennsylvania.

2 and assume a heavy-duty mast of 2-in. O. D. X. 1/4-in. wall steel tubing. From fig. 4,

- F_a = antenna wind load
- F_m = mast wind load
- D = height of mast
- d = mast diameter

The effect of mast wind loading is evenly distributed along the mast; therefore this force may be considered to act at a point half-way up the mast.

Maximum allowable bending moment:

$$M = f \times S$$

$$f = 40,000 \frac{\text{lb}}{\text{in.}^2} \text{ (from table 2)}$$

$$S = 0.537 \text{ in.}^3 \text{ (from table 3)}$$

$$= 40,000 \frac{\text{lb}}{\text{in.}^2} \times 0.537 \text{ in.}^3$$

$$= 21,500 \text{ in.} \cdot \text{lb} \times \frac{\text{ft}}{12 \text{ in.}} = 1,800 \text{ ft}\cdot\text{lb.}$$

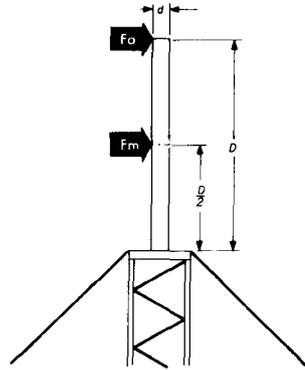
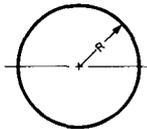


fig. 4. Extending a mast above the tower. Parameters are defined in the text.

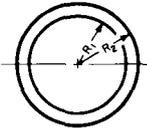


$$I = \frac{\pi \times R^4}{4}$$

$$C = R$$

$$S = \frac{\pi \times R^3}{4}$$

SOLID CIRCLE

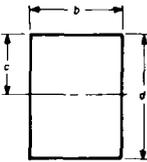


$$I = \frac{\pi (R_2^4 - R_1^4)}{4}$$

$$C = R_2$$

$$S = \frac{\pi (R_2^3 - R_1^3)}{4R_2}$$

HOLLOW CIRCLE

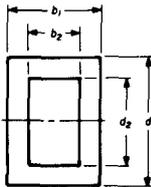


$$I = \frac{b \times d^3}{12}$$

$$C = \frac{d}{2}$$

$$S = \frac{b \times d^2}{6}$$

SOLID RECTANGLE



$$I = \frac{b_1 d_1^3 \times b_2 d_2^3}{12}$$

$$C = \frac{d_1}{2}$$

$$S = \frac{b_1 d_1^2 \times b_2 d_2^2}{6d_1}$$

HOLLOW RECTANGLE

fig. 3. Data for determining I, S, and C of different shapes.

Antenna wind load:

$$F_a = 6 \text{ ft}^2 \times 20 \frac{\text{lb}}{\text{ft}^2}$$

$$= 120 \text{ lb}$$

Mast wind load:

$$F_m = D \times d \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2}$$

$$= D \times .167 \times .6 \times 20$$

$$= 2D$$

Total bending moment:

$$M_{\text{total}} = F_a \times D + F_m \times \frac{D}{2}$$

$$= 120 \text{ lb} \times D + 2D \times \frac{D}{2}$$

$$= 120D + D^2 \text{ (ft}\cdot\text{lb.)}$$

Maximum allowable mast height:

Setting the maximum allowable bending moment equal to the total bending moment, we can now solve for the maximum allowable mast height.

$$1,800 = 120D + D^2$$

$$D^2 + 120D - 1,800 = 0$$

Using the general solution for a quadratic equation,

$$\frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

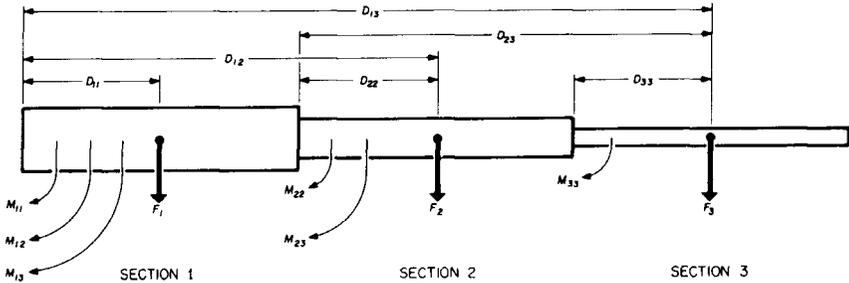


fig. 5. Parameters for a typical beam element designed to withstand 70-mph wind loads.

where $a = 1$
 $b = 120$
 $c = -1,800$

we have

$$D = \frac{-120 \pm \sqrt{120^2 - 4 \times 1 \times (-1,800)}}{2}$$

$$= \frac{-120 \pm 147}{2} = \frac{27}{2}$$

$$= 13.5 \text{ ft maximum}$$

A safety factor to suit the taste of the designer may now be applied to come up with the final height. If it is desired to allow for contingencies such as ice loading or inferior materials, the more conservative approach would be to use values in the initial calculations to account for them and not try to lump them all into a magic safety factor.

Use the safety factor to account for contingencies you have not thought of yet! If such an idea should appeal to you, be sure to consider also the allowable bending moment for your tower (should be available from manufacturer) and the added load on your rotator (15 ft of 2-in. diameter 1/4-in. wall steel tubing will weigh about 70 lb).

example 4 Next let's look at the design of a beam element, say a 20-meter element

supported at the center and designed to withstand 70-mph wind loading. Since the element is supported at the middle, we need consider only a half element (fig. 5).

Section 3: 1/2-in. diameter, 0.031-in. wall

$$M = S \times f$$

$$= .005 \text{ in.}^3 \times 35,000 \frac{\text{lb.}}{\text{in.}^2}$$

$$= 175 \text{ in.-lb} \times \frac{\text{Ft}}{12 \text{ in.}}$$

$$= 14.6 \text{ ft-lb allowable}$$

wind loading per linear foot

$$= .5 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2}$$

$$= .5 \frac{\text{lb}}{\text{ft}}$$

weight per linear foot

$$= .054 \frac{\text{lb}}{\text{ft}}$$

total load per linear foot

$$= .554 \frac{\text{lb}}{\text{ft}}$$

If section 3 is 7 feet long:

$$F_3 = 7 \text{ ft} \times .554 \frac{\text{lb}}{\text{ft}}$$

$$= 3.87 \text{ lb}$$

$$M_{33} = D_{33} \times F_3$$

$$= 3.5 \text{ ft} \times 3.87 \text{ lb}$$

$$= 13.5 \text{ ft-lb loaded}$$

This is a safe loading when compared to the calculated allowable load of 14.6 ft-lb.

Section 2: ¾-in. diameter, 0.062-in. wall

$$\begin{aligned}
 M &= S \times f \\
 &= .021 \text{ in.}^3 \times 35,000 \frac{\text{lb}}{\text{in.}^2} \\
 &= 735 \text{ in.}\cdot\text{lb} \times \frac{\text{ft}}{12 \text{ in.}}
 \end{aligned}$$

Total moment at 2-3 = 61.2 ft-lb allowable

$$\begin{aligned}
 \text{wind loading per linear foot} \\
 &= .75 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2} \\
 &= .75 \frac{\text{lb}}{\text{ft}}
 \end{aligned}$$

$$\begin{aligned}
 \text{weight per linear foot} \\
 &= .159 \frac{\text{lb}}{\text{ft}}
 \end{aligned}$$

$$\begin{aligned}
 \text{total load per linear foot} \\
 &= .909 \frac{\text{lb}}{\text{ft}}
 \end{aligned}$$

If section 2 is 6 feet long:

$$\begin{aligned}
 F_2 &= 6 \text{ ft} \times .909 \frac{\text{lb}}{\text{ft}} \\
 &= 5.45 \text{ lb}
 \end{aligned}$$

$$\begin{aligned}
 M_{22} &= F_2 \times D_{22} \\
 &= 5.45 \text{ lb} \times 3.0 \text{ ft} = 16.4 \text{ ft}\cdot\text{lb}
 \end{aligned}$$

$$\begin{aligned}
 M_{23} &= F_3 \times D_{23} \\
 &= 3.87 \text{ lb} \times (6 + 3.5) \text{ ft} = 36.8 \text{ ft}\cdot\text{lb}
 \end{aligned}$$

Total moment at 1-2 = 53.2 ft-lb

This is a safe loading when compared to the calculated allowable load of 61.2 ft-lb.

Section 1: 1-in. diameter, 0.062-in. wall

$$\begin{aligned}
 M &= S \times f \\
 &= .041 \text{ in.}^3 \times 35,000 \frac{\text{lb}}{\text{in.}^2} \\
 &= 1440 \text{ in.}\cdot\text{lb} \times \frac{\text{ft}}{12 \text{ in.}} \\
 &= 120 \text{ ft}\cdot\text{lb} \text{ allowable}
 \end{aligned}$$

$$\begin{aligned}
 \text{wind loading per linear foot} \\
 &= 1 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2} \\
 &= 1.0 \frac{\text{lb}}{\text{ft}}
 \end{aligned}$$

$$\text{weight per linear foot} = .216 \frac{\text{lb}}{\text{ft}}$$

$$\text{total load per linear foot} = 1.216 \frac{\text{lb}}{\text{ft}}$$

If section 1 is 5 feet long:

$$\begin{aligned}
 F_1 &= 1.216 \frac{\text{lb}}{\text{ft}} \times 5 \text{ ft} \\
 &= 6.08 \text{ lb}
 \end{aligned}$$

$$\begin{aligned}
 M_{11} &= F_1 \times D_{11} \\
 &= 6.08 \text{ lb} \times 2.5 \text{ ft} = 15.2 \text{ ft}\cdot\text{lb}
 \end{aligned}$$

$$\begin{aligned}
 M_{12} &= F_2 \times D_{12} \\
 &= 5.45 \text{ lb} \times (5 + 3.0) \text{ ft} = 43.6 \text{ ft}\cdot\text{lb}
 \end{aligned}$$

$$\begin{aligned}
 M_{13} &= F_3 \times D_{13} \\
 &= 3.87 \text{ lb} \times (5 + 6 + 3.5) \text{ ft} = 56.0 \text{ ft}\cdot\text{lb}
 \end{aligned}$$

Total moment at center of element = 114.8 ft-lb

This is a safe loading when compared to the calculated allowable load of 120 ft-lb.

From these calculations the general method may be seen for solving this type of problem. It should be noted that the size of tubing used for each section was arrived at by trial and error; i. e., picking a size, calculating the allowable moment and loaded moment for a reasonable length, until a good and reasonable combination was obtained.

If it's desired to account for possible ice loading, simply increase the element diameter for the wind loading calculation and compute the weight of the ice and add it to the weight of the tubing

$$(\text{Weight of ice} = 62.5 \frac{\text{lb}}{\text{ft}^3} \times .9 = 56.2 \frac{\text{lb}}{\text{ft}^3}).$$

If you wish to design a self-supporting vertical antenna, the same technique is also used. In this case, however, the weight causes compressive stress rather than bending moments and in most cases may be neglected.

example 5 As another example, therefore, let us look at the design of a 20-meter self-supporting vertical, 18 feet high, allowing for 70 mph wind and 1/4 in. of radial ice. **Fig. 6** illustrates the parameters.

Section 3: ½-in. diameter, 0.125-in. wall, (plus ¼ in. of ice)

$$\begin{aligned}
 M &= S \times f \\
 &= .012 \text{ in.}^3 \times 35,000 \frac{\text{lb}}{\text{in.}^2} \\
 &= 420 \text{ in.-lb} \times \frac{\text{ft}}{12 \text{ in.}} \\
 &= 35 \text{ ft-lb allowable}
 \end{aligned}$$

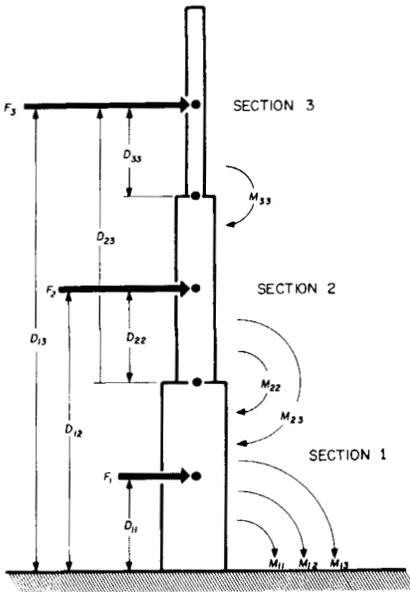


fig. 6. Parameters for a self-supporting vertical antenna allowing for 70-mph wind and ¼ in. of ice buildup.

wind loading per linear foot

$$\begin{aligned}
 &= 1 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2} \\
 &= 1.0 \frac{\text{lb}}{\text{ft}}
 \end{aligned}$$

If section 3 is 8 feet long:

$$\begin{aligned}
 F_3 &= 1 \frac{\text{lb}}{\text{ft}} \times 8 \text{ ft} \\
 &= 8 \text{ lb}
 \end{aligned}$$

$$\begin{aligned}
 M_{33} &= D_{33} \times F_3 \\
 &= 4 \text{ ft} \times 8 \text{ lb} \\
 &= 32 \text{ ft-lb}
 \end{aligned}$$

This is safe loading when compared to the

calculated allowable load of 35 ft-lb.

Given the procedure for finding the wind load with 1/4 inch of ice on section 3 of the example shown in fig. 6, I'll leave as an exercise the calculation of moments for the other two elements. Assume the following: section 2: 3/4-in. diameter, 0.125-in. wall, 5 ft long section 1: 1-in. diameter, 0.125-in. wall, 5 feet long

If your calculations are correct, you should obtain a value of 177.5 ft-lb at the base, which is a safe load compared to the calculated allowable of 195 ft-lb.

compressive stress

Assume 1-in. diameter elements with 1/4-in. ice load and 18 feet high:

$$C = \frac{W}{A}$$

where C = compressive stress

W = total weight

A = cross section area of restraining material

weight of tubing

$$\begin{aligned}
 &= 0.405 \frac{\text{lb}}{\text{ft}} \times 18 \text{ ft} \\
 &= 7.3 \text{ lb}
 \end{aligned}$$

weight of ice

$$\begin{aligned}
 &= \pi(r_2^2 - r_1^2) \text{ in.}^2 \times \frac{\text{ft}^2}{144 \text{ in.}^2} \\
 &\quad \times 18 \text{ ft.} \times 56.2 \frac{\text{lb}}{\text{ft}^3} \\
 &= \frac{3.14(.75^2 - .50^2) \times 18 \times 56.2}{144}
 \end{aligned}$$

$$= 6.9 \text{ lb}$$

total weight = 7.3 lb + 6.9 lb

$$= 14.2 \text{ lb}$$

$$\begin{aligned}
 A &= \pi(r_2^2 - r_1^2) \text{ in.}^2 \\
 &= 3.14(.500^2 - .375^2) \text{ in.}^2 \\
 &= .344 \text{ in.}^2
 \end{aligned}$$

$$\begin{aligned}
 C &= \frac{W}{A} \\
 &= \frac{14.2 \text{ lb}}{.344 \text{ in.}^2} \\
 &= 41.3 \frac{\text{lb}}{\text{in.}^2}
 \end{aligned}$$

This is negligible when compared to the strength of aluminum at $35,000 \frac{\text{lb.}}{\text{in.}^2}$

From this example we observe two important characteristics. First, the additional wind area caused by ice loading results in the requirement for heavier

guying a 65-foot tower that's supporting the antenna in **example 2**. See **fig. 7**.

example 6 Force of wind on tower:

$$\begin{aligned} \text{area of one face} &= 2(1 \text{ in.} \times 12 \text{ in.}) = 24 \text{ in.}^2 \\ \frac{1}{4} \text{ in.} (12 \text{ in.} + 16 \text{ in.}) &= 7 \text{ in.}^2 \\ \text{total} &= 31 \text{ in.}^2 \end{aligned}$$

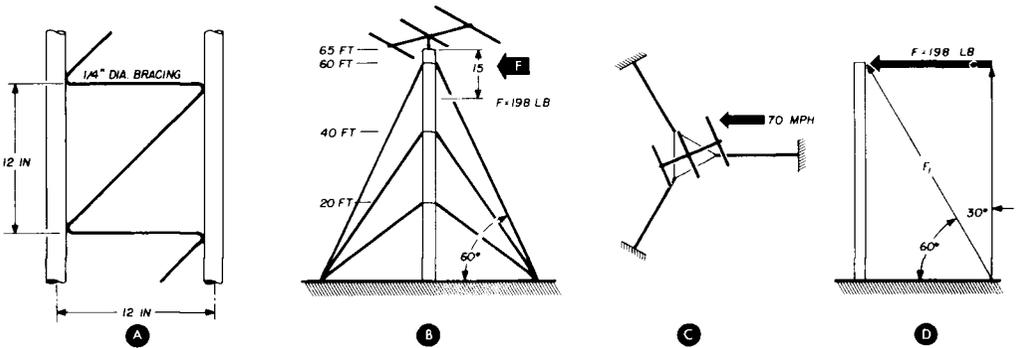


fig. 7. The guying problem for a 65-ft. tower supporting a typical 3-element beam. In A, tower elements are 1 in. dia. with 1/4 in. bracing. B and C illustrate text description of wind load on top guys. D is geometry for calculating force on the top guy.

construction than the beam element of **example 4** despite its being supported vertically instead of horizontally. Second, the compressive stress in the vertical element, even with ice loading, is quite negligible.

If you would like to design a husky beam element, you might try designing for 70 mph wind with simultaneous 1/4 in. of ice!

guying

Another problem for many hams is determining the proper size cables to use for guying an antenna support structure. A simplified approach will be taken to this problem, which should allow adequate safety.

To deal with this problem it's necessary to understand the concept of resolving a force into components. The reader is urged to refer to almost any high school physics book to review this technique.

Let's now consider the problem of

effective area of one face (cylindrical surfaces)

$$\begin{aligned} &= .6 \times 31 \text{ in.}^2 \\ &= 18.6 \text{ in.}^2 \end{aligned}$$

effective wind area of open latticed triangular tower with wind applied perpendicular to a face = $18.6 \text{ in.}^2 \times 2$

$$= 37.2 \text{ in.}^2 \frac{\text{ft.}}{144 \text{ ft.}^2}$$

effective total wind area per linear foot of tower = $.258 \frac{\text{ft.}^2}{\text{ft}}$

Tower wind load per linear foot

$$\begin{aligned} &= .258 \frac{\text{ft.}^2}{\text{ft}} \times 20 \frac{\text{lb}}{\text{ft.}^2} \\ &= 5.16 \frac{\text{lb}}{\text{ft}} \end{aligned}$$

The top set of guys will have to take all of the wind force above that set of guys plus the antenna, and about half the force between the top and middle set of guys (**fig. 7B**).

wind load of antenna (from example 2)
= 120 lb

wind load of tower = $5.16 \frac{\text{lb}}{\text{ft}} \times 15 \text{ ft}$
= 78 lb

total wind force = 198 lb

Component of force in the top front guy
(see fig. 7C):

$$F_1 = \frac{198 \text{ lb}}{\sin 30^\circ} = \frac{198 \text{ lb}}{.50} = 396 \text{ lb}$$

Added to this would be about 100 lb of force from tightening up the guy, or a total of just about 500 lb of tension. Therefore, using 1/8 in. aircraft cable with a breaking strength of 2100 lb would offer a comfortable safety margin. The strain on the lower guys would be somewhat less: the calculation is left as an exercise for the reader.

When selecting guying cable, careful attention should be paid to the load rating. There are many different types of 1/8 in. cable — some flexible stainless steel, some flexible winching cables with fiber core, less-flexible 5- or 7-strand galvanized, and of course the single strand type — all with different load ratings. Also, a given cable will have many ratings depending on the service; i. e., breaking strength, yield strength, and working strength. And for that matter, the working strength will vary depending on the intended use; e. g., winch service rating would be lower than that for guying service. Suffice to say — check with your supplier and be sure of what you're buying.

These are the basics. If you've been guessing up to now, you might try some calculations to see how safe you are. If you've been putting off a project because of fear of disaster, perhaps you can try it now (with a *little* less fear)!

reference

1. *Reference Data for Radio Engineers*, International Telephone and Telegraph Corp., New York, New York.

ham radio

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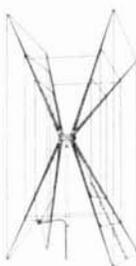
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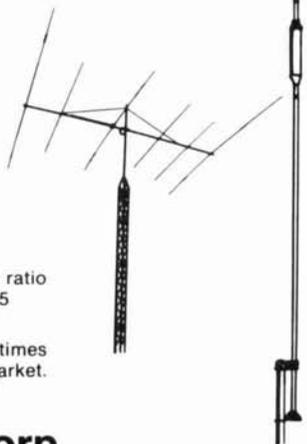
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solid-state RTTY monitor scope

This solid-state
RTTY monitor scope
was designed
specifically as a
companion to the
ST-6 terminal unit

Anthony Sperduti, WB2MPZ, 4740 Newton Road, Hamburg, New York 14075

In the past amateurs who turned to RTTY usually used a vacuum-tube type terminal unit such as the TTL/2 or the circuit designed by W2PAT. Now the trend is toward the high performance solid-state ST-5 or ST-6 designed by W6FFC.^{1,2} When I decided to go solid state I decided to go all the way and include a monitor scope that was based on solid-state components.

Although the oscilloscope circuit shown in fig. 1 is designed primarily for use with the ST-5 or ST-6 it can be used with any RTTY terminal unit as long as the vertical and horizontal amplifiers are not over-driven.

circuit

The amplifier circuits are quite simple. In the vertical amplifier an emitter-follower input stage (Q1) provides high input impedance. The CRT driver stage (Q3) is rated at 250 volts. Vertical gain is controlled by the 5000-ohm pot (R9) in the emitter circuit of the driver stage.

The horizontal amplifier circuit is identical to the vertical lineup. Note in fig. 1 that even-numbered components are part of the horizontal amplifier chain; components in the vertical amplifier are designated by odd numbers. The gain of

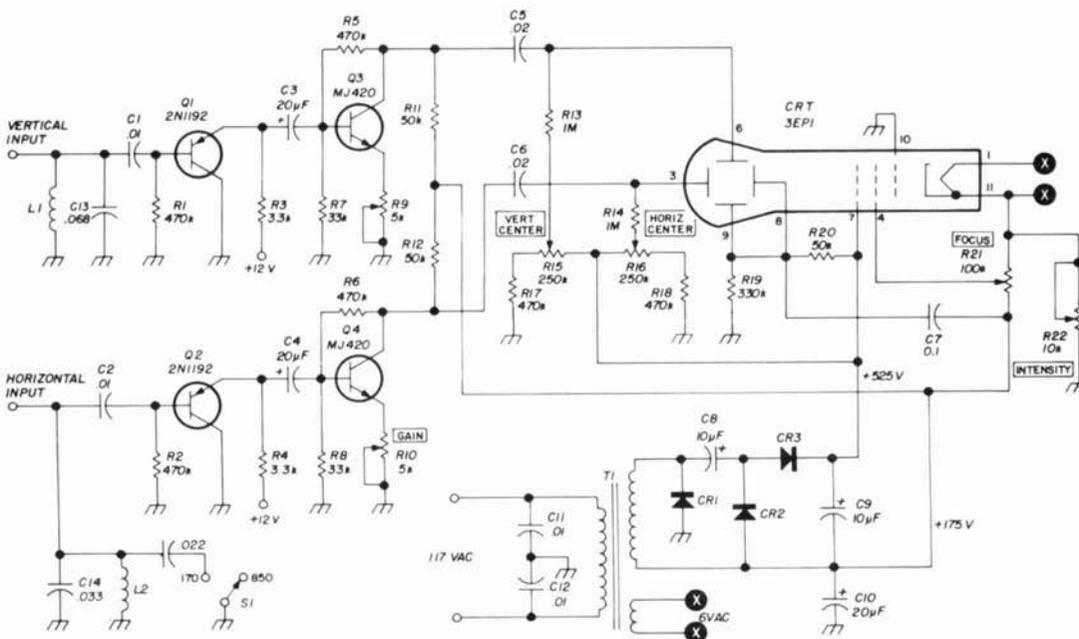


fig. 1. Solid-state RTTY monitor scope is designed for the ST-6 RTTY terminal unit.

the amplifier system is more than adequate to fill the screen of a 3-inch CRT.

The filters used in the Mainline ST-5 and ST-6 terminal units result in a very poor oscilloscope display since there is voltage on the *mark* toroid when a *space* signal is present, and vice versa. When the ST-6 is used on 850 shift the scope display looks like two bananas; on 170 shift the display looks like two fat footballs.

For example, when tuned to *mark* there is still vertical deflection voltage which opens up the horizontal trace. The higher the vertical voltage, of course, the more oval the display. This problem is inherent in simple single-tuned channel filters and not at all bad as far as the terminal unit is concerned. To clean up the monitor scope traces W6FFC recommended a high-Q tuned circuit at the input to the scope; this is shown in fig. 1.

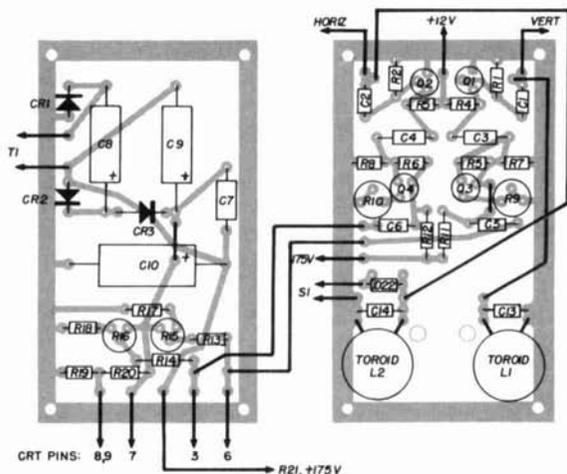


fig. 2. Printed-circuit board layout for the RTTY monitor scope.

Toroid L1 is tuned to 2125 Hz; L2 is tuned to 2975 Hz with S1 open. Close S1 and pad the 0.022- μ F capacitor to tune L2 to 2295 Hz. These simple circuits are tuned so that maximum deflection of the scope trace coincides with maximum deflection of the ST-6 tuning meter.

construction

The RTTY monitor oscilloscope circuitry is built on two 3 x 6-inch printed-circuit boards. Circuit-board layout is shown in fig. 2. If you look at the boards carefully you will see that I have drilled two holes through the amplifier board so the centering controls (R14 and R15) on the power-supply board can be easily adjusted.

If you want, these controls could be located on the front panel, although they seldom require attention after the circuit has been initially set up. I put the focus control (R21) and intensity control (R22) on the front panel of my unit.

The power transformer I used for the 12-volt supply is the same as that used in the ST-5 and ST-6. However, use a separate transformer for the monitor scope; don't try to make the terminal-unit transformer do double duty.

For potentiometers R15 and R16 I used Ohmite type RV6NAV that I obtained surplus. You can use the Mallory MTC254L1 instead, but be careful because the whole potentiometer frame will be *hot* to the tune of 500 volts.

summary

This monitor scope is a nice companion to the Mainline ST-6 RTTY terminal unit. At my station I included it in the same cabinet as the ST-6. Performance is excellent, and the scope traces are clean and easy to read.

references

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2. Irvin M. Hoff, "The Mainline ST-6 RTTY Demodulator," *ham radio*, January, 1971, page 6.

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resistor performance at high frequencies

A comparison of the high-frequency response of solid carbon-composition and metal-film resistors

Most amateurs are aware of what happens to a capacitor as the operating frequency is increased — it becomes series resonant at the frequency where its internal inductance resonates with its capacitance. However, many amateurs may not realize that carbon resistors, as well as metal-film types, show resistive and reactive changes as the frequency is varied from dc to vhf.

Since resistors are the most common components used in electronic equipment it is helpful to know as much as possible about this *so-called* passive element. You may find that those resistors are not nearly as "passive" as you may think — under the right circumstances they may even exhibit a rather capricious nature.

In this article I will discuss the high-frequency characteristics of resistors. I will not cover wirewound types because of their high inherent inductance, even "non-inductive" types.

Before looking at high-frequency resistor performance we should briefly re-

view the construction of both *composition* and *metal-film* types and the general characteristics of each. The *composition* resistor consists of a mixture of resistive material and a dielectric binder that is molded into a cylindrical shape. *Metal-film* resistors are composed of a resistive film deposited on, or inside, an insulating ceramic cylinder.

These two resistor types differ from each other in size, resistance range, cost, power dissipation and general characteristics. One type may be better than the other for particular purposes but neither type has all the best characteristics. Therefore, resistor choice depends upon the circuit requirements, the environment in which it must operate and many other factors.

composition resistors

Generally speaking, solid composition resistors have poorer stability since their resistance is a function of body temperature, circuit voltage, moisture content and previous history.

The high-frequency characteristics of composition resistors are good, but not quite as good as film types. Composition resistors generate a noise voltage when current flows through them; this is inherent in the construction of the resistor and is much greater than in any other type. Therefore, composition types are not recommended for circuits handling low-level signals.

The reliability of composition resistors is good since they seldom open up unless badly overloaded; however, they do change value by several percent with changing operating conditions. For example, they should not be used in voltage dividers where accuracy and voltage

M. A. Ellis, K1ORV, 61 Marlboro Road, Sudbury, Massachusetts 01776

stability is required. Also, they shouldn't be used where a long-term permanent resistance change of ± 10 percent or more cannot be tolerated.

In addition, composition resistors should be derated 50 percent* to increase life and stability in moderate temperatures, and even more than this for operation in high ambient temperatures.

film. However, exposure to moisture may seriously affect resistance if the element is not well protected by its casing.

The metal-film resistor should also be protected to reduce the possibility of physical damage. The temperature coefficient of resistance is extremely low (that is, the resistance change with change in operating temperature).

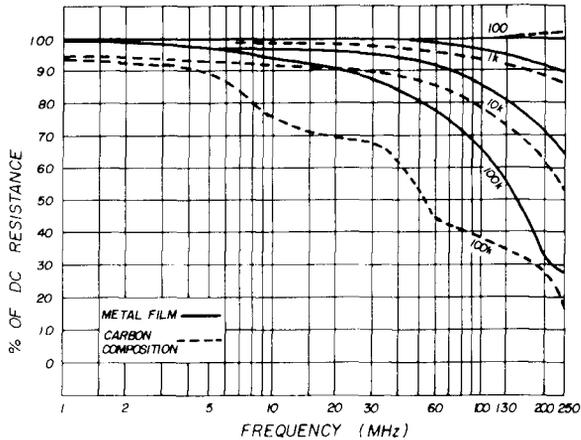


fig. 2. Resistance of both metal-film and carbon-composition resistors decreases with frequency. Larger values are affected more than smaller values.

metal-film resistors

Metal film type resistors generate considerably less noise voltage than composition types since the conducting paths are more homogeneous. Stability is high, environmental changes have little effect, and they have high reliability. Also, the combined effects of climate and operation will not change initial resistance by more than about 3 percent under normal operating extremes.

The maximum full-power operating temperature of metal-film resistors may be allowed to reach 70 to 100 degrees C. Skin effects are negligible since the entire resistance path is made up of a surface

high-frequency operation

A resistor that can be assumed to exhibit only resistance at low frequencies develops inductance and shunt capacitance as frequency is increased. See fig. 1 for a simplified equivalent circuit of an ordinary resistor at high frequencies. In a well constructed composition or film resistor the inductance is due solely to the connecting leads.

Shunt capacitance is due to the capacitance between end caps (or the lead terminations) and the shunt capacitances formed by the conducting particles which are held in contact by the dielectric binder. The solid composition resistor naturally has a greater number of such contacts than the film type resistor.

resistance/reactance measurements

I have made a series of measurements on a set of good quality carbon composi-



fig. 1. Simple equivalent circuit of a resistor includes series inductance and shunt capacitance.

*A 1/2-watt resistor should be used to dissipate 1/4-watt maximum.

tion resistors and an equal number of quality metal-film types,* having values of 100, 1000, 10k and 100k ohms.

A Boonton RX Meter, model 250A, was used for the measurements. This is a wide frequency range impedance meter that is designed to give an accurate reading of the equivalent parallel resistance and parallel reactance of two-terminal networks or components. Two

within the specified tolerances ($\pm 5\%$ for the composition types and $\pm 2\%$ for the metal-film resistors).

Graphs of resistance and capacitance versus frequency for the four resistor values are shown in figs. 2 and 3. It can be seen that the resistance of the higher resistance units decreases at higher frequencies, with the metal-film types appearing to be slightly better performers.

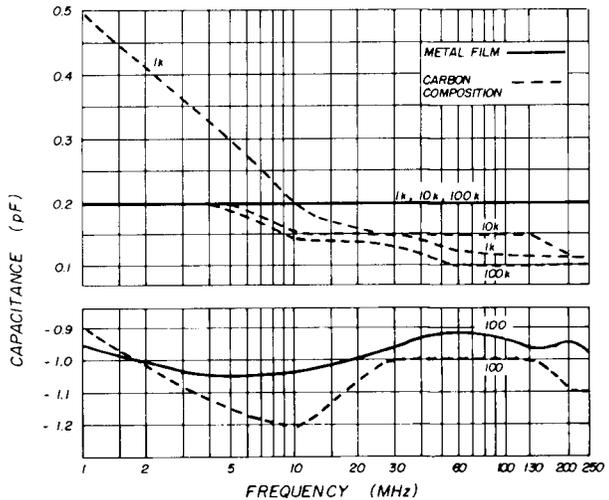


fig. 3. Shunt capacitance of carbon-composition resistors changes much more with increasing frequency than the shunt capacitance of metal-film types.

calibrated dials, labeled R and C, are used to balance a bridge.

The equivalent parallel resistance is then read from the R dial. The positive or negative resonating capacitance of the resistor (\pm pF) is indicated by the C dial.

high-frequency characteristics

All resistor leads were clipped to the same length (3/8 inch) before making any measurements. The first measurement was for dc resistance — all resistors were

This drop-off in resistance with frequency was first discovered by Boella and is called the "Boella effect." Howe¹ suggested a transmission-line theory to describe this behavior, and this is generally considered correct.

At high frequencies the resistor looks like a network of resistances and capacitances according to Howe, so the resistance reduction with increasing frequency is due mainly to the shunting effect of distributed capacitance in the resistor (see fig. 4).

For the best high-frequency performance the controlling conditions are geometry (size and shape of the resistor) and minimum dielectric losses. It has

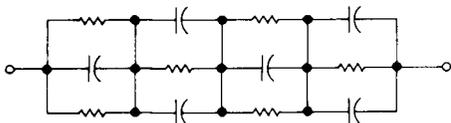


fig. 4. Realistic equivalent circuit of a resistor at high frequencies actually looks like an RC network.

*The composition resistors were Allen-Bradley type CB, 1/4 watt, 5% tolerance. The metal-film resistors were Corning Glass type C4, 1/4 watt, 2% tolerance.

been proven that the smaller the diameter of the resistor (minimum cross-sectional area), the better will be its high-frequency response, all other things being equal.

The small-diameter resistor will have fewer contacts to contribute capacitance than a larger unit. Generally the terminals and lead terminations will also be smaller and therefore contribute less capacitance. The dielectric losses are kept low by a

conclusions

If you consider other criteria, such as stability, environment, inherent noise, high reliability, susceptibility to physical damage or cost, then one type or the other should be used for that reason and the high-frequency response should not generally be a major consideration.

If, however, high-frequency response is important, you should use the smallest

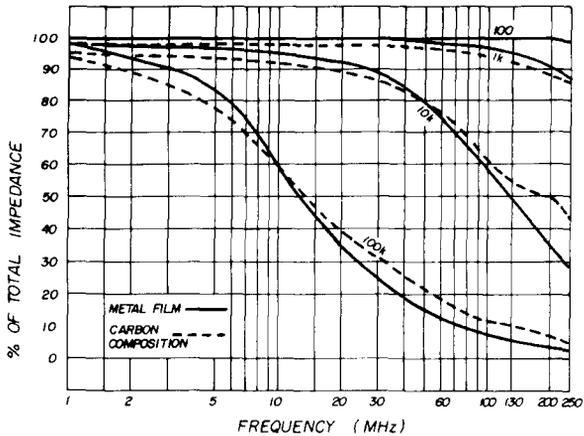


fig. 5. Total resistor impedance decreases with frequency. This is especially apparent with the larger values. Resistor impedance is a much more meaningful measurement than simple ac resistance.

good choice of base material. If binders are used their total mass should be kept to a minimum.

total impedance

The total impedance of a resistor with shunt capacitance appears to be a more meaningful indicator of resistor performance at high frequencies than the basic ac resistance measurement.* A comparison of the curves in fig. 5 indicates that very little difference exists between the two types of resistors at high frequencies if the basis of comparison is total impedance alone.

*The total impedance of the resistor is

$$Z = \frac{R_p X_p}{\sqrt{R_p^2 + X_p^2}}$$

where R_p is the equivalent parallel resistance and X_p is the equivalent parallel reactance (see fig. 2). X_p may be calculated for each test frequency using the standard reactance formula, $X = 1/2\pi fC$. The value of C is found in fig. 3.

physical size possible consistent with good design practice, e. g., if a 1/8-watt resistor will do the job safely, use it rather than a 1/4-watt or larger value.

The resistor leads and interconnecting wires should be as short as feasible and resistor placement should be chosen with care. Resistor capacitance to ground, for example, may greatly increase shunt capacitance and accentuate high-frequency roll-off.

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3. Keith Henney, Craig Walsh, "Electronic Components Handbook," Chapter III, Technical Writing Service; McGraw-Hill, New York, 1957.
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ham radio

modified inverted-V antenna

This modified
inverted-V antenna
provides complete
multiband operation
from 40 through 10

Many amateurs I have worked have expressed interest in my rather unconventional multiband antenna. I call it a modified inverted-V because it began as an inverted-V for 40 meters. However, it has been modified considerably and now provides good performance on all bands from 40 through 10 meters.

For those amateurs who are already using an inverted-V antenna, this design may give them some ideas for covering other bands. For the amateur with limited space this antenna offers good efficiency with minimum size.

The multiband inverted-V antenna shown in fig. 1 offers many advantages, including small physical size, relatively low height, broadband response with low swr and requires no traps or tuning devices. In addition, it appears to be nearly omnidirectional.

The vswr curves plotted in fig. 2 were measured with a Knight P2 swr bridge. As you can see, the antenna is cut for phone

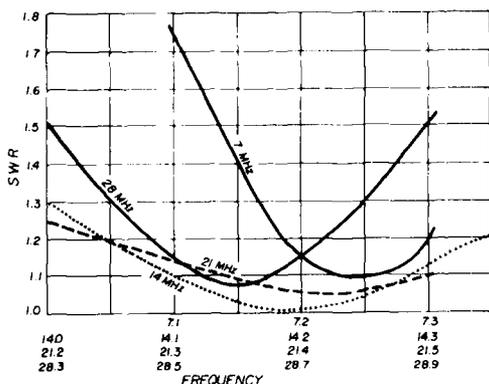


fig. 2. Swr performance of the W2KTW modified inverted-V.

Bob Austin, W2KTW, 120 Hamilton Avenue, Dumont, New Jersey 07628

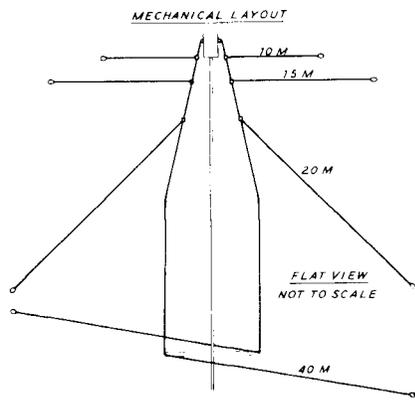
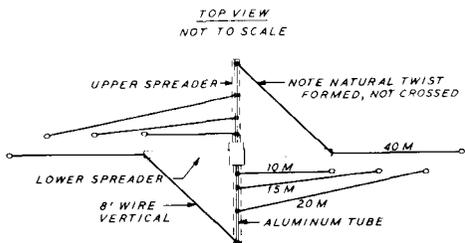
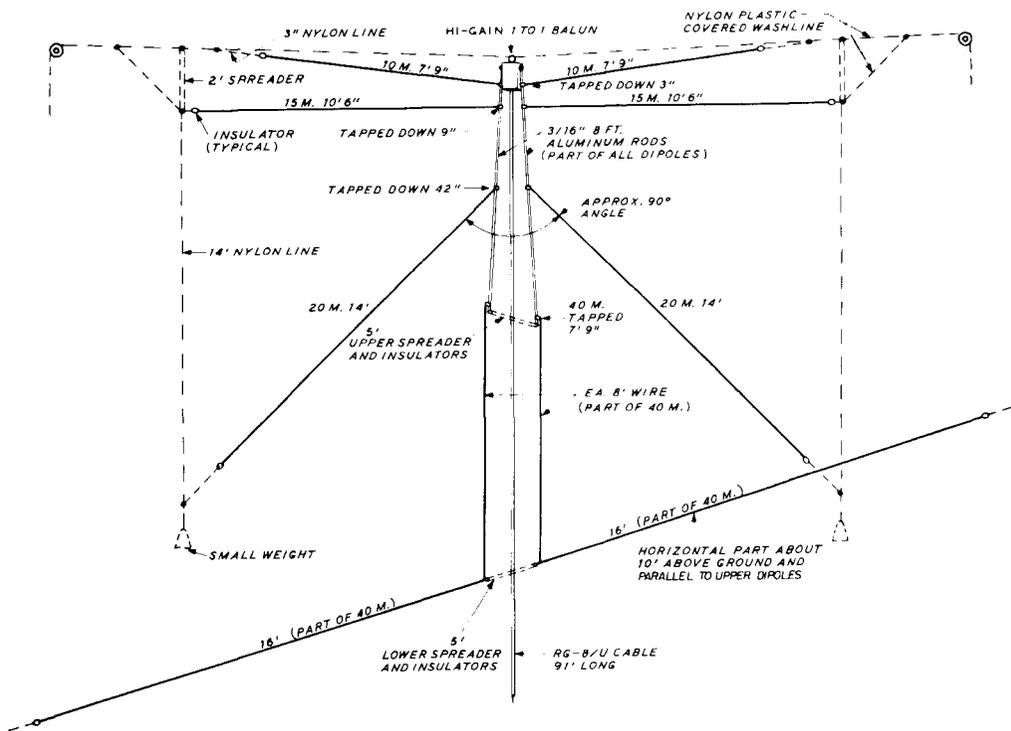


fig. 1. Modified inverted-V antenna provides good operation on 40, 20, 15 and 10 meters.

portions of the amateur bands. Although my modified inverted-V has the dimensions given in fig. 1, at other locations it will probably be necessary to trim each of the sections to resonate at the desired frequency.

On 20 meters, where the antenna appears as a more conventional inverted-V it was noted that decreasing the angle between the elements raised the resonant frequency.

On 40 meters it was noted that the horizontal part of the elements must be

180° away from the higher-frequency elements. Also of importance is the fact that the vertical support provides more of a twist than a transposition.

With this antenna I have obtained optimum loading on all bands with my TR-4 transceiver. In the future I hope to devise a way of including a 75-meter antenna within the limited space I have available.

ham radio

series-tuned pi networks

An output
tuning network
that increases
the upper
frequency limit
of many
power-amplifier
circuits

One of the problems confronting the rf power amplifier designer in the upper hf and the lower vhf range from 20 to 150 MHz has been the achievement of tank-circuit constants that provide normal operating Q and decent power transfer. For years the bandswitched pi-network has been the accepted tank circuit for most amateur multiband rf power amplifiers. Unfortunately, this has precluded some otherwise excellent inexpensive rf power amplifier tubes from use above 15 meters because the tube plate output capacitance, added to the minimum capacitance of the input tuning capacitor and other stray circuit capacitance, has made it impossible to achieve a reasonable operating tank circuit Q.

Frank N. Van Zant, W2EGH, 101 Somers Place North, Moorestown, New Jersey 08057

When high-C, low-L tanks are tried with these tubes the results are generally unsatisfactory due to high circulating currents in the tank inductor which result in heating and loss of efficiency.

Attempts have been made to overcome this deficiency by eliminating the pi-network input tuning capacitor; to tune the tank the inductance is adjusted by a moveable slug or shorted turn. Usually this results in complicated or unwieldy mechanical assemblies that are limited to single-band amplifiers, and power losses are inherent in the slug or short-circuit turn.^{1,2} For example, one of the problems that detracted from the use of parallel 813s in bandswitched pi-network amplifiers was that of obtaining reasonable operating Q on 10, 15 and occasionally, 20 meters. Reasonable Q was particularly unattainable on 10 meters due to the high plate output capacitance of two 813s in parallel (28 pF). Therefore, use of two 813s on ten meters was virtually restricted to push-pull, plug-in coil, link-coupled amplifiers or to single-band configurations. The single 2E26, or two 6146s, on two meters is normally restricted to compromise LC tank circuits which use link output coupling. Since the

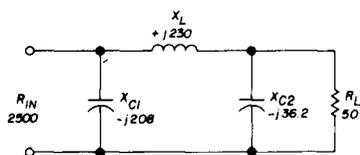


fig. 1. Basic pi network. Component values are discussed in text.

link has to be series tuned the transmitter ends up with as many tuning controls as a pi-network, but without its advantages.

tank circuit design

The operating Q of most plate tank circuits is chosen in the range of 10 to 20. In the pi-network the operating Q is a function of tube plate resistance and capacitance at the input to the network. Varying either of these parameters affects the component values of the pi-network.

A careful review of the reference material will reveal that to maintain a fixed value of operating Q as plate voltage and current (and therefore plate resistance) are changed, the values of C and L must be changed. For a given plate voltage/current ratio the Q will vary directly as the tank capacitance; ie, doubling tank capacitance doubles the Q.

For a given value of Q the input tuning capacitor in a pi-network can be larger with a low plate voltage/current ratio than with a high plate voltage/current ratio.

To achieve an operating Q of 12 in a pi-network with a plate voltage/current ratio of 5 (1500 volts at 300 mA or 2000 volts at 400 mA), the reactance of the input tuning capacitor is expressed by

$$X_c = \frac{R_p}{Q} \quad (1)$$

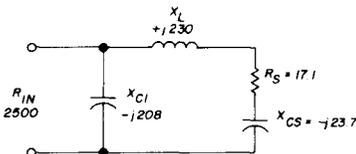
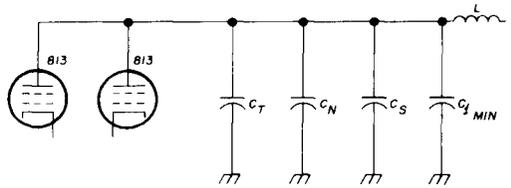


fig. 2. Series-tuned pi network that is electrically equivalent to the circuit in fig. 1.

where R_p is the tube plate resistance and is equal approximately to $E_p/2I_p$. For the case of the amplifier operating at 1500 volts and 300 mA with a Q of 12 the input tuning capacitor should exhibit 208 ohms capacitive reactance. This represents a 27.3-pF capacitor at 28 MHz. The other pi-network values may be determined from the formulas given in the

ARRL Handbook.³ For this case the reactance of the output capacitor is 36.2 ohms (157 pF at 28 MHz). The required inductance is 1.31 μ H at 28 MHz (230 ohms inductive reactance).



type of capacitance	bandswitched amplifier	single-band amplifier
C_{tube}	28 pF	28 pF
$C_{\text{neutralizing}}$	6 pF	6 pF
C_{stray}	20 pF	6 pF
C_1 minimum	<u>25 pF</u>	<u>15 pF</u>
Total C	79 pF	55 pF
X_c	72 ohms	103 ohms
Minimum realizable Q for 2500-ohm plate load	34.7	19.25

fig. 3. Estimates of input tuning capacitance for bandswitched and single-band rf power amplifiers.

The pi-network has the basic form shown in fig. 1 with the output load R_L in parallel with the loading capacitor X_{C2} . The reactance values are those given above. This basic circuit arrangement can be converted into the series-tuned configuration in fig. 2 by finding the equivalent series impedance of the two output components:

$$Z_s = \frac{R_L X_{C2}}{R_L + X_{C2}}$$

For the circuit constants in fig. 1, the equivalent series impedance is 17.1 - j23.7 ohms.

At the resonant frequency the network looks like a pure resistance at the

input terminals R_{in} . Although this matches the plate resistance of a typical pair of 813s it is impossible to use this circuit on 10 meters because the tube output capacitance and stray circuit capacitance are greater than the minimum value of C1. This is the limiting factor in tank circuit Q at the highest frequency of operation.

It becomes extremely important to arrive at a circuit configuration and operating voltage/current ratio that will allow acceptable Q on ten meters. Fig. 3 presents an estimate of total pi-network input circuit capacitance in two actual parallel 813 amplifiers. As can be deduced from the data, on 10 meters the tank circuit of the bandswitched amplifier will probably get hot enough to boil water.

series pi-network

Since capacitive reactance cancels a like amount of inductive reactance in a series combination of C and L, a method of varying L in a circuit is suggested. If the inductive reactance of L has a certain value, and a capacitive reactance of some smaller value is placed in series with it, the inductance of L will be effectively reduced. If the capacitive reactance is variable, then L can be varied just as effectively as if it were a roller inductor

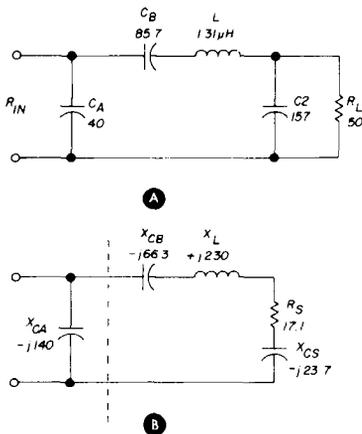


fig. 4. Tuning capacitor may be placed in series with the inductor (A). Equivalent reactance values are shown in (B). This circuit is electrically equivalent to circuit in fig. 5.

or tapped coil.

If the input tuning capacitor of the pi-network (C1) is insulated from ground and placed in series with the inductor the minimum capacitance limitation is virtually eliminated, reducing the total

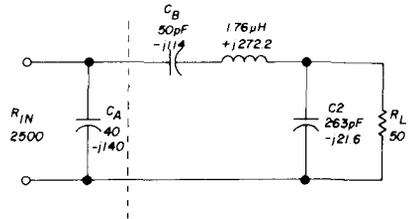


fig. 6. Input impedance of circuit in fig. 5 may be increased to 2500 ohms by changing component values as shown here.

fixed capacitance to no more than 40 pF (see fig. 4).

If the reactances to the right of the dotted line in fig. 4B are added together (since they are in series) the total series

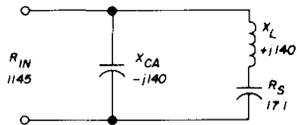


fig. 5. Electrical equivalent of the series-tuned circuit in fig. 5.

impedance is $17.1 + j140$. Thus the circuit of fig. 4 may be simplified to the form shown in fig 5. This parallel resonant circuit is exactly the same result you would obtain if you reduced the standard pi-network to its most simple equivalent form.

At resonance the input resistance to the parallel tuned circuit is

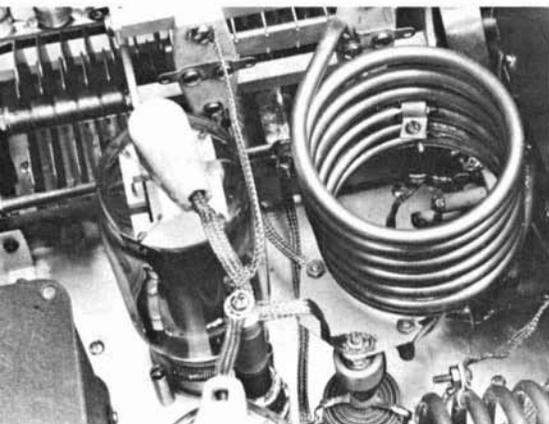
$$R_{in} = \frac{X^2}{R_S} \frac{(140)^2}{17.1} = 1145 \text{ ohms}$$

Tank circuit Q, from eq. 1, is 17.85. An impedance match can be obtained by increasing the value of the tank coil (fig. 4A) to 1.76 μ H and decreasing the series tuning capacitor to 50 pF. The network is matched to the load by increasing C2 to

263 pF. The new component values and their reactances are shown in **fig. 6**.

The equivalent series impedance of R_L and C_2 is $7.85-j18.2$ ohms. Total reactance to the right of the dotted line is still $+j140$. However, R_{Sis} is now 7.85 ohms, so the input resistance, from *eq. 2* is

$$R_{in} = \frac{(140)^2}{7.85} = 2500 \text{ ohms}$$



Bandswitched parallel 813 rf power amplifier using the series-tuned pi network.

The operating Q is 17.85; this value is perfectly acceptable.

In cases where there is enough tube output capacitance plus nominal stray capacitance to equate to the normal input tuning capacitor in a regular pi network the series-tuned configuration warrants prime consideration as an efficient tuning scheme. The series-tuned pi-network has the same number of components as a regular pi-network, and they all have reasonable, attainable values.

There is actually better harmonic attenuation with the series-pi than with the regular pi for a given situation due to the lower shunt reactance to ground on the output side of the network. The network will match a broad range of load resistances for a given Q and plate resistance.

If a Q of 20 is considered the top acceptable limit, the maximum fixed tube output plus nominal stray capacitance for

table 1. Maximum tube output capacitance and stray capacitance (C_a) that can be accommodated by the series-pi-network for a Q of 20.

Rp (ohms)	Xca (ohms)	Ca (pF)		
		28 MHz	50 MHz	144 MHz
1000	50	114	64	22.1
2000	100	57	32	11.1
2500	125	45.5	25.5	8.85
3000	150	38	21.3	7.4
3500	175	29.2	18.4	6.3

a given plate resistance that can be accommodated with the series-pi network is given in **table 1**.

bandswitching

Most of the input capacitance in a normal tank circuit, other than tube output capacitance, is a result of long leads associated with bandswitch circuitry. The preceding network analysis was based on the absence of bandswitch circuitry at the input to the network thereby keeping stray capacitance to an absolute minimum.

You can build a bandswitching amplifier with the series-tuned pi-network if you place the switch on the output side of the network as shown in **fig. 7**. Since there are no particular restrictions on the input tuning capacitance on 40 and 80 meters, a standard pi-network is used on those bands. The lead from the tube plates through the coupling capacitor to the switch adds only a few pF to the stray capacitance and can be virtually ignored.

Additional inductors are switched into the circuit for 15 and 20 meters. The photograph shows the installation of the 10/15-meter inductor. A section of the 20-meter coil can be seen in the lower right-hand corner of the photo.

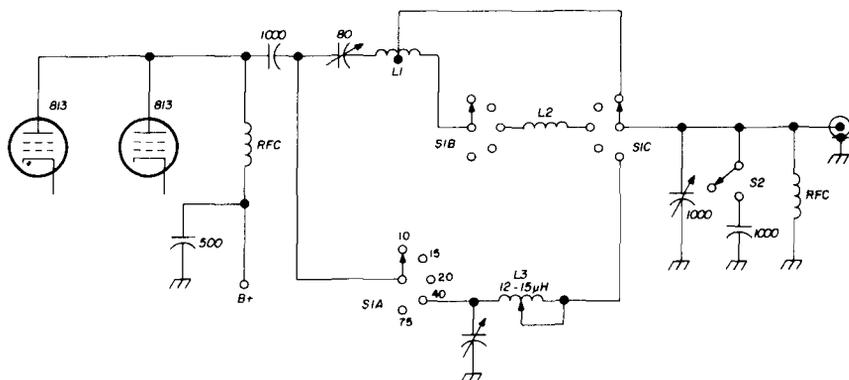
alternate series-pi

For even greater reductions in stray capacitance the series tuning capacitor is placed between the inductor and the load as shown in **fig. 8**. With this arrangement the stray capacitance is on the output side of the network where it becomes a

very small fraction of the total shunt capacitance to ground.

To verify the alternate approach, a two-meter 2E26 transmitter was altered to the series-tuned pi-network con-

circuit was slightly off resonance. The final results with the series-tuned tank were well worth the 30 minutes it took to move the coil and re-solder a few connections.



L1 6 turns 1/4" copper tubing, 2 turns per inch, 3" OD; 10-meter tap 2 turns from output end

L2 6 turns 1/4" copper tubing, 2 turns per inch, 3" O. D.

fig. 7. Using the series-tuned pi network in a multiband band-switched power amplifier.

L3 12 - 15 μ H roller inductor, 500- to 1000-watt rating

figuration in fig. 8. The plate output capacitance of a single 2E26 is 7 pF, exactly the magnitude of tank capacitance required for a Q of 12 with a plate voltage of 300 and plate current of 75 mA.

The original two-meter tank circuit was a push-pull arrangement with a series-tuned link-coupled output. Fig. 9A shows the original circuit; the photograph shows the original circuit in operation with a vtvm indicating *relative* rf voltage across a 50-ohm dummy load.

The same circuit components were reconnected into the arrangement of fig. 8. Fig. 9B illustrates the improvement in efficiency as demonstrated by the relative output rf voltage reading. Every attempt was made to keep all test conditions with the exception of the tank circuit the same for both measurements.

In addition to the improved efficiency of the series-tuned tank circuit there were other advantages. Tuning was much smoother than the original, and there were no rf feedback problems when the tank

design procedure

The following step-by-step procedure will help you design a series-tuned pi network for your own particular requirements. Steps 1 through 4 will indicate whether the series pi is a practical solution to your design problem.

1. Determine tube operating conditions to achieve the lowest possible tube plate resistance consistent with power input requirements

$$R_p = \frac{E_p}{2 I_p}$$

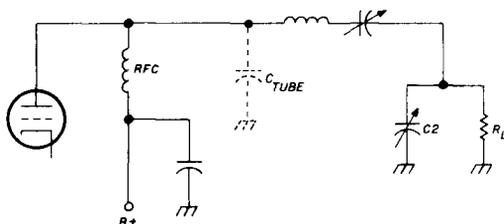


fig. 8. Effects of stray capacitance may be further reduced by placing the series tuning capacitor between the inductor and load.

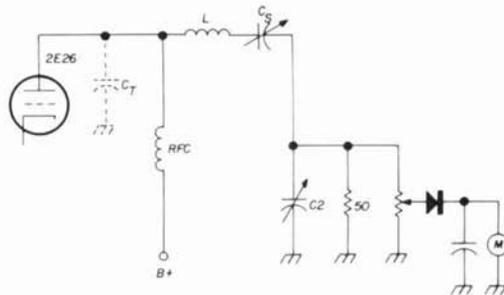
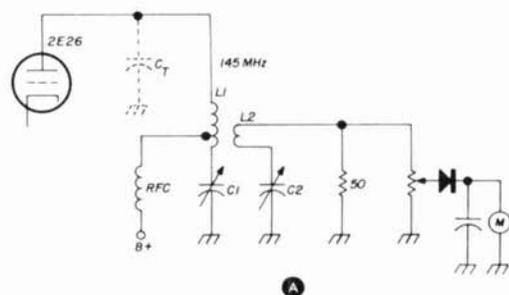
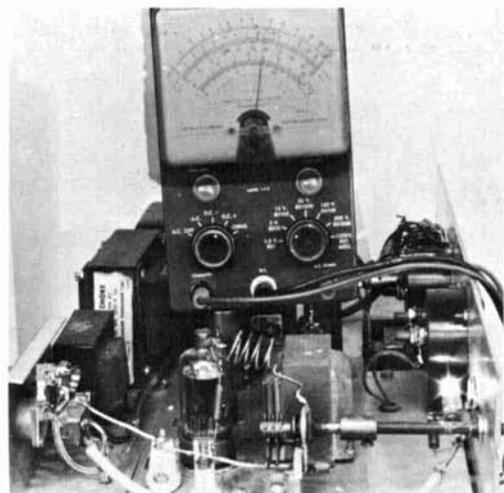
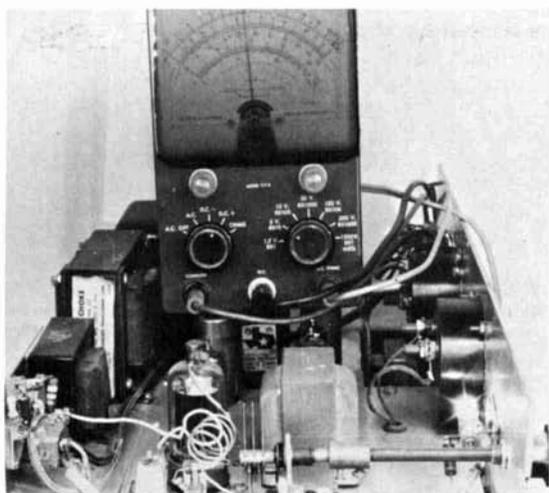


fig. 9. Original two-meter power-amplifier circuit is shown in (A). Series-tuned pi network in (B) provides increased rf output.

2. Determine total fixed input capacitance to the network by adding tube output capacitance to approximately 15 pF stray capacitance

$$C_{in} = C_{tube} + C_{stray}$$

3. Convert C_{in} to its equivalent reactance

$$X_c = \frac{1}{2\pi f C_{in}}$$

4. Compute Q

$$Q = \frac{R_p}{X_c \text{ in}}$$

If Q is not less than 20, re-examine plate resistance R_p and attempt to find a new set of operating conditions to arrive at a lower value.

5. X_c is equal to X_L' , that part of X_L remaining after the reactance of both the series tuning capacitor (X_{cb}) and equivalent series loading capacitor (X_{cs}) are subtracted from the total inductive reactance

$$X_c = X'_L$$

6. Compute the required series output load resistance R_s

$$R_s = \frac{(X'_L)^2}{R_p}$$

7. Knowing R_s , the corresponding X_{cs} portion of the series equivalent load impedance can be selected from fig. 10. This graph is calculated for *only* a 50-ohm antenna at 28 MHz.

8. Select a reasonable value for the series tuning capacitor C_B and calcu-

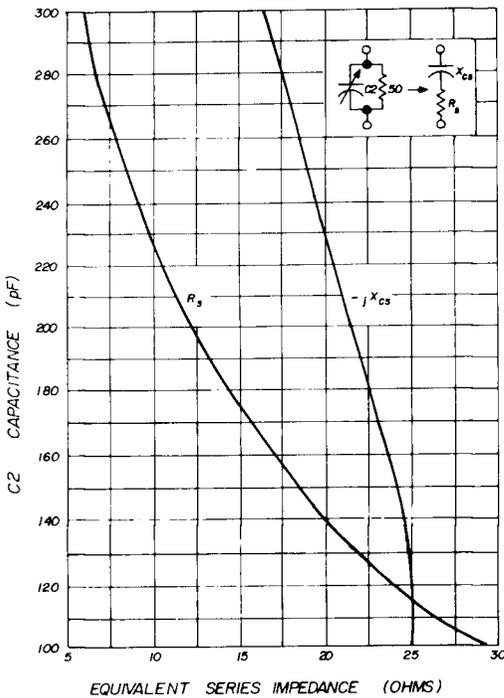


fig. 10. Graph for converting pi network parallel impedance to equivalent series impedance. (Capacitor C_2 in parallel with 50-ohm load at 28 MHz.)

late its reactance using the formula in step 3.

A value of around 100 ohms for X_{CB} is suggested since this results in about 57 pF at 28 MHz for C_B . This is a reasonable midrange value for a tuning capacitor that will cover 10 through 20 meters.

9. Determine X_L by adding the absolute values of X_L' , X_{CB} and X_{CS} (ignore the plus/minus j factors).

$$X_L = X_L' + X_{CB} + X_{CS}$$

10. Calculate the value of the tank inductor from the reactance value determined in step 9

$$L = \frac{X_L}{2\pi f}$$

11. Determine the value of the output

loading capacitor which results in the equivalent series impedance found in step 7.

$$X_{C2} = \frac{R_s^2 + X_s^2}{X_s}$$

The capacitance value of C_2 is calculated with the reactance formula given in step 3.

This completes the design of the series-tuned pi-network. Double check all calculations.

summary

The results of using series-tuned pi-networks has been very gratifying. Tuning and loading with the series circuit is no different than with a standard pi-network. Dial settings for tuning capacitance, inductance and loading capacitance for a resistive 50-ohm load are right on the calculated optimum values.

The series-tuned pi-network should permit the use of inexpensive power tubes in bandswitched linear cathode-driven grounded-grid amplifiers. The problem of plotting a set of design curves for the series tuned pi-network (similar to the pi-network charts in the ARRL Amateur's Handbook) will be left to an industrious engineering student with access to a digital computer.

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ham radio

power in reflected waves

A discussion
of power
in reflected waves
along a
transmission line,
and its effect
upon equipment
operation

Hubert Woods, Calle Las Nubes 1760, Guadalajara, Jalisco, Mexico

Discussions of reflected waves on radio-frequency transmission lines sometimes present the concept that energy in these waves is absorbed by the transmitter; this would adversely affect the operation of the output tuned circuit, or the output tube, or both, by generating heat therein. Of course, if this concept were correct, the result would be the transfer of less rf energy from the transmitter to the load at the far end of the line, usually an antenna. The loss of energy in such case would depend on the magnitude of the reflected wave in relation to that of the forward or incident wave (the swr) and on the proportion of energy in the reflected wave that is absorbed by the transmitter.

As to the proportion of the energy in the reflected wave that is absorbed by the transmitter, opinion varies. Some say that this proportion may range as widely as that of the forward wave at the load end, all the way from zero to 100%, depending on the degree of match between the impedance of the output tuned circuit and the input impedance of the line.

If this match is good, the absorption would be large according to this concept, and nearly all of the energy of the reflected wave would be absorbed. This seems to be a strange situation because the adjustment of the transmitter output circuit is to make the match to the line as good as possible for best forward transfer, whereby most of the energy in the reflected wave would then be absorbed and wasted! (It should be remembered of course that adjustments at the transmitter have no effect on the swr near the load; this is determined solely by the degree of match between the line and the load.)

absorption concept

It may be useful to examine this concept of energy absorption at the transmitter to see whether it stands up under such examination, and to see if there is an acceptable alternative to it. (Hereafter I will refer to this as the *absorption concept*.) To do this, it will be necessary to consider specific transmission lines and impedance-matching

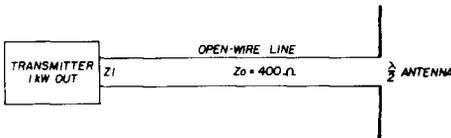


fig. 1. Feeding a half-wave antenna with open-wire transmission line results in a high standing-wave ratio. However, efficiency of this system is very good.

devices such as tuned circuits or their equivalents.

At first we will consider these as idealized circuit elements with no inherent losses, and later consider actual circuit elements with their small but unavoidable losses, to see what effect if any these losses may have on our conclusions. This will make it much easier to discuss the main issue: The possibly much larger losses that may result from high swr if the absorption concept is correct.

First, let us see whether the absorption concept leads to results that are found in practice. Consider fig. 1. There we have a transmitter with a power output of 1 kW with a tuned output circuit capable of being tuned to resonance and of providing a good impedance match between the plate resistance of the tube and the input impedance Z_i of the parallel-conductor open transmission line. This line, which is specified to have a characteristic impedance Z_0 of 400 ohms, feeds the end of a half-wave antenna so there are standing waves on the line because of a large mis-match between the line and the antenna feed point.

The swr could be very high, well above 10, but we will assume that it has the

reasonable value of 9. Because of these standing waves the line input impedance Z_i , as seen by the transmitter, may depart considerably from the nominal value of 400 ohms, depending on the electrical length of the line; it may be larger or smaller and it may exhibit reactance as well as resistance. Whatever this line input impedance may be we will specify quite reasonably that the transmitter output circuit provides a good match for it. Thus, according to the absorption concept the transmitter absorbs nearly all the energy in the reflected wave. This, calculated as usual, turns out to be represented by a power of 640 watts!

Many amateurs, including myself, have successfully operated amateur transmitters using similar feed lines which were, in all essential respects, equivalent to the arrangement shown in fig. 1 without the slightest evidence that anything approaching 64% of the energy going into the transmission line was being returned to the transmitter, appearing as heat in the tuned circuit or tube or both. If even a fraction of that energy was coming back into the transmitter it would have been immediately evident to a knowledgeable amateur that something was very wrong.

Some amateurs have experienced overheating in amplifier tuned circuits or tubes; this can easily happen if the tuned circuits are not properly designed or built, and particularly if the coupling between the tuned circuit and the line is not sufficient to bring the loaded Q of the circuit down to the proper value, usually about 12. With insufficient coupling the circulating current in the tank circuit can rise to a high value and may damage the inductor or other parts. However, as such, this is not the fault of the high swr.

If the absorption concept were correct these parallel-conductor open lines with their high swr would give incessant trouble. In fact, these same open transmission lines are very efficient in spite of their high swr. Thus, the absorption concept is not confirmed by qualified experience.

power factor

Now consider **fig. 2**. This arrangement uses the same transmitter on the same frequency but with output circuits adjusted or modified to match a short 50-ohm coaxial cable which is connected to the input of an impedance-matching device. This device accurately matches the 50-ohm cable to whatever may be the input impedance Z_2 of the same open-wire transmission line as used before with the same swr of 9. (Z_2 may not equal Z_1 because the transmission line is not necessarily the same electrical length.)

The transmitter cannot now absorb any energy that may exist in the reflected wave in the open transmission line for there is no reflected wave in the 50-ohm cable connected to the transmitter since the 50-ohm line is perfectly matched at both ends.

Where does any such energy, if it exists, appear? It cannot disappear. Are we to believe that it must appear as heat in the impedance-matching device? That would be doing the designers and constructors of such devices a grave injustice because it is readily feasible to design and build such devices having a very much smaller loss than seems to be implied in our example.

The small residual loss in matching between lines can be reduced nearly to zero by using a suitable tuning stub or in some cases a quarter-wave linear transformer, in place of a network with lumped constants. We cannot reasonably believe that any considerable proportion of the calculated power in the reflected wave on the open-wire line is resulting in the dissipation of heat in our matching device. Then where is it?

The simple answer is that it never existed, and therefore, there is no problem of power loss or dissipation of heat to be solved. The only real power in the transmission line (barring the previously mentioned small inherent losses) is that which results in the dissipation of energy in the load. That is equal to the difference between the *indicated* values for forward and reflected power.

It is usually understood that the forward wave and the corresponding reflected wave are each made up of two associated waves, one of current and the other of voltage. The usual discussion proceeds to show how the forward voltage wave adds to and subtracts from the reflected voltage wave, taking phase as well as magnitude into account, to form a voltage standing wave.

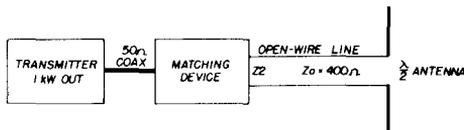


fig. 2 Antenna matching device provides good match to transmitter although there is still a high swr on the line to the antenna.

Similarly, it is shown how the forward and reflected current wave interact in the same fashion to form a current standing wave. What is not always made clear is that the two resultant standing waves (one of voltage and the other of current) have a phase difference of 90° . The importance of this fact becomes clear when it is recalled that in an ac circuit power is equal to voltage times current times the power factor:

$$P = EI \times \text{power factor}$$

Power factor is the cosine of the phase angle and the cosine of 90° is zero. Therefore, the power in standing waves along a transmission line is zero.

In the case of a lossless line (either shorted or open at the load end or ending in a pure reactance) reflection is complete. The two standing waves constitute the entirety of electric waves on the line. That is, the only current in the line is that in the standing wave of current and the only voltage on the line is that of the standing wave of voltage. As a result of the phase angle of 90° between the two powers in the line of zero.

The usual directional rf wattmeters do not take power factor into account; in the case of our lossless line with complete

reflection, the rf wattmeter will indicate equal forward and reflected "power," their difference being zero. An analogous situation exists if we connect an ideal (lossless) capacitor to an ac source. An ac voltmeter across the capacitor will indicate volts and an ac ammeter in series will indicate amperes, but there is no energy dissipation in the capacitor, and therefore, no power in the connections to it. The phase angle between the voltage and current waves in a capacitor is 90° so the power factor is zero.

Next consider a lossless transmission line terminated in a load that is purely resistive and equal to the characteristic impedance of the line. Under these conditions there is no reflection, no standing waves, and all the energy into the line appears as energy into the load. The power factor for the line in this case is 1.0. That is, the current and voltage waves coming from the transmitter to the load are in phase.

Another termination for our lossless line is one having a resistance differing from the characteristic impedance of the line or having some resistance and some reactance. such cases there will be some reflection but some energy will be dissipated in the load. The analysis of this situation is somewhat more complex but will be facilitated by considering that a *part* of the total line current is assignable to the standing wave of current and that a *part* of the voltage on the line is likewise assignable to the standing wave of voltage. These portions of the total line current and voltage are 90° out of phase as in the previously described cases and therefore do not represent power. The remaining current and voltage are in phase and represent real power which results in dissipation of energy in the load. Thus, the *only* power in the line is that which flows from the transmitter toward the load.

The line as a whole will have a power factor at the input end of just the value needed to satisfy the relation

$$\text{power factor} = \cos \theta$$

where the angle θ is the phase angle

between the total line current and total line voltage measured at that point. This same statement holds even if the line has some losses. The only power in the transmission line at its input end is that power which accounts for line losses as well as for energy dissipated in the load.

conclusion

The concept of reflected "power" is a useful fiction to help us visualize the formation and nature of standing waves, but it can get us into trouble if taken too literally. If you insist on considering the reflected wave as real power then you must adopt another fiction, namely that when it gets back to the transmitter it is *completely* re-reflected toward the load. (The alternative is the absorption concept which I hope by now has been given a decent burial.) Complete re-reflection of power at the input end is impossible to accept since the necessary conditions of impedance mismatch are not present. This is perhaps an example of one untruth, or even two, never being enough!

Although open transmission lines were used in the previous discussion, exactly the same arguments apply to systems using coaxial lines and result in exactly the same conclusion: Power in a transmission line flows in one direction, from transmitter toward the load.

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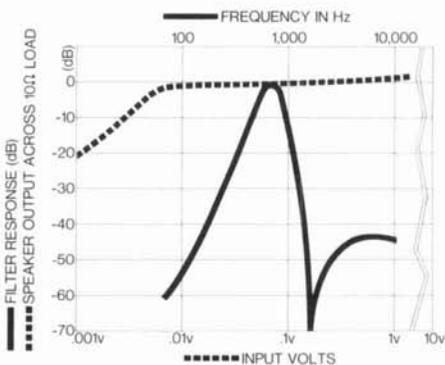
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Although there are many swr indicator circuits and construction techniques, most are unnecessarily complicated. The one described here is simple and easy to make, requiring no special hardware. In addition it provides, on one meter, simultaneous indications of both forward and reflected voltage, leaving both hands free to adjust transmitter controls.

operation

The indicator (fig. 1) is somewhat more intricate than most, but its advantages outweigh its complications.* Its operation is based on the use of a zero-center dc microammeter. The diodes are connected to give opposing polarities, unlike most circuits, so that FWD gives an indication to one side of the meter; REF to the other. When the switch is in FWD or REF position, operation is the same as with other bridges. In the center position, the voltages are combined in a resistive divider, and the meter gives an indication of the relative predominance of either forward or reflected voltage. Be sure to connect the switch so that forward voltage will be read in the same direction as the switch when it is turned to FWD.

construction

The pickup line is the result of a search for an easy-to-make, effective, and small unit. It has been described before.² It uses the inner conductor of a piece of RG 58/U or RG 11/U.

The pickup line is constructed as shown in fig. 2. Tin the edges of the hole in the braid to prevent shorting to resistor R_T , then slip the inner part of the coax inside the braid. Solder the ends of the

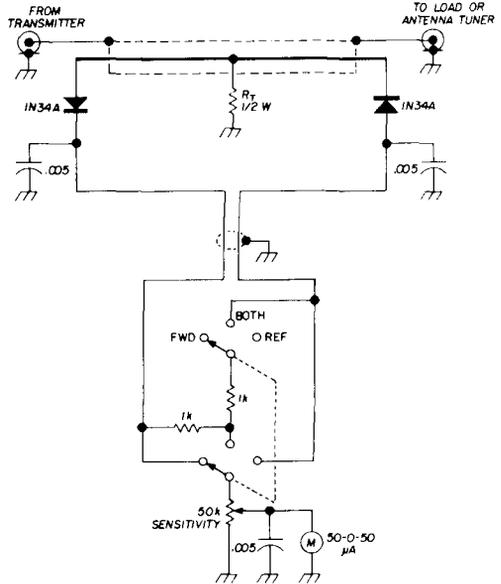


fig. 1. Schematic of bridge and indicator. Resistor R_T is equal to the value in ohms of the load impedance (see text).

*This design is similar to the "Monimatch" directional coupler,¹ which uses a capacitance-resistance bridge. The value of the terminating resistor, R_T , is critical and must be determined experimentally to obtain good bridge balance (null). The adjustment procedure and method of determining R_T in reference 1 should be followed before attempting to operate the instrument. Editor.

Gregory P. Widin, WB2ZSH, 108 Valley Drive, Watchung, New Jersey 07060

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braid to pieces of stiff wire, which go to the connectors for the transmitter and transmission line. Next the diodes and resistor R_T are soldered to the ends and center, respectively, of the pickup inner conductor.

The pickup line and diodes were built in a shielded box, which connects in the transmission line between transmitter and antenna, or between transmitter and antenna tuner. The diodes I used were

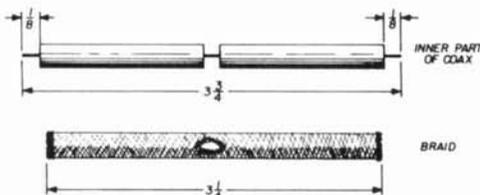


fig. 2. Construction of pickup line. Inner part of coax cable should be insulated where it protrudes through hole in braid; check with ohmmeter to make sure wire and braid are not short circuited.

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Hoffman type 1N261, but the general-purpose 1N34A will do. A shielded two-conductor line was run to the indicator unit, which was built in an unshielded bakelite minibox. This box did not seem to pick up any stray rf. The shielded line need not be rf-type cable, since it carries only dc. The entire unit could be built in one box to eliminate a separate box and cable.

An ordinary 0-50 dc microammeter was tried and worked in the combined circuit without the switch. This would eliminate both the switch and a zero-center meter.

The bridge seems to have adequate sensitivity. Construction is straightforward and uncomplicated. Best of all, using it is as simple as tuning for maximum on the meter.

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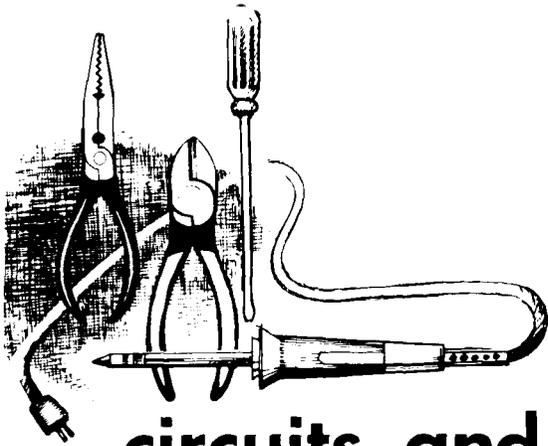
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circuits and techniques

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experiments with phase-locked loops

Do you like to experiment? To improve? You can start now with these little magic boxes, phase-locked-loop integrated circuits. I covered the fundamentals in last month's column; now you can get involved with some practical circuits.

no tuned circuit

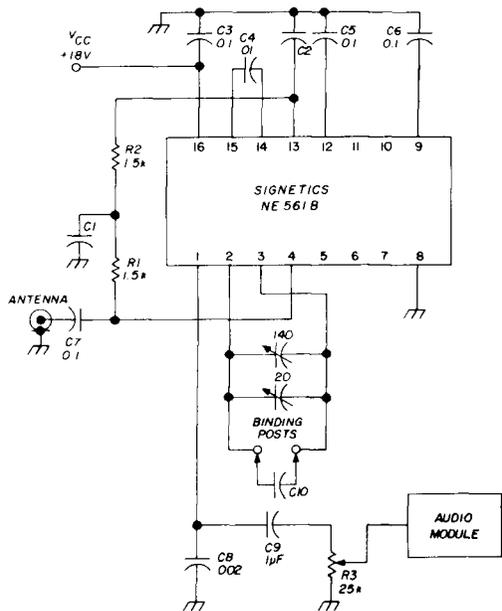
Can you conceive of a tunable receiver without a single resonant circuit. Take a look at fig. 1. All of the signals are fed into the magic box in one lump. The signal that is demodulated is the one with its carrier phase-locked to the voltage-controlled oscillator of the integrated circuit. You pick the station you wish to receive by changing the frequency of the oscillator. Even this oscillator does not have a resonant circuit because its frequency is determined by resistor-capacitor time constants.

The oscillator is stable if adequate carrier signal is applied to the phase-comparator, fig. 2. The output of the vco

IC is supplied to a small solid-state audio module with an output rating of 1 watt; loudspeaker is a 4-inch pm unit. More than adequate volume level is obtained when using a good antenna system.

At this point in my experiments the unit has been used on the a-m broadcast band, the 160-, 80- and 40-meter amateur bands and the 31- and 49-meter short-wave broadcast bands. Vco tuning capacitor values and quadrature capacitor values are given in fig. 1.

Two variables, a 140 pF and a 20 pF, are mounted on the breadboard. The smaller capacitor is used for bandspread tuning. Two binding posts are included for convenience in changing the fixed-value capacitor associated with vco tuning. For operation on the broadcast band and 160 meters a 365- or 400-pF variable is appropriate. When using the values shown complete broadcast band



	C1	C2	C10
frequency	pF	pF	pF
BC band	150	150	200
			300
160 meters	100	100	150
80 meters	50	50	none
40 meters	50	50	none

fig. 1. Integrated-circuit phase-locked-loop receiver uses no resonant circuits.

operation is obtained by using several fixed capacitors, inserting them across the two binding posts when shifting among segments of the band.

A good antenna is important. A 160-meter dipole is excellent for 160 meters and the broadcast band; it also does well on 40 and 80 meters. However, dipoles cut for 40 and 80 meters do better on these bands as well as the shortwave broadcast bands.

If there is an overpowering local broadcast station that bulls its way through you may have to experiment with antenna lengths or install a resonant trap ahead of the antenna input.

more signal and selectivity

A simple resonant transformer at the input improves selectivity and blocks out any overpowering local, **fig. 3**. The low-impedance primary link matches a low-impedance antenna system. The secondary is resonant and is connected to the ic input by a low-value capacitor, C7. This capacitor prevents the low-impedance ic input from loading down the resonant circuit, insuring greater selectivity. A double-tuned input transformer would offer even greater selectivity.

When using a short random-length antenna a simple tuner such as the simple T-network in **fig. 4** will deliver more signal. It is inserted between the receiver input and the antenna system.

More sensitivity. An amplifier ahead of the receiver (fet, bipolar or ic) will increase the sensitivity. However, the signal delivered to the PLL should not

exceed 0.5 volts rms. Next month I will discuss my experiences with an amplifier.

Preset frequency operation. The PLL system lends itself to preset operation for particular stations on the broadcast band, WWV reception, net frequencies, etc. You need only use a switching arrangement and preadjusted trimmer capacitors as shown in **fig. 5**. You can then very quickly switch off the tunable position to one or more preset receive positions.

Idiosyncrosies. Everything is not ideal. Sufficient signal must be delivered to the input. Under weak signal conditions there is a swishing sound which results from the difference frequency between carrier and vco when there is an unstable lock.

There is also a hand-capacitance effect. This results from non-grounded capacitive tuning of the vco. Use a vernier dial and insulated shaft on the fine-tuning variable. The phase-locked loop ic also lends

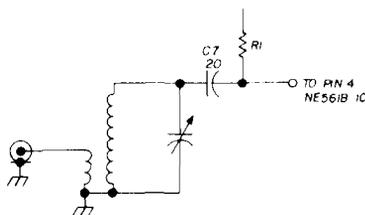


fig. 3. Resonant input circuit for improved selectivity.

itself to frequency change by voltage change, especially for fine-tuning. This may eliminate hand-capacitance effects. Suitable circuits will be discussed next month.

At higher frequencies input resonant circuits and tuning influence the vco frequency. It appears that the ideal arrangement would have an untuned amplifier just ahead of the ic input. Perhaps a better plan would be to use an isolating no-gain stage at this position and precede it with a tunable gain amplifier.

The PLL device in **figs. 1 and 3**, is suitable only for a-m demodulation, not cw or sideband. However, it can be used as a practical fm detector. A simple

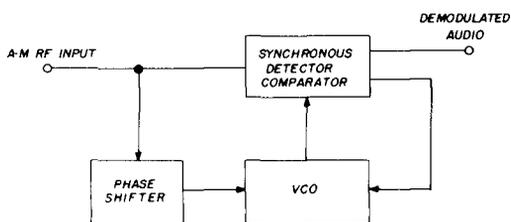


fig. 2. Basic phase-locked am demodulation scheme.

switching arrangement would permit selection of either a-m or fm, making it ideal for 6- and 2-meter operation.

theory of operation

Important considerations in setting up the external circuit for the PLL synchronous a-m detector are the selection of the vco tuning capacitance and choosing appropriate values for the two RC combinations that establish the proper 90° phase relationship between the incoming carrier and the vco.

A very simple formula permits you to select a suitable value for the vco tuning capacitor:

$$\text{vco capacitor (pF)} = 300/\text{frequency in MHz}$$

Values should be calculated for minimum and maximum frequency you want to cover. For example, if the receiver is to tune between 1 and 2 MHz, the minimum and maximum capacitance values would be 150 pF (for 2 MHz) and 300 pF (for 1 MHz). A 125-pF fixed capacitor and

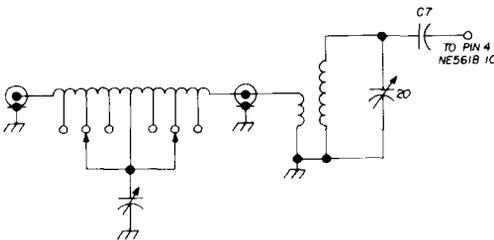


fig. 4. T-network antenna tuner for random-length antennas.

200-pF variable would provide band coverage plus a little bit of overlap at each end.

A net 90° phase shift is obtained by connecting two RC combinations in cascade. Each contributes a 45° shift. The phase shift of 45° is obtained at the frequency where the resistance and reactance are equal.

$$RC = \frac{1}{2\pi f}$$

In the actual calculation the frequency, f , is selected as the median frequency between the two desired frequency extremes. In our example this

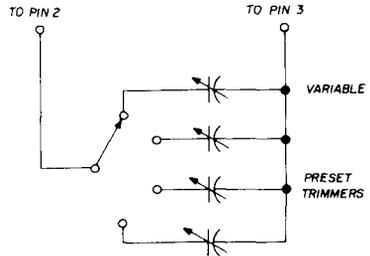


fig. 5. Presetting the phase-lock IC to specific frequencies.

frequency would be 1.414 MHz or:

$$f_0 = f_H f_L = X 1 - 1.414$$

Where f_H and f_L are the high and low frequencies respectively.

If a resistor value of 2000 ohms is selected, the value of the associated capacitor must be:

$$C = \frac{1}{2\pi f_0 R} = 56 \text{ pF}$$

Vco locking takes place over a considerable angular range; specifications are 90° ±30°. Therefore, it stays in lock between the high- and low-frequency tuning limits provided these limits are not spread too far from the median frequency. This means that the phase-shift of each RC combination should not be greater than ±15°. At these extremes the phase angle values would be 60° (45 + 15) and 30° (45 - 15). From a natural function table it can be seen that the tangents of 30° and 60° are 0.577 and 1.732 respectively. In practical applications this means that the reactance of the capacitor at the highest frequency should not be less than 0.577R, and the reactance at the lowest frequency should not be higher than 1.732R.

In our example, then, reactance at 2 MHz should not be less than 1154 ohms

(0.577×2000). At 1 MHz the reactance should not exceed 3465 ohms (1.732×2000). Calculations for 1 and 2 MHz show that 56 pF exhibits 2843 ohms reactance at 1 MHz and 1422 ohms

The Attwood technique can be used with conventional amplitude-modulated rf signals but is especially adaptable to suppressed carrier and sideband modulation modes. Experimental work was done

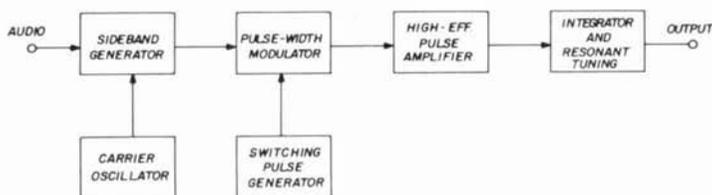


fig. 6. Basic arrangement for the Attwood high-efficiency switching modulator.

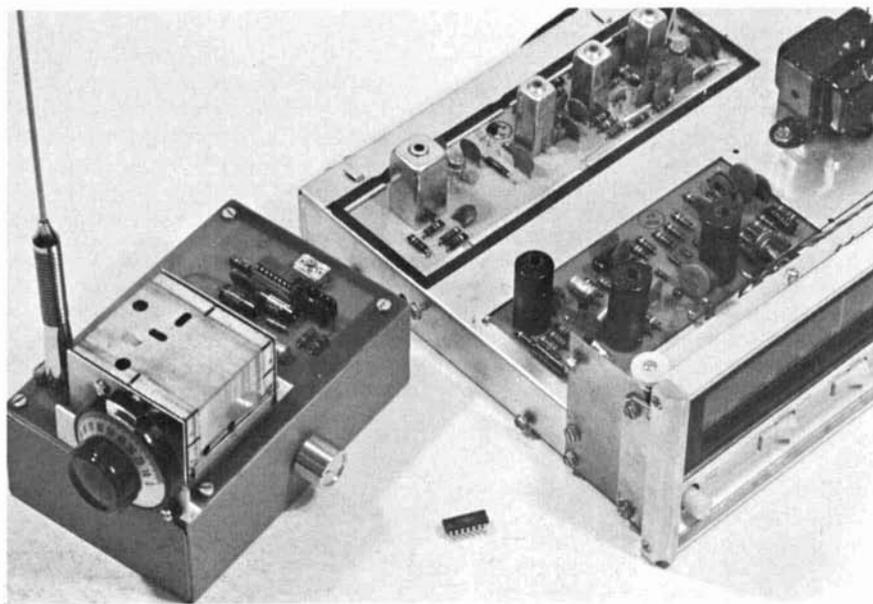
at 2 MHz. These values are well within the angular locking requirements.

switching and linear amplification

A linear amplification system with efficiency as high as 90% has been developed by Brian Attwood of Mullard of England. This is indeed a startling figure when you consider the difficulty in obtaining 50% efficiency with conventional class-AB linear amplifiers.

with solid-state devices but the idea is just as appropriate for vacuum-tube applications and perhaps even more so for hybrid combinations of semiconductors and vacuum tubes.

In the Attwood process the modulated signal is initially formed at low power levels in a conventional manner. The block diagram in fig. 6 shows the functional plan for a sideband transmitter. It includes the usual carrier generator and



The phase-locked loop IC at the bottom center replaces the entire i-f/discriminator section outlined in black on the receiver chassis to the right. The unit at the left was built in the Signetics applications lab to demonstrate the reduction in size when a PLL is used in an fm tuner.

follow-up sideband modulator and mixer; the modulated signal at the transmit frequency is applied to a pulse-width modulator.

Two switching stages perform an additional modulation function. First there is a switching frequency generator which operates at a frequency a number of times higher than the transmit frequency. For good results this generator should be five times carrier frequency or higher although the system will function with the switching frequency only twice the carrier frequency. The switched modulation system produces an output that has the duration of its pulses varying with the modulating information.

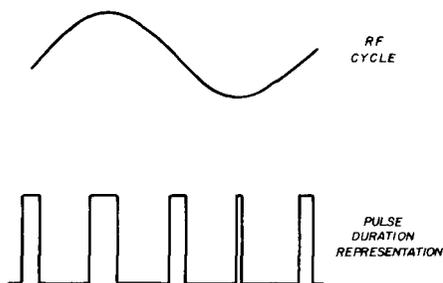


fig. 7. Example of pulse-width modulation.

A simplified drawing of a pulse-width modulation process is shown in fig. 7. As the modulated rf wave varies with modulation the pulse-width modulator generates a train of high-frequency pulses, with pulse width varying with modulation.

As an example, consider the single-tone modulation of a lower-sideband transmitter with a carrier frequency at 3.9 MHz. Under this condition a single rf wave with a frequency of 3.899 MHz (3.9 MHz minus 1000 Hz) is generated. As the 3.899 MHz rf wave goes through its cycle the pulse-width modulator produces a series of pulses. When the rf wave is on its positive crest the width of the output pulse is greater than the pulse that represents its negative trough. In fact, the width of the pulse varies in accordance with the instantaneous amplitude of the

rf wave on each side of its zero axis. Of course, if there were voice modulation there would be complex pulse duration changes which would follow the amplitude gyrations of speech.

Note that pulse output has constant amplitude. All the desired information is in the form of pulse duration changes. The succeeding amplifier can then be made to operate at high efficiency because it can be designed to function as a *high-powered pulse amplifier*; pulse levels can swing between cut-off and saturation limits. The switched nature of the information conserves average power and results in greater power-handling capability and efficiency.

In an amateur transmitter it is conceivable that all stages prior to the final are solid-state; the final would be a high-powered vacuum-tube amplifier. This stage might use high-power television sweep tubes which are designed primarily for pulse and nonsinusoidal power amplification.

More power output could be obtained than is now possible using the same tube types as class-AB linears (common practice in many ham transmitters). The technique might lend itself to the design of mobile equipment with high output light-weight and minimum power demand.

After the pulses are amplified they are not transmitted; the signal is converted back to the form it had before it was introduced to the pulse-width modulator. This can be done with a suitable resonant system since resonant circuits have energy-storing ability and function as effective integrators. Thus, the pulse information is stretched out and converted to sinusoidal form.

The output circuit, of course, must be designed to remove and attenuate any switching frequency components which are on a higher frequency than the transmit frequency. It is conceivable that a suitable multisection pi-network output system would do a satisfactory job. Perhaps an m-derived addition may be necessary to thoroughly notch out the switching frequency.

Because of the requirement for the high switching frequency this modulation mode is currently most suitable for low-frequency operation. However, vhf possibilities exist when you consider that low-cost ics are now available with switching rates up to 500 MHz or so.

version 1 used one of the commonly available 1-watt output audio modules. A transistor audio-output transformer connected in reverse (low impedance to high impedance) permitted the necessary voltage drive and reasonable match to the source input circuit; the center tap of the

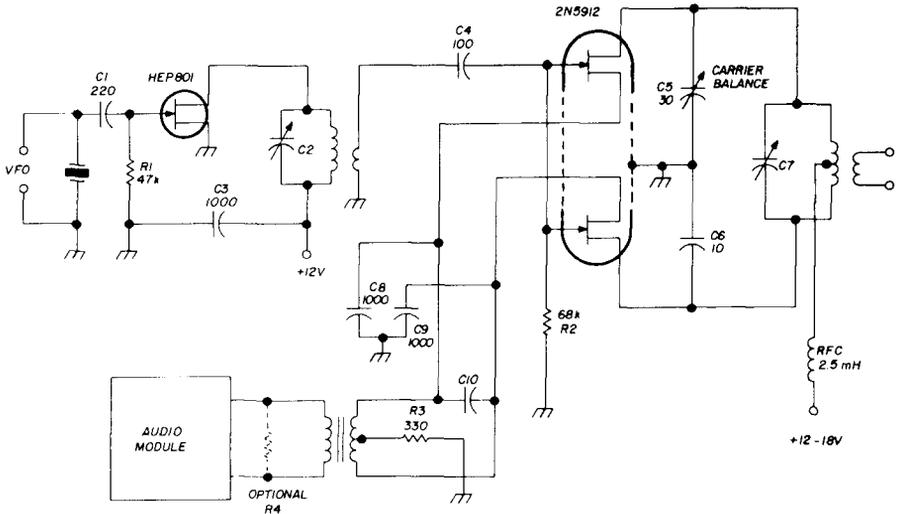


fig. 8. Dual-fet balanced modulator.

dual-fet balanced modulators

A dual field-effect transistor consists of two fet's with identical characteristics mounted in the same case. Such units are designed mainly for use in differential amplifiers, but they are also ideal for balanced modulator and demodulator circuits. Devices are available which function up to several hundred megahertz.

An effective circuit that I have operated on 80 through 10 meters using plug-in coils is shown in fig. 8. In this circuit the paralleled gates of the Siliconix 2N5912 dual fet are supplied with carrier from a crystal oscillator. The drains are connected in push-pull to obtain carrier cancellation. Removal of the crystal from the oscillator permits use of a stable vfo. Output level is correct for driving a low-power vacuum-tube linear amplifier using a 6AK5, 6BA6, etc.

The modulating signal is applied to a push-pull source circuit. In my pegboard

output winding permitted a balanced feed system. A capacitor across the secondary (C10) is used to control the high-frequency audio roll-off.

The average audio module may supply more audio than is necessary. If this is the case a loading resistor (R4) of 12-ohms or higher (depending upon how much the audio amplitude must be cut back) can be connected directly across the output of the module. The microphone gain control is connected between the mike terminals and the audio module input.

Carrier balance is handled by capacitors C5 and C6. The trimmer capacitor is adjusted for minimum carrier output; carrier suppression is excellent. No carrier reduction control was included in the source circuit although some additional suppression may be possible. However, if the dual fet's are really identical it should not be necessary.

ham radio



32S-3 audio

The two audio amplifier stages in the Collins 32S-3 exciter run full-blast, inasmuch as the *mic gain* control follows them. With high microphone levels the second audio stage is overdriven and adds distortion; this can be verified by placing a voltmeter between V1B pin 6 and ground. This distortion may be reduced by allowing slightly higher *mic gain* settings, and adding desirable degenerative feedback in the audio stages to reduce the level at which the second audio stage operates.

Collins Service Bulletin No. 2 (dated 21 September 1967) says, "disconnect and discard the bus wire connected between tube socket XV1, pin 8, and switch S8C, pin 2. Replace the bus wire with a new 680-ohm resistor, R1." R1 is ½ watt. This change was incorporated in the fifth edition of the 32S-3 transmitter instruction book.

Incidentally, Collins Service Bulletin No. 1 was revised on 13 October 1967, making a number of changes. After tests, I restored the 1-megohm resistor R101 in the first audio stage, V1A, rather than use the 100k resistor. Many of the changes already were incorporated in my equipment, which is covered by the third edition of the instruction book (dated 15 September 1964).

My attention was called to another change to increase degenerative feedback. The third edition of the instruction book shows a 20- μ F capacitor, C183, between

pins 10 and 11 of switch S9F and ground. This capacitor bypasses the second audio cathode resistor, R4. It did not appear in the fifth edition of the book; Collins recommended that it be disconnected. The capacitor is located conveniently behind the words *freq control* on the upper left of the front panel so the change can be made without removing the set.

If there still is excessive audio gain Collins has recommended a ½- or a 1-megohm resistor between pins 1 and 6 of the audio stages, V1. This is between the two plates. I tried it using a Vector test socket under the tube, and found that it is desirable, resulting in excellent audio quality after removing the capacitor mentioned above. The level in the second audio stage was reduced to a distortionless level and the *mic gain* was at a reasonable half-scale position. I had been using the Electro Voice 676 dynamic cardoid microphone, with 10-dB roll off.

Bill Conklin, K6KA

narrow-shift RTTY reception with Heath SB receivers

To receive 170-Hz narrow-shift RTTY signals with the Heathkit SB-100, SB-101, SB-300 or SB-301 using the optional 400-Hz filter, the bfo crystal must be 3393.19 kHz. In the SB-series receivers the lower-sideband crystal is 3393.6 kHz; the frequency may be lowered to

3393.19 kHz by placing a small capacitor from the grid of the oscillator tube to ground. This shunt capacitor must be chosen so that both the *mark* and *space* tones have equal amplitude; 20 pF worked satisfactorily in several sets tried here.

In the SB-300 and SB-301 receivers the 3393.6-kHz crystal must be placed in the upper-sideband crystal socket so it will be selected when the 400-Hz filter is switched in.

Robert Clark, K9HVW

dynamic transistor tester

If you already have an oscilloscope on your workbench here is a simple dynamic transistor tester which will measure both in- and out-of-circuit transistors. Total cost of the device is less than \$5. In addition, this tester will check diodes, although it cannot be used with mosfets, junction fets or uhf varistors and diodes.

In the circuit in **fig. 1** a small ac voltage is fed to the transistor junction. This alternately forward and reverse biases the junction. With the oscilloscope test leads open-circuited a horizontal trace is displayed; when the test leads are shorted a vertical trace is shown. A good transistor junction, either emitter-to-base or base-to-collector, when connected between the two test leads, will show a sharp-cornered trace as shown in **fig. 2D**. A rounded corner (**fig. 2E**) indicates

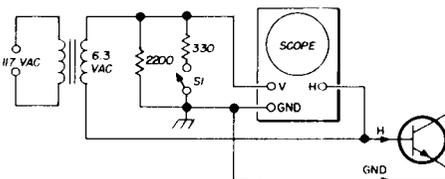


fig. 1. Simple dynamic transistor checker tests both in- and out-of-circuit transistors and diodes.

leakage current. Junction resistance shows up as a sloping vertical trace (**fig. 2F**).

To calibrate the oscilloscope, switch

the horizontal sweep to *external*, plug in the test leads, and with the leads open-circuited, adjust the horizontal control

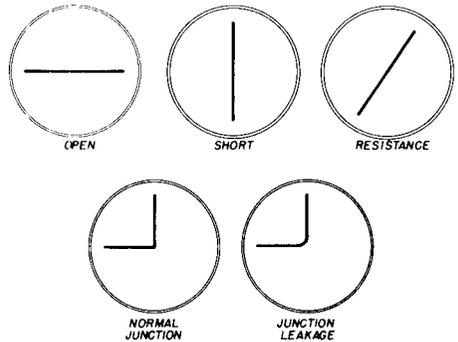


fig. 2. Basic traces obtained with the test circuit of **fig. 1**.

for approximately 2/3 scale deflection. Now short-circuit the test leads and adjust the vertical control for approximately 2/3 scale deflection. The equipment is now ready for testing.

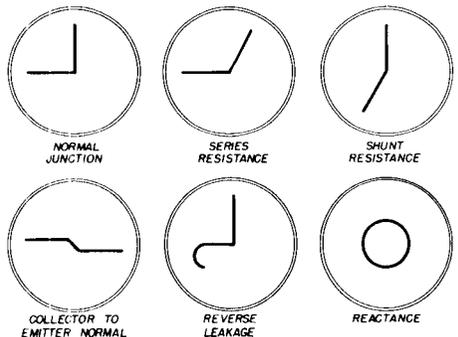


fig. 3. Typical circuit traces provided by the dynamic transistor tester.

Power transistors require more test current—this is provided by closing switch S1. Most in-circuit transistors will indicate some resistance such as shown in **fig. 3B** or **3C**. This transistor tester will not check an in-circuit transistor when the impedance across the junction is extremely low.

Vern Epp, VE7ABK

radio-controlled morse sounder

The circuit shown in fig. 4 provides operation of a Morse telegraph sounder off the air. There are many amateurs who are also Morse operators and would like to hear the music of a telegraph sounder again. With this circuit the relay picks up with the reception of a signal.

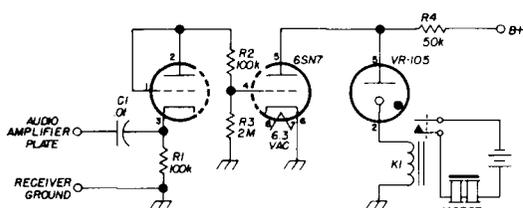


fig. 4. Circuit for operating Morse sounder off the air. K1 is 10,000-ohm spst relay.

The first half of the 6SN7 is connected as a diode rectifier. The incoming audio signal is rectified to bias off the second half of the 6SN7, which is a clamp tube. When the tube is clamped off, current flow stops, and the voltage at pin 5 of the VR105 increases, igniting the VR tube and causing the relay to close. If more audio gain is desired a one-stage amplifier could be inserted between the receiver and the first half of the 6SN7 although this should not usually be necessary.

For proper operation, current flow through the relay should be measured between the relay coil and ground. The value of R4 may have to be reduced to approximately 40k for snappier sounder operation. Capacitor C1 may be increased to 0.1 μ F if desired. Total current drain from a 250 to 300-volt receiver power supply is about 50 mA.

Jack Proefrock, K6QEQ

nylon guy rope

Most amateurs who have experimented with antennas realize the problems generated when steel guy wires are used in close proximity to the antenna they

support. Swr and pattern distortions are just the beginning.

After a recent residence move I decided that the only location on the property suitable for a tower was the roof. This meant that guying would be essential. In an effort to bypass the usual interference problems I considered two possible remedies. One was the use of strain insulators to break up the guy wires into non-resonant lengths. I ruled this out for several reasons: time involved, added potential weak links and the increased cost.

The second consideration was the use of nylon rope. I finally decided on a rope with a working strength of 148 pounds (in excess of 750 pounds test). My

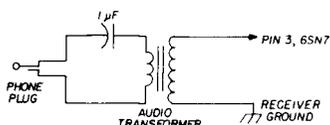


fig. 5. Interface circuit for plugging Morse sounder circuit into your receiver.

24-foot aluminum tower went up from its tilt-over base without a hitch. However, when it was time to begin tightening the nylon guy ropes, the first hint of trouble appeared. We couldn't tighten each leg enough; the nylon, especially the end knots, was stretching.

With hopes of rectifying the situation, we switched from knotted ends to split bolts. This helped, but not as much as I would have liked. Even so, I decided to give it a trial run. I kept the installation *status quo* for the entire winter and found that nylon was every bit as strong as first anticipated, and held in winds gusting over 90 mph. The nylon guys showed little, if any, signs of weathering; the split bolted ends were still secure.

After the tower was up a short time I discovered that strength was not the only criteria to consider. Although the tower guys held safely in high winds, stretching became more and more evident. The tower and array often swayed violently as

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if suspended by six huge rubber bands. This generated a lot of unpleasant noise and may have weakened the base mount. Needless to say, this sort of installation was unacceptable.

As is often the case, the eventual cure was a compromise. Nylon rope was used for the top set of guys, while steel was chosen for the lower supports. This arrangement reduces noise and provides acceptable strength with minimum guy-to-antenna interaction.

Morrie S. Goldman, WA9RAQ

drill guide

While building various projects I occasionally find it necessary to accurately drill a hole in the center of a 1/4-inch volume-control shaft. I first obtain a short length of soft-iron stock, 1-inch square or smaller, depending on what is available; then I drill a hole through the iron with a small drill (7/64-inch or 6-32

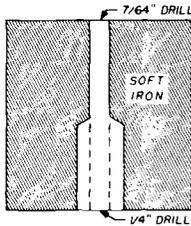


fig. 6. Simple jig for drilling hole in the center of a 1/4" shaft.

tap drill). This can be done with a hand drill.

Using this small hole as a pilot hole, I drill halfway through the metal with a 1/4-inch drill. This completes the drilling jig. To use the jig, insert the 1/4-inch shaft into the larger hole and drill through the smaller. Centering accuracy is quite good with this technique and it does not require any elaborate tools.

This method is suitable for drilling other size shafts as well. Simply choose the correct size drills for each of the holes.

Felix W. Mullings, W5BVF

improving linear amplifier performance

In a recent article¹ it was noted that unsatisfactory performance of transistorized equipment could often be traced to poorly regulated power supplies. This is also true of linear amplifiers. The point was very much stressed when ssb first became popular but is often overlooked by the new generation of amateurs. My experiences in performing a simple test for linearity and observing the increase in power output that resulted from improving power supply regulation may be instructive.

My linear, the Heath HA-10, had a single 8- μ F capacitor in the power supply. Rf output is normally monitored with a Heath HO-10 Monitor Scope. When I keyed the exciter with a succession of dots at about 40 wpm it was obvious from the scope display that there was an enormous variation in output within a single dot, following an approximately sinusoidal pattern. It illustrated precisely the conditions described years ago in *GE Ham News*.²

The correction was simply to add 20 μ F of filtering capacitance (there is plenty of room in the spacious old HA-10 linear). The result was that both long and short keyed pulses were nearly flat on the monitor scope. Furthermore, there was a substantial increase in average power output. I suspect that improving the dynamic performance of a linear amplifier's power supply will be an economic way of increasing power for many amateurs.

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1. W. Wildenhein, W8YFB, "Regenerative Detectors and a Wideband Amplifier for Experimenters," *ham radio*, March, 1970, page 66.
2. "About Power Supplies: Better Dynamic Characteristics Mean Better Output on CW, AM and SSB," *GE Ham News*, January/February, 1954.*

Guy Black, W4PSJ

*Copies of this article are available from the author for 10 cents and a self-addressed stamped legal-size envelope. *editor*.



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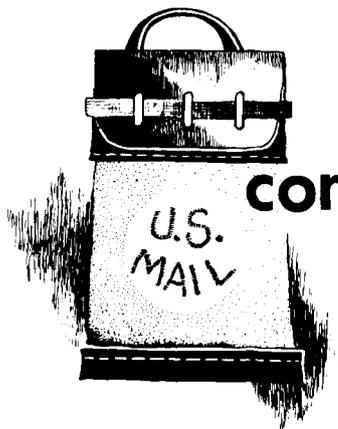
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collinear antenna

Dear HR:

I doubt that WB6KGF is getting 14-dB gain from the 4-element 144-MHz collinear described in the May, 1971 issue. A single $5/8$ -wave section gives 3-dB gain over a ground plane; theoretically, two sections provide 6-dB gain, and four sections yield 9 dB — *if current is equal in all sections*. When fed at the bottom, top sections have less current. I would guess that 7-dB gain would be a more realistic figure.

William I. Orr, W6SAI
San Carlos, California

Dear HR:

W6SAI is quite right in saying that it is impossible for the gain of a 4-element collinear to be entirely responsible for the 14-dB increase in field strength. As I said, some of this increase could be attributed to an increase in transmitter power input; I should have emphasized this more.

During the tests the collinear was at the end of a long run of old RG-8/U coax (over 100 feet) with fixed 50-ohm transmitter output. However, I don't think

gain would be as low as 7 dB — don't forget, mutual coupling between elements plus the low-loss characteristics of the copper-tubing elements would reduce any current imbalance caused by feeding the antenna at the end as opposed to the center.

The antenna system is highly resonant with circulating currents far exceeding the input rf current. Unfortunately, since the antenna described in the article is in Viet Nam, it is not available for further tests with input power held constant.

Robert A. Dahlquist, WB6KGF
Kingsburg, California

old-time radio

Dear HR:

K4NW's excellent article, "Those Were the Days," brought back memories of the good times I used to have with ham radio. I have had a transmitter of some sort on the air since May, 1913, starting out with a Ford spark coil powered by Columbia 6 dry cells, obtained from the telephone company when they installed new ones. I will never forget the thrill that came with my first contact — it was only seven miles away, but what a thrill!

My first receiving crystal was homemade. I used a teaspoon for mounting; the holder was a piece of pure lead about the size of a large pea. The crystal itself was made by melting flower of sulphur until it formed a crystal. Believe it or not,

rf interference

Dear HR:

When selecting the transistor radio as an i-f for the converter used as a noise locator (*QST*, June, 1966), it will prevent birdies if a trf instead of a superheterodyne receiver is chosen. However, be sure that the trf set has an unused broadcast frequency available on it without interference.

The *QST* converter has no connection between the oscillator and the mixer in the high-frequency range. Tests with a signal generator resulted in adding a coupling capacitor for some bands in order to obtain adequate sensitivity. Also, the whip antenna was too small. A bamboo pole, wound loosely with wire, with a long lead to a plug, provided sensitivity comparable to the home receiver. In a few cases, a wire coil used as a loop antenna, even on the fm band, could be used for determining the direction of the noise.

Bill Nelson, WA6FQG, emphasizes the need to shift to a higher frequency as the noise is approached, thus limiting its area. On the 7- and 3.5-MHz bands, the noise may be heard for a mile or two, but this can happen even on 28 MHz if the noise source is high and in line of sight. As the receiving frequency is increased when the noise is louder, there may be a substantial reduction in noise range, permitting accurate noise location. Some noises, however, may be heard in a relatively narrow frequency range. Fluorescent lights are likely to have a high-frequency cutoff around 7.5 to 8 MHz. It is helpful to have a general-coverage noise locator because of the possible frequency limitations of the noise.

On the upper high-frequency bands there may be standing waves so it is necessary to walk beyond the point of minimum noise to determine whether there is another peak further along — possibly a stronger one! Reduction of the rf gain may help particularly where the agc cannot be disabled. Beware of increased noise levels near power poles having a ground wire; this just adds antenna length

to the noise-locator receiver.

It is well to have a sledge-hammer, or at least a large rock, with which to "thump" a power pole in the noisy area. Usually, the pole "thumped" will cause the noise to break up (and sometimes to come on, or go off), but movement sometimes travels down the wires to an adjacent troublesome pole. Thick poles don't respond to this treatment, but frequently it is possible to make a guy wire swing sufficiently to produce the necessary movement of a pole to affect the radio interference.

New electric lines appear to be worse than old ones. Here, this has resulted from installation where the nuts were left loose all over a pole! Sometimes, a noisy pole again becomes noisy a half-year later. One capacitor bank on a pole near my home has created noise three different times.

One swimming-pool pump motor put out noise for a hundred feet. It was reported to the owner. Within about two weeks, it failed completely and was replaced.

The Navy purchased shield boxes for fluorescent lights; in addition to line filters and metal shields, the light passed through a conducting-glass cover. They were very effective, even in a radio receiving location underground.

Bill Conklin, K6KA
La Canada, California

vhf fm receiver

Dear HR:

In the September issue there was a 2-meter fm receiver project that really took my fancy. I built it, and it worked so well I wanted to tell you how pleased I am with the finished receiver.

Since I used printed-circuit boards I only had to worry about collecting the necessary parts, and alignment. Collecting parts was the hardest part of the whole project — I enlisted the help of everyone I know. Most of the parts were not too difficult, but the Sprague integrated circuit and i-f cans were tough. I also had

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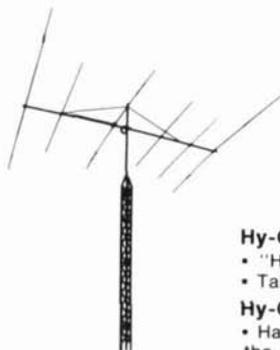
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Butler, Missouri 64730
816-679-3127

some difficulty finding shields for the handwound coils. I finally spent a dollar and ordered six inexpensive i-f cans from a dealer on the East coast; after discarding the goodies inside, the cans worked fine for coil shields.

The i-f transformers specified in the original article were too expensive for me. The author now advises the use of Calectro D1823 i-f transformers. These work fine and cost less than one dollar each. The Sprague integrated circuit was the most elusive rascal I have ever hunted, but I finally found one.

The sensitivity of the receiver really surprised me – 0.3 μ V for full quieting. If you decide to build this receiver I hope you have as much fun as I did. It is a very worthwhile project, and one that will be in constant use.

Larry Pepple, WA9RRZ
Fort Wayne, Indiana

audio filters

Dear HR:

That simple audio filter described by W4NVK in the October issue of *ham radio* really works. I put one together out of the junk box, about 0.3 μ F across the 88-mH toroid, and it entirely cuts out the QRN. Tuning was very sharp – I estimate about 100 Hz wide; so sharp, in fact, that I had to tune with the receiver pitch control. There is about 10 dB reduction in volume, which is not too bad.

I put it together with no solder, just jumpers, and it about floored me when I kicked it into the circuit. I worked a novice on 80 meters with no problem – when I turned the filter off he disappeared into the QRN.

I don't get excited very easily, but this filter really works. My SX-101A and 500-Hz filter with bridged-T notch are sick next to it. I added a 1000-ohm pot across the filter to adjust the amount of filtering – at zero resistance the filter is shunted out, with maximum resistance you get full filtering.

Duane Schnur, WB8EEJ
Caro, Michigan 48723

1296-MHz moonbounce

Dear HR:

I am using two Motorola 1N5150 varactor diodes to produce 10 watts output on 1296 MHz; this will be used to drive a pair of 3CX100A5s to 100 watts into a circularly symmetric feed on my 28-foot dish. Hopefully I will be able to work G3LTF and W2NFA via moonbounce during March.

Peter, G3LTF, has sent me a preamp which has a measured noise figure of 5.3 dB and has been used to receive his own echoes on his 15-foot dish. The preamp will be mounted at the feed point with the converter in the shack. I have some 1-5/8" Andrews heliax which will be used on the transmitting side, and as my dish is more accurately constructed than Peter's I should be able to hear him between 6 and 10 dB above the noise. My own echoes should be 2 to 3 dB weaker as he is running four 3CX100A5s and inferior cable.

I hope that some modifications to my dish will be made and the dish remounted by late February or early March. The polar mount will enable accurate tracking automatically any time the moon is above the horizon.

I also have plans for 32 18-foot long crossed *Swan* yagis mounted 15-feet apart in both planes and phased for vertical, horizontal, clockwise or counter-clockwise circular polarization. This will be polar mounted on another tower that I hope to have operational by the end of this year.

My ultimate aim is to provide moonbounce antenna capability on 144, 432 and 1296 MHz to any interested vhf group or individuals in Australia at no charge to foster international goodwill and promote moonbouncing to many who would otherwise be unable to partake of this fascinating aspect of amateur radio.

Ray Naughton, VK3ATN
Birch, Australia

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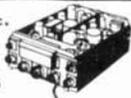
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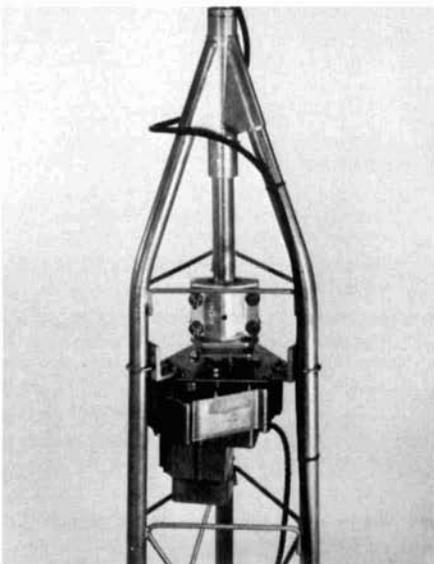
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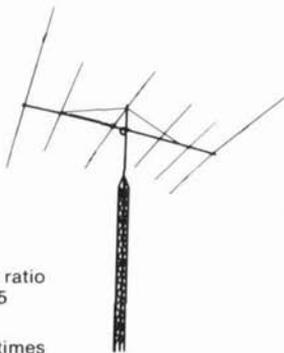
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The new Tempo/fmv from Henry Radio is setting new highs in performance and value. The solid-state two-meter transmitter features 12-watts rf output with spurious signals more than 60-dB down. Frequency deviation is adjustable from 5 to 15 kHz. The dual-conversion receiver has 0.6 μ V sensitivity for 20-dB quieting (0.3 μ V usable threshold). Selectivity at full quieting is ± 6 kHz at -6 dB and ± 15 kHz at 170 dB. Receiver audio power output is 1 watt.

The Tempo/fmv provides eight-channel coverage. Power supply requirements are 12 to 15 Vdc at 2 amps; the unit weighs 4.5 pounds. Extra features of this fm transceiver include a build-in metering test socket, an operation/maintenance manual and an optional test-set accessory. The built-in test socket can be used with a sensitive microam-

meter to monitor all stages including the discriminator. The large instruction manual covers complete checkout and alignment of the transceiver. The optional test set mates with the built-in test socket and includes a sensitive microammeter.

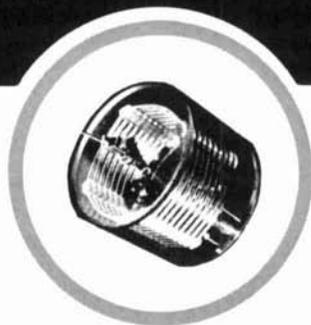
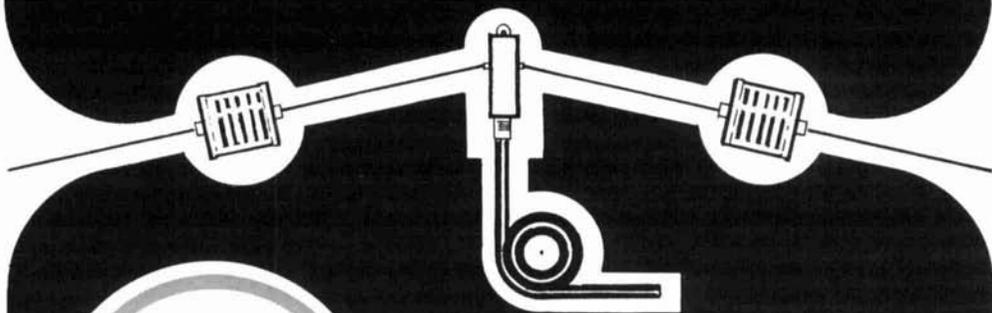
The Tempo/fmv two-meter fm transceiver is priced at \$249 including microphone from Henry Radio, 11240 West Olympic Boulevard, Los Angeles, California 90064. The optional test-set accessory is priced at \$29. For more information use *check-off* on page 126.

gem-quad antennas

The fiberglass quad antennas from Gem-Quad in Canada offer a number of interesting design features, including light weight, a cone-shaped design that maintains critical measurements under severe weather conditions, non-corrosive construction, single 52-ohm feedline on all bands, low swr, and nylon tension tubes at the corners of the quad to eliminate sharp angles in the wire, thus assuring longer antenna life. Gem-Quads come in two-, three- and four-element arrangements for use on 10, 15 and 20 meters. Forward gain for the 2-element quad is reported to be 8 dB on DX signals; with an optional third element (which may be easily installed with no conversion) gain is increased to about 8.9 dB. Front-to-back ratio of the 2-element model is 25 dB; for 3 elements, front-to-back is 30 dB.

The Gem-Quad antenna is designed for optimum performance and can easily be rotated by an ordinary tv rotator. When properly assembled, the Gem-Quad is capable of withstanding winds up to 100 mph. The 2-element quad is \$107.00 complete; 3-element quad, \$167.00; and 4-element quad, \$227.00 complete. A third or fourth element, if purchased separately, is priced at \$60.00. For more information write to Structural Glass Limited, 20 Burnett Avenue, Winnipeg 16, Manitoba, Canada, or use *check-off* on page 126.

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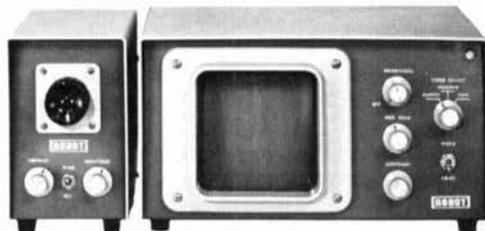
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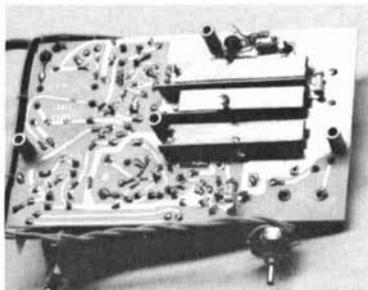
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432-MHz converter



The new Janel 432-MHz converter provides many design features for the 432-MHz operator, including noise figure of 5.5 dB, any 4-MHz band from 420 to 470 MHz, 35-dB gain (adjustable) and image rejection (with 28-MHz i-f of 40 dB; i-f rejection is 95 dB). The new high performance converter uses all silicon and mosfet transistors, high-Q air-line tuned circuits for low loss and built-in zener-regulated power supply for extreme frequency stability; the built-in power supply may be used for 117 Vac or 12 Vdc operation.

The converter circuit uses five npn bipolar silicon transistors plus one mosfet. The rf amplifier uses a 40235 in a common-emitter configuration with a broadband input circuit to tune out input reactance. Two silver-plated stripline output circuits provide maximum selectivity and image rejection. The local oscillator chain begins with a crystal in the 100-MHz region, thus reducing the number of multipliers and spurious responses.

A 40235 oscillator excites two 40235 doublers to obtain injection to the base of the 40235 grounded-emitter mixer. The output of the mixer is capacitively coupled to the three-stage mosfet i-f amplifier that provides zero-to 27-dB i-f gain by adjustment of a control on the front panel.

I-f output frequencies available off the shelf are 26-30 MHz, 28-32 MHz and 50-54 MHz. One-year guarantee. \$64.95 from Janel Laboratories, Post Office Box 112, Succasunna, New Jersey 07876. For more information use *check-off* on page 126.

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solid-state qrp projects

QRP operation is rapidly becoming one of more popular facets of amateur radio, largely because it means a return to homemade equipment and economical operation. This new book by Ed Noll, W3FQJ, covers a variety of solid-state transmitting gear, with power ratings from less than 100 milliwatts up to about 20 watts. A variety of solid-state crystal oscillators and vfos are included as well as multistage cw transmitters. There are also a-m and single-sideband circuits.

The emphasis in this book is on solid state with circuits using bipolar transistors, field-effect transistors and integrated circuits. Also included are an introduction to practical solid-state circuit theory, QRP test gear and simple antennas for QRP operation.

128 pages, softbound. Published by Howard W. Sams & Co., Inc. \$4.25 from Comtec Books, Box 592, Amherst, New Hampshire 03048.

RTTY control terminal

The new WCI phase-locked-loop detector-type RTTY demodulator converts the audio-frequency shift tones from a communications receiver to dc pulse data that operates a teleprinter. This demodulator has the unusual ability to track the input audio signal frequencies automatically if they change frequency due to transmitter or receiver drift. In addition, this unit will automatically copy any shift from 100 to 1000 Hz. Also included in the detection circuitry is an automatic threshold computer.

If the input signal is degraded past a preset level (due to a fading or a very poor signal-to-noise ratio) an automatic noise squelch circuit places the unit in *mark hold* to hold the selector-magnet armature closed to prevent the machine from printing unwanted erroneous characters.

The RCT-2D RTTY control terminal uses all solid-state circuitry with plug-in printed-circuit boards. Each printed-cir-

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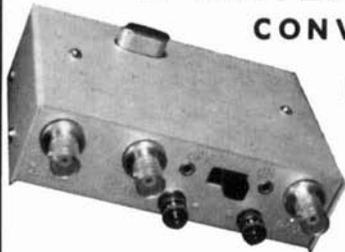
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A full description of this fantastic converter would fill this page, but you can take our word for it (or those of thousands of satisfied users) that it's the best. The reason is simple — we use three RCA dual gate MOSFETs, one bipolar, and 3 diodes in the best circuit ever. Still not convinced? Then send for our free catalog and get the full description, plus photos and even the schematic.

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cuit board has built-in top-access test points and adjustments for easy maintenance. The RCT-2D can also be supplied with digital autostart and selective call. The digital autostart keeps the teleprinter motor turned off until a message is being received. Within the first three characters the motor turns on and the message is printed. Ten seconds after the last character is received, the motor is turned off again.

The RCT-2D RTTY control terminal is available from WCI, Post Office Box 17, Schaumburg, Illinois 60172. For those amateurs who want to build the WCI phase-locked demodulator themselves, the company is offering a kit of pre-drilled circuit boards with instructions for \$45. For more information use *check-off* on page 126.

high-frequency receiver



The new Galaxy FFR-230/6 receiver is completely solid state and crystal controlled for operation on any six channels in the 2- to 18-MHz range. Front panel controls are provided for channel selection, mode—usb and lsb (2.1 kHz bandwidth) and cw (1.5 kHz bandwidth), clarifier (ssb), bfo, af and rf Gain. Features include rack mounting, internal speaker and 117-Vac power supply, external muting and audio (4 ohms and 600 ohms), modular construction and plug-in circuit boards for changing frequency. Priced at \$950.00. Optional filter bandwidths are available.

For more information on the new Galaxy FFR-230/6 receiver, use *check-off* on page 126 or write to Galaxy Electronics, Subsidiary of Hy-Gain Electronics Corporation, Route 3, Lincoln, Nebraska 68505.

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RF550A contains high accuracy watt meter; calibrated in 400 and 4,000 watt scales; switch for forward or selected power; switch to select 5 antennas or dummy load. Order No. 857 Ham Net \$75.00

RV550A is a solid state VFO. Function switch selects the remote unit to control Receive-Transceive-Transmit frequency independently. Order No. 856 Ham Net \$95.00

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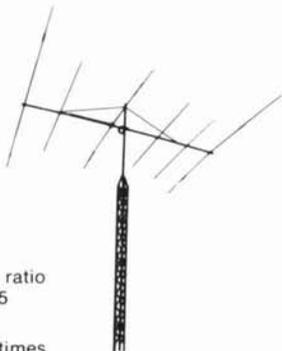
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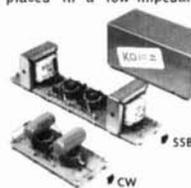
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The CW filter has a spot frequency of 780 Hz and a passband of 1100 Hz with a reference level, 40 decibels below the signal level at the design frequency. The peak of the passband is 100 Hz wide at the -3 decibel reference points. The CW filter is specifically designed for low-impedance input and high-impedance output. High-impedance crystal earphones are recommended. However, with low impedance earphones a small auxiliary amplifier or impedance matching transformer may be used.

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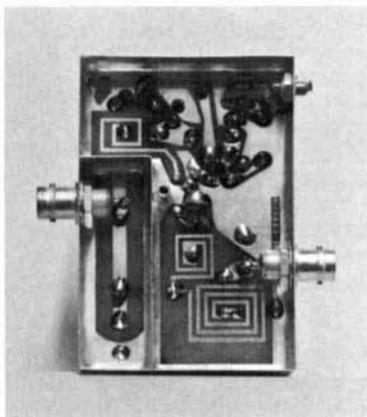
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144/432-MHz transverter

The new Braun TTV-1270 solid-state transverter seems like the ideal way to become operational on 432 MHz. The TTV-1270 was developed specifically for mobile, portable and field-day operation, although it is also suitable for low-power fixed-station operation on 432 MHz.

To put the TTV-1270 on the air simply connect it between a 2-meter a-m, fm or cw transceiver and a 432-MHz antenna. No antenna switching is required; a 12-volt dc power supply is needed only for receive. Input frequency range is 144 to 146 MHz; output range is 432 to 438 MHz.



The transmit section of the TTV-1270 transverter uses a factory-aligned varactor tripler. Input power should be from 50 mW to 1.2 watts. Output is 30 to 720 mW at 60% efficiency. In the receive mode the output of a 96-MHz voltage-regulated crystal-controlled oscillator is tripled to 288 MHz. This signal serves as the injection for a 432-MHz diode mixer stage. Noise figure on receive is 9 dB; receiver gain is -9 dB. To change from transmit to receive, the 12-V power supply is connected; current drain in the receive mode is 12 mA.

The Braun TTV-1270 is built into a silver-plated enclosure with rf-tight partitions. The stripline 432-MHz circuits are

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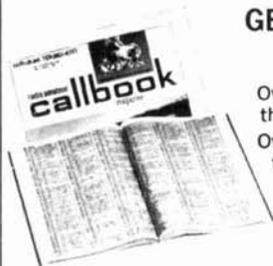
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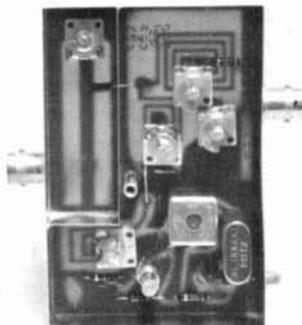
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be obtained by retuning the unit with your two-meter source and 432-MHz antenna; complete instructions are provided with the unit.

The Braun TTV-1270 is specially introductory priced at \$37.50 from Spectrum International, Post Office Box 87, Topsfield, Massachusetts 01983. For more information on Braun vhf and uhf equipment, use *check-off* on page 126.

transistor substitution handbook

Although bipolar transistors are noted for their low failure rate occasionally they do have to be replaced. As long as the specific type number required is available replacement of the transistor is no problem because a duplicate of the original should be used whenever possible. All too often, however, an exact replacement cannot be obtained without considerable delay. Furthermore, the great variety of transistor types make it difficult to determine which transistor can be substituted for the original.

The new 11th edition of "Transistor Substitution Handbook" is possible be-

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- SWR less than 1.5:1 on all bands
- Takes maximum legal power
- 24-foot boom. Order No. 389 Ham Net \$179.95

Hy-Gain's 18 AVT/WB

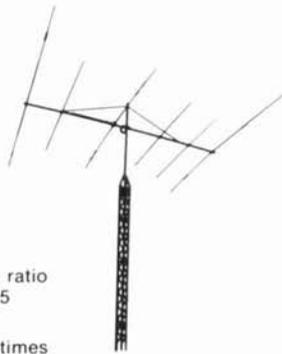
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Hy-Gain's Thunderbird TH3Mk3 (not shown)

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cause of the ability of modern-day electronic computers to handle a large quantity of information in a relatively short length of time. The computer selected the substitutes listed in this handbook in much the same manner that an individual would select a transistor replacement. The electrical and physical parameters shown in the manufacturer's published specifications for each bipolar transistor were given to the computer, and then each transistor was compared with all the others. Over one billion data comparisons were made in the preparation of this book.

The transistors which matched within given limits are listed as substitutions. A second section contains additional information on general purpose replacement transistors: the manufacturer, the polarity (npn or pnp), the material (germanium or silicon), and the recommended applications. The information in this handbook can be used by anyone concerned with transistor replacement — be it in amateur, industrial, commercial, or home-entertainment equipment. 160 pages, soft-bound. \$2.25 from Comtec Book Division, Box 592, Amherst, New Hampshire 03031.

marine radiotelephone



For shipboard and shore applications RF Communications has produced a marine single-sideband radiotelephone incorporating the latest electronic advances and compatible with all existing and future marine stations. The new RF-201M series includes compatible a-m, speech clarifier, squelch, balanced audio and complementary accessories including remote controls, 1-kw amplifier, antenna couplers and power supplies. The unit is

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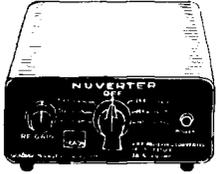
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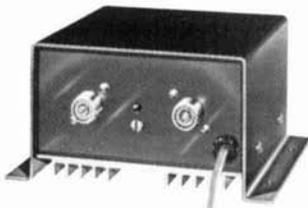


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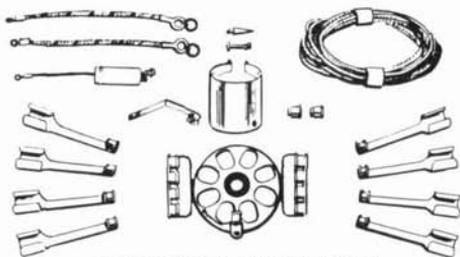
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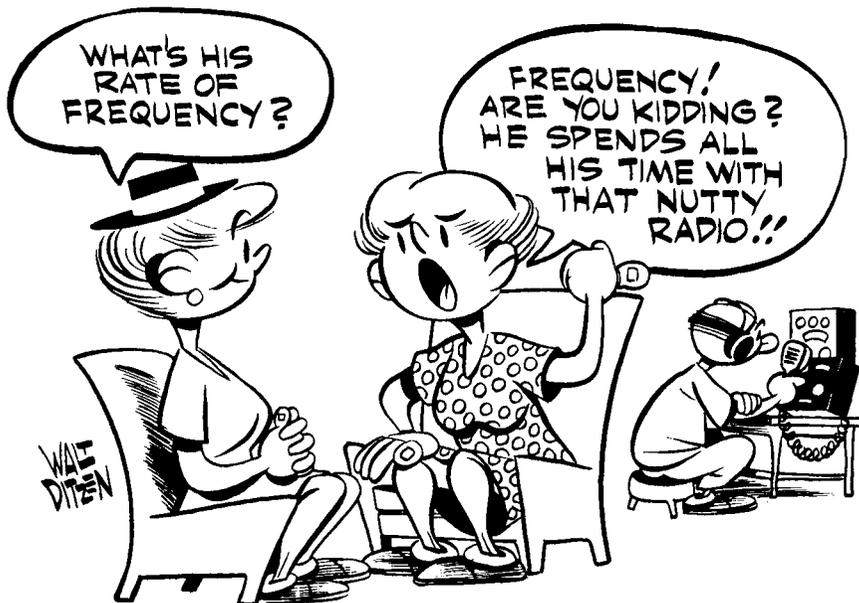
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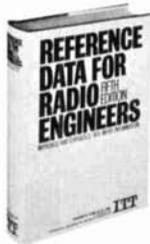
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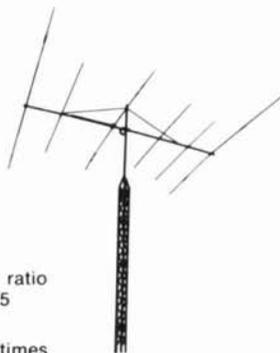
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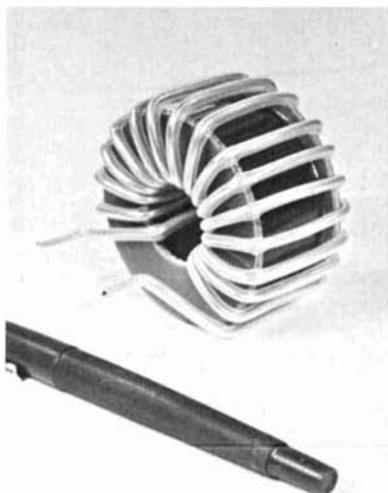
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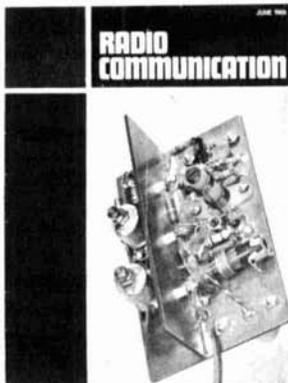
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Many thousands of you have become very familiar with the various Radio Society of Great Britain books and handbooks, but very few of you are familiar with their excellent magazine, **Radio Communication**.

This is the oldest and most widely read British amateur radio magazine. Published monthly it provides complete coverage including such popular features as: Technical Topics, a monthly survey of the latest ideas and circuits, Four Meters and Down, a rundown of the latest in VHF and UHF and much more.

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Design **A** is capable of handling 2 kW PEP; however, it should not be used if 1 kW continuous key-down operation is contemplated. Design **B** is electrically similar to Design **A**. However, it has an outboard 20-meter coil to reduce excessive core heating. This configuration will handle key-down inputs of 1 kW for prolonged periods.

Design **C** is capable of handling the greatest power (3 kW PEP) and is the most efficient design. An outboard coil is used for 10, 15 and 20 meters. The toroid is used on only 80 or 40.

All circuits are designed to be used with tubes which operate at voltages between 3,000 and 1,800 Vdc at peak current up to 1 ampere. For higher voltages and lower currents tap positions will require adjustment.

The no. 115 2-kW toroid kit contains a special 2-inch toroid core, end spacers, Teflon sleeve and wire plus complete instructions; \$16.95 from Redline Electronics, 3498 East Fulton Street, Columbus, Ohio 43227. Also available are a bifilar filament choke with 30-amp rating, catalog no. 421, priced at \$7.95; and 1.5-amp plate rf choke (2.5 to 55 MHz), catalog no. 417, priced at \$6.95.

base-station control extension



The new XR-4 remote extension from Alpha Electronics is designed to enhance the efficiency and convenience of your

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RF550A contains high accuracy watt meter; calibrated in 400 and 4,000 watt scales; switch for forward or selected power; switch to select 5 antennas or dummy load. Order No. 857 Ham Net \$75.00

RV550A is a solid state VFO. Function switch selects the remote unit to control Receive-Transceive-Transmit frequency independently. Order No. 856 Ham Net \$95.00

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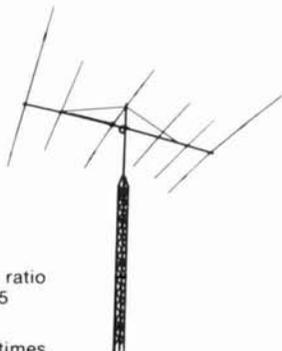
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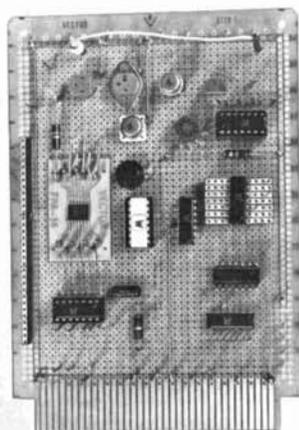
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repeater system. The unit features center button push-to-talk handset, a choice of colors, automatic monitor of the tone function, transmit indicator light and volume control. In addition, the XR-4 control extension is available with several options, including intercom system, multifrequency switching, handset volume control and choice of microphone types.

Of special interest is the provision that makes it possible to use the unit (with an add-on module) as a fully amplified remote on a telephone line. The XR-4 may be easily connected into any base station or dc remote through an 8-conductor cable. Cable length is limited only by line loss and receiver or remote amplifier output power; distances greater than 300 feet are often achievable.

The XR-4 base-station control extension is priced at \$89.50 from Alpha Electronic Services, Inc., 8431 Monro Avenue, Stanton, California 90680. For more information use *check-off* on page 126.

high-density plugboards



Vector Electronic Company has just announced a new high-density Plugboard series for mounting DIPs and discrete components. The Plugboards are made from glass epoxy or phenolic and are punched with an overall grid of .042-inch diameter holes located on 0.1-inch

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centers. The boards will accommodate any component with lead spacing on 0.1-inch multiples. As a result, integrated circuit packages or sockets may be mounted in any required density. Discrete components or hardware items such as test jacks and clips may be located anywhere on the board without fear of interfering with any pre-etched circuitry.

The new boards are offered with a choice of etched nickle and gold-plated contacts, Elco Varicons or Vector edge pins. Along with the Plugboards Vector is also supplying mating receptacles with a choice of solder-eye termination or square posts for wrapped wire connections.

The company also makes a full line of other accessories for the cards including rack-mounted cages and module cases for housing the boards, IC sockets, and various types of terminals which will fit the boards. Pricing of the new Plugboards ranges from \$6.00 to \$8.00, depending upon the type of board. For more information, use *check-off* on page 126, or write to the Vector Electronic Company, 12460 Gladstone Avenue, Sylmar, California 91342.

crystal cw filter

The latest addition to the crystal-filter line offered by Spectrum International is the new KVG XL10M, a ten-pole 500-Hz CW filter with superb skirt selectivity. Near-Gaussian response to -6 dB eliminates ringing; built-in matching transformers eliminate the need for external inductors.

The center frequency of the new crystal filter is 9.0 MHz; bandwidth at -6 dB is 500 Hz. Shape factor (6:60 dB) is 2. Insertion loss of the XL10M is 10 dB maximum; ultimate rejection is 80 dB minimum. The XL10M is priced at \$59.95 from Spectrum International, Post Office Box 87, Topsfield, Massachusetts 01983. For more information on the XL10M as well as other crystal filters and crystal discriminators, use *check-off* on page 126.



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RF550A contains high accuracy watt meter; calibrated in 400 and 4,000 watt scales; switch for forward or selected power; switch to select 5 antennas or dummy load. Order No. 857 Ham Net \$75.00

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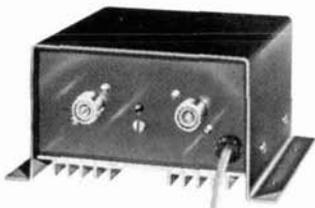


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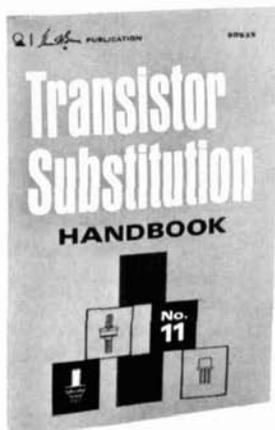
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The new *QSK MK II* is a station control system that permits you to hear other stations between your transmitted dots and dashes at speeds up to 60 wpm. The unit can be used with virtually any exciter/power-amplifier and receiver combination to provide high-speed antenna switching, exciter keying and receiver isolation. The *QSK MK II* will operate on all bands with no adjustment; the unit has been proven in traffic, contest and DX work.

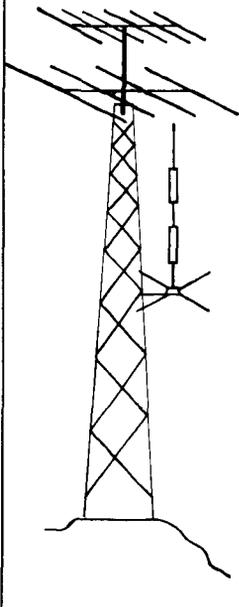
The 8-transistor, 9-diode timing and sequencing circuit assures cold switching of the vacuum relay (5-kW rating). The unit features negligible losses and includes a built-in 117-Vac power supply. \$119.95 from your dealer, or write to Dynamic Technology International, Inc., 8 Fellowship Road, Cherry Hill, New Jersey 08034. For more information use *check-off* on page 126.

communications receiver



The new Drake model DSR-1 receiver is a commercial grade communications receiver employing the most up-to-date solid-state devices and circuitry offering continuous coverage from 10 kHz to 30 MHz. The received frequency is indicated by six Nixie tubes to 100 Hz. All frequency injections are controlled by a phase-locked digital synthesizer which allows incremental frequency selection in 10, 1 and 0.1 MHz steps. The remaining zero to 0.1 MHz is continuously adjustable from the front panel by a highly stable variable oscillator. Modular construction on easily accessible printed-circuit boards is used throughout.

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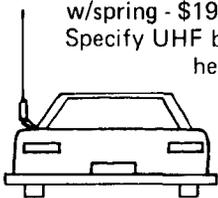
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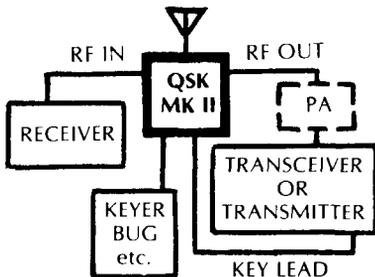
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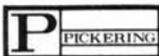
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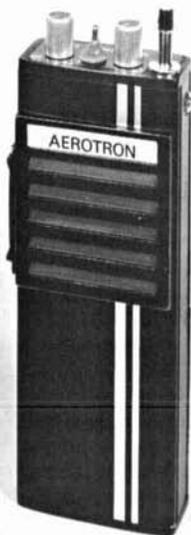
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portable vhf-fm radio



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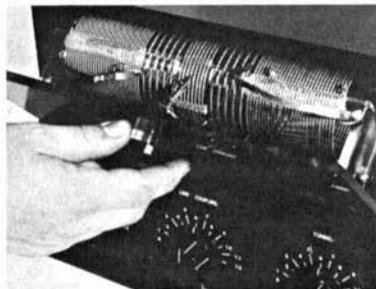
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HOT CARRIER DIODES		HP2800	801	12-1110-02	Matched by HAL	6-14-79
GENERIC		1N4726(3) Rev.	1N4726(3) 1st	1N4726(3) 2nd	1N4726(3) 2nd	1N4726(3) 1st
LINEAR IC'S		709M	756	709M, 710M	11 25	741N
DIGITAL IC'S		MC14292	33 75	MC1488C	33 75	MC13800
MIXER		MC1790P	11 30	MC1790P	11 30	MC1790P
DIP TTL		7400	7401	7402	7410	7430
7404		7405	7406	7407	7410	7412
7412		7413	7414	7415	7416	7417
7406		31 15	7480	7482	7483	42 10
7484		11 10	7490	7492	7493	10 10
7494		11 10	7498	7499	7500	7501
7502		11 10	7504	7505	7506	7507
7508		11 10	7509	7510	7511	7512
7513		11 10	7514	7515	7516	7517
7518		11 10	7519	7520	7521	7522
7523		11 10	7524	7525	7526	7527
7528		11 10	7529	7530	7531	7532
7533		11 10	7534	7535	7536	7537
7538		11 10	7539	7540	7541	7542
7543		11 10	7544	7545	7546	7547
7548		11 10	7549	7550	7551	7552
7553		11 10	7554	7555	7556	7557
7558		11 10	7559	7560	7561	7562
7563		11 10	7564	7565	7566	7567
7568		11 10	7569	7570	7571	7572
7573		11 10	7574	7575	7576	7577
7578		11 10	7579	7580	7581	7582
7583		11 10	7584	7585	7586	7587
7588		11 10	7589	7590	7591	7592
7593		11 10	7594	7595	7596	7597
7598		11 10	7599	7600	7601	7602
7603		11 10	7604	7605	7606	7607
7608		11 10	7609	7610	7611	7612
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7633		11 10	7634	7635	7636	7637
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7643		11 10	7644	7645	7646	7647
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7663		11 10	7664	7665	7666	7667
7668		11 10	7669	7670	7671	7672
7673		11 10	7674	7675	7676	7677
7678		11 10	7679	7680	7681	7682
7683		11 10	7684	7685	7686	7687
7688		11 10	7689	7690	7691	7692
7693		11 10	7694	7695	7696	7697
7698		11 10	7699	7700	7701	7702
7703		11 10	7704	7705	7706	7707
7708		11 10	7709	7710	7711	7712
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7738		11 10	7739	7740	7741	7742
7743		11 10	7744	7745	7746	7747
7748		11 10	7749	7750	7751	7752
7753		11 10	7754	7755	7756	7757
7758		11 10	7759	7760	7761	7762
7763		11 10	7764	7765	7766	7767
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TELETYPE #28 LRB4 reperfurator-transmitter "as is" \$100; checked out \$175. Includes two 3-speed gearshifts. Alltronics-Howard Co., Box 19, Boston, Mass. 02101. 617-742-0048.

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QSL'S — BROWNIE W3CJI — 3111B Lehigh, Allentown, Pa. 18103. Samples 10¢. Cut catalogue 25¢.

WORLD QSL BUREAU — see ad page 123.

RADIO SOCIETY OF ONTARIO CONVENTION 1971. Hosted by Guelph Amateur Radio Club. October 22-23, 1971. Holiday Inn — Hespeler, Ontario. Jcts. Hwys. 401 and 24. 46 miles from Toronto, 174 miles from Windsor. For further details contact D. Gore, VE3DGA or R. Jenning VE3CZE, Box 342, Guelph, Ontario.

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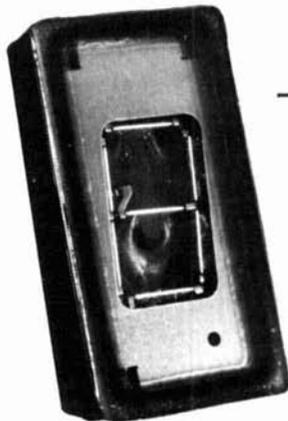
SALE: SIGNAL GENERATOR GR804C, 8-330 MHz, 5 bands plus spare, microvolt attenuator, CW, AM \$125. Frequency counter, Berkeley 554S, 100 Kc, 5 column neon readout \$100. Perf-board pre-scaler to extend range to 10 MHz \$25. Mico pantograph engraving machine with chest of 7 fonts \$150. Vanguard Labs, 196-23 Jamaica Avenue, Hollis, N. Y. 11423.

MECHANICAL FILTERS: 455 kHz. 2.1 kHz \$18.95. 300 Hz \$22.95. J. A. Fredericks, 314 South 13th Avenue, Yakima, Washington 98902.

VIRGIN ISLANDS TRANSMITTING AND LISTENING SOCIETY (VITALS) will conduct a mindxpedition during the period 9-17 October to PJ8/FS7/VP2 working/MM as well on 20-15-10 CW/SSB. QSL sase to Command Communications, Box 3374, St. Thomas, VI 00801. Principals are: KV4BV, KV4EY, KV4FR and other possibles.

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TV & RADIO TUBES 36¢. FREE catalog. Cornell, 4219 N. University, San Diego, California 92105.

SAROC Seventh Anniversary January 6-9, 1972. Advance Registration \$9.00 per person entitles registrant to SAROC Special room rate \$12.00 per night plus room tax, single or double occupancy, effective January 4 thru 12, 1972; tickets for admission to technical seminars, **Ham Radio Magazine** and **SAROC Happy Hour Thursday, Swan Electronics and SAROC Social Hour Friday, Hy-Gain/Galaxy Electronics and SAROC Champagne Party Saturday, Buffet Hunt Breakfast, Sunday.** Ladies who register will receive transportation for shopping tour, luncheon and Crazy Hat program at the New Union Plaza Hotel downtown Las Vegas, Saturday. Advance Registration, with Flamingo Hotel mid-night show, two drinks, \$14.50. Advance Registration, with Flamingo Hotel Dinner Show (entrees Brisket of Beef or Turkey) no drinks, \$17.50. Tax and Gratuity included except for room. Frontier Airlines SAROC group flight package planned from Chicago, St. Louis, Omaha, Denver, send for details. Fifth National FM Conference, ARRL, WCARS-7255, WPSS-3952, MARS, meetings and technical sessions scheduled. Accommodations request to Flamingo Hotel, Las Vegas, Nevada before December 15th. Advance Registration to SAROC, Southern Nevada ARC, Inc., Box 73, Boulder City, Nevada 89005, before December 31st.

JAMAICA (6Y5) Reciprocal licensing authorizations are now available to W/K amateurs. Write Ham Radio Magazine for application.

TAMPA HAMFEST — Hillsborough Amateur Radio Society, Inc. announces its Annual Tampa Hamfest to be a gala 2 day affair Saturday and Sunday October 16 and 17 at the Electrical Building of the Tampa Fair Grounds. Plenty of free parking. Awards galore.

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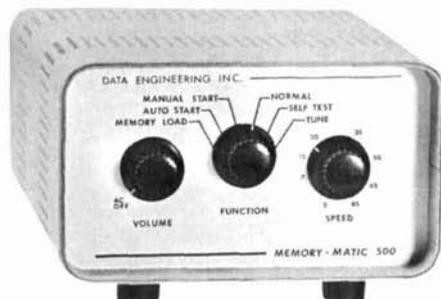
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NEW HAM MAGAZINE: Interested in public service, humanitarian actions and international friendship? Sample issue free. Worldradio, 2509 Donner Way, Sacramento, Calif. 95818, WB6AUH.

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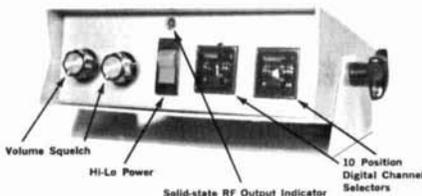
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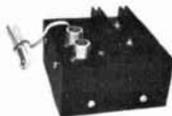


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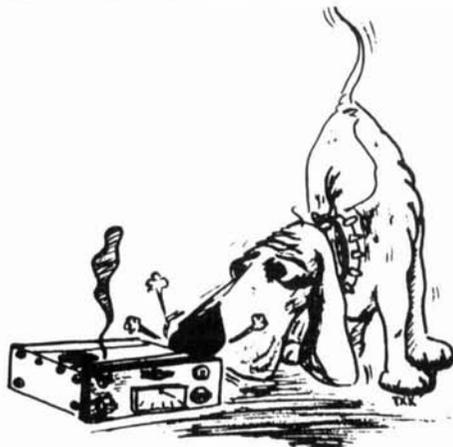
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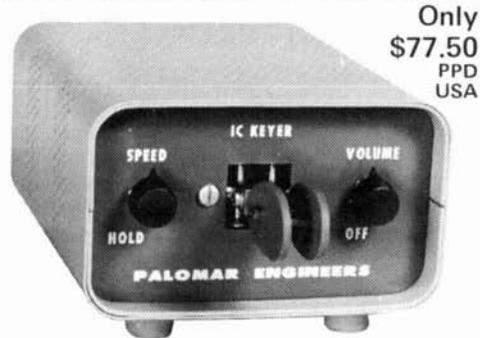
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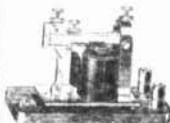
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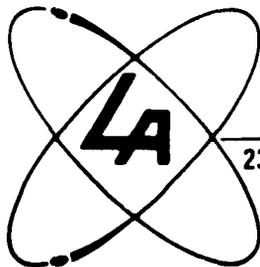
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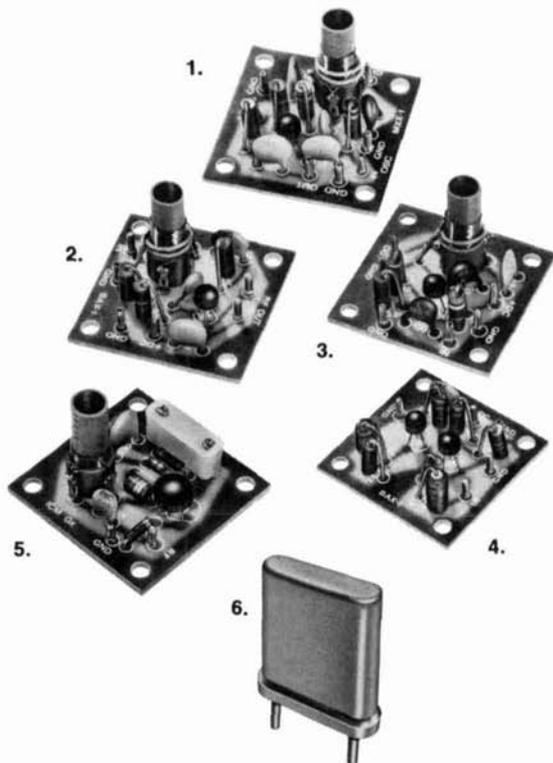
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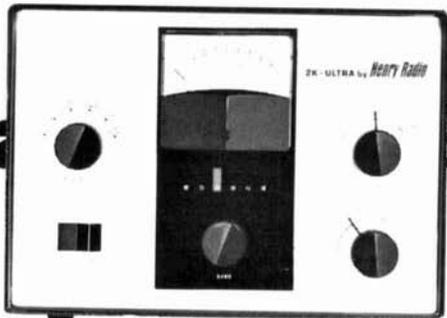


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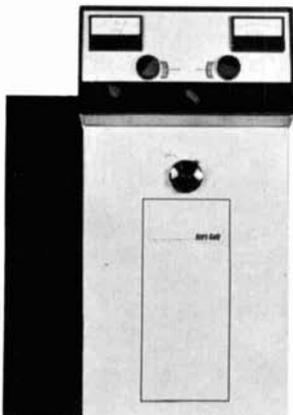
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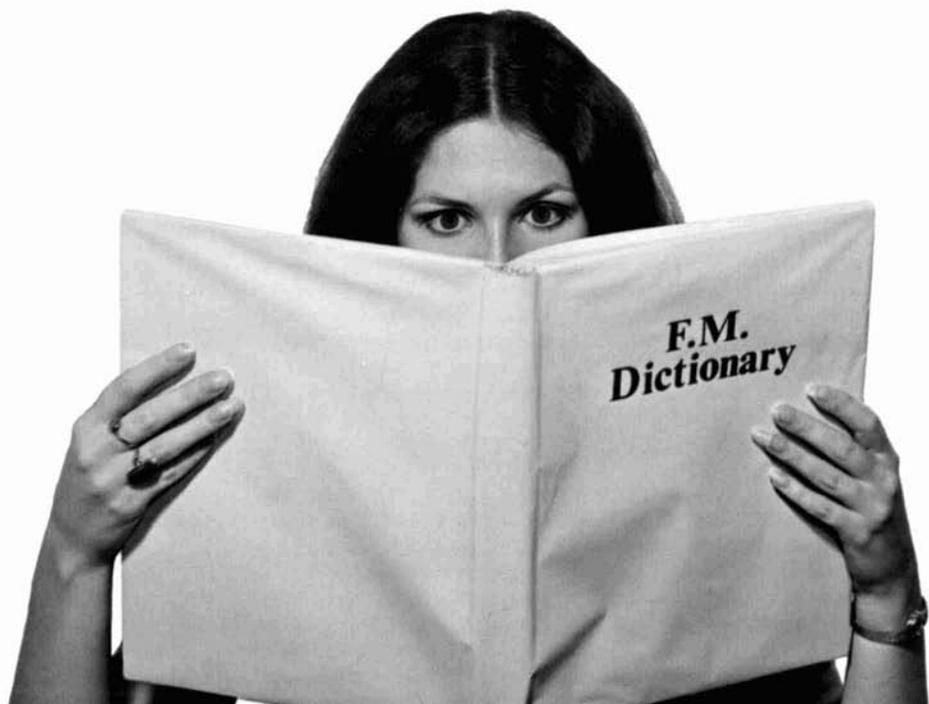
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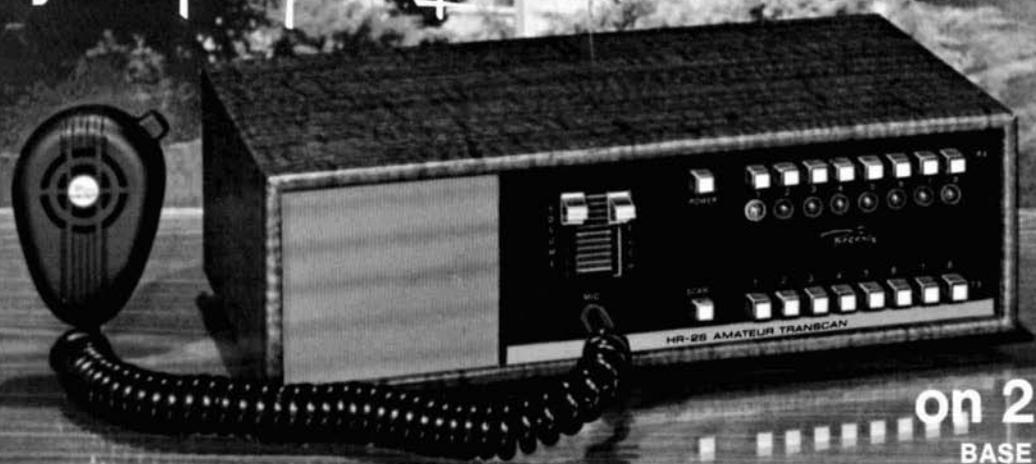

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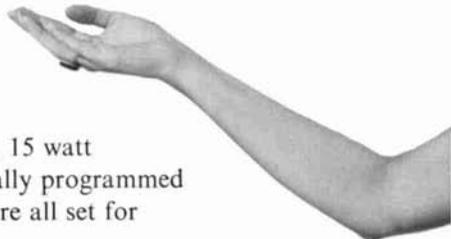
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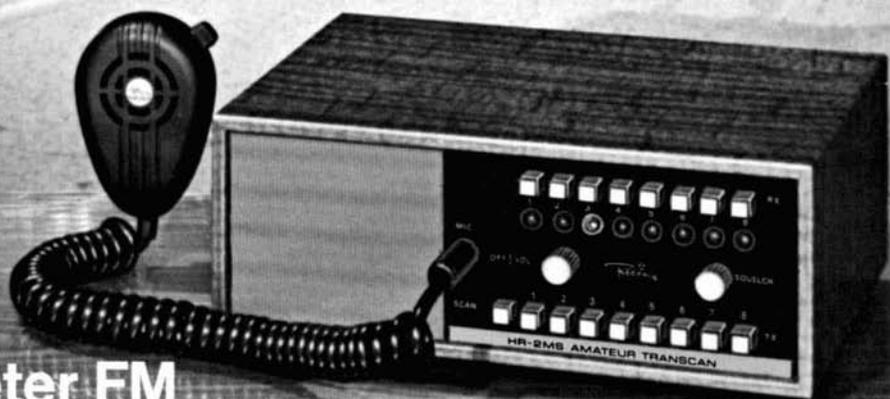


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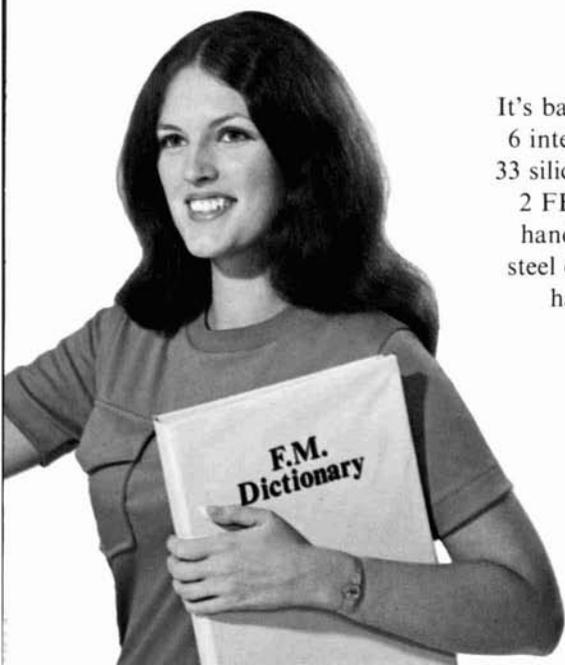
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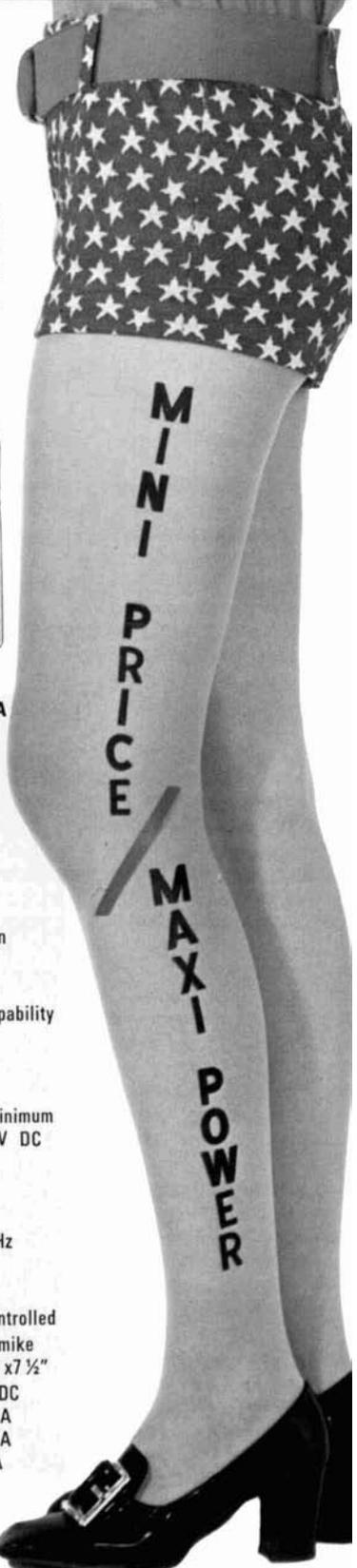
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