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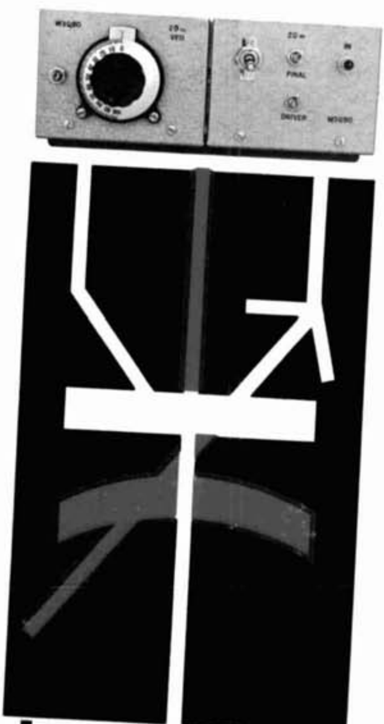

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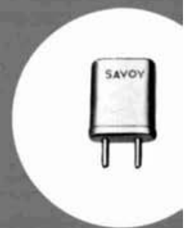

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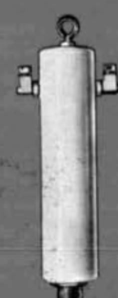

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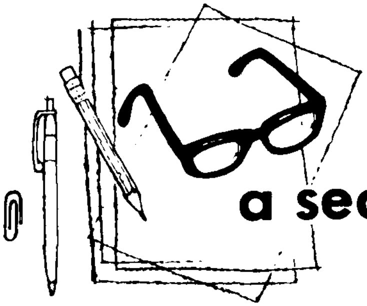

a second look

by jim fisk

Because of an excessive number of problems being encountered by the FCC with amateur repeater license applications, and the lengthy processing delay, the Commission has extended the deadline date for all amateur stations licensed prior to October 17, 1972, which were automatically retransmitting radio signals from other amateur stations, *and* for which a timely and sufficient application has been filed. An application will be considered as being timely if it was received by the FCC on or before August 30, 1973.

According to the Commission, there apparently has been some confusion among amateurs as to the actual effective date of the rules adopted in Docket 18803. The FCC has reiterated that the rules became effective on October 17, 1972, and further, that full compliance was expected as soon as possible but not later than June 30, 1973. At the request of the American Radio Relay League this period was extended to August 30, 1973.

The FCC adheres to the view that all licensees have had adequate time in which to modify their repeater stations to fully comply with the rules set forth by Docket 18803, although there may not have been sufficient time to obtain the licensing authority for a repeater station, control station and/or auxiliary link station. Therefore, although the licensing deadline has been extended (apparently indefinitely), amateurs operating repeaters under previous authorization are cautioned that their operations must otherwise fully comply with the rules. The Commission has also pointed out that licensees and control operators of stations not operated in compliance with the rules of Docket 18803 are subject to appropriate enforcement action.

The FCC has complained that the main problems contributing to the lengthy processing delays are lack of standardization, failure to supply the required information and failure to present the information in a manner permitting rapid processing. In as much as there has been considerable confusion as to what information was actually required and the Commission was apparently reluctant to supply any guidelines, this is not surprising. After a considerable amount of arm twisting a few suggestions were forthcoming but they were nebulous at best. To add to the muddle, FCC staff members reached the epitomy of bureaucratic vacillation by approving one application and then disapproving another which used exactly the same format. If the problem wasn't so serious it would be amusing.

The Commission is now, finally, developing suggested application forms that are designed to eliminate the most frequently encountered errors. This should have originally been done at the time they adopted Docket 18803 — it would have saved everyone a lot of grief. Although it has not yet been decided if these suggested forms will be adopted as official FCC forms, properly prepared license applications using them will be acceptable for processing. You can also help by using universally accepted terms and standard symbols in your repeater license applications. If there is any question about an abbreviation, for example, spell it out — terms commonly used in one part of the country may mean something completely different somewhere else.

In the meantime, if your repeater was licensed prior to October 17, 1972, *and* you have filed a new license application prior to August 30, 1973, you can continue to operate your repeater under its present license *providing* it complies with all the rules of Part 97 of the regulations, as amended by Docket 18803.

Jim Fisk, W1DTY
editor

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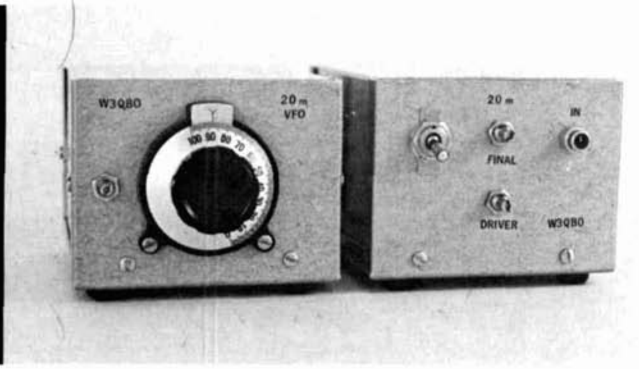


low-power solid-state vfo transmitter for 20 meters

Complete construction details for a two-watt QRP transmitter and Vackar vfo for CW operation on 20 meters

C. Edward Galbreath, W3QBO, 8326 Still Spring Court, Bethesda, Maryland

This article describes the construction of *a 20-meter solid-state transmitter using a* jfet Vackar vfo with a power output of two watts. This QRP transmitter has been used successfully for several months in many contacts, ranging geographically from New Zealand to the USSR. This has been done in a relatively few hours of operating time and — surprisingly — without the benefit of a beam; I use a 130-foot dipole, center-fed with open-wire line.

This past winter my kilowatt friends, skeptical as usual, challenged me to get on the air and try to work DX in a DX contest with this rig. I accepted their challenge, and despite lack of contest experience, I gave it a try for about two hours one afternoon during a recent DX contest. To the astonishment of all concerned I worked VP9HC, OK1TA, HB9KC, I5CFY, HA5KBM and I3ASE. Such accomplishments are only mentioned to show what can be done with two watts — I hope that more amateurs will give QRP a try.

The transmitter is built as two units, each in a 3x4x5-inch Minibox. The vfo, with the three buffer stages, is housed in a separate box only to meet my own needs. The entire transmitter can easily be built in a single Minibox of appropriate size if you prefer.

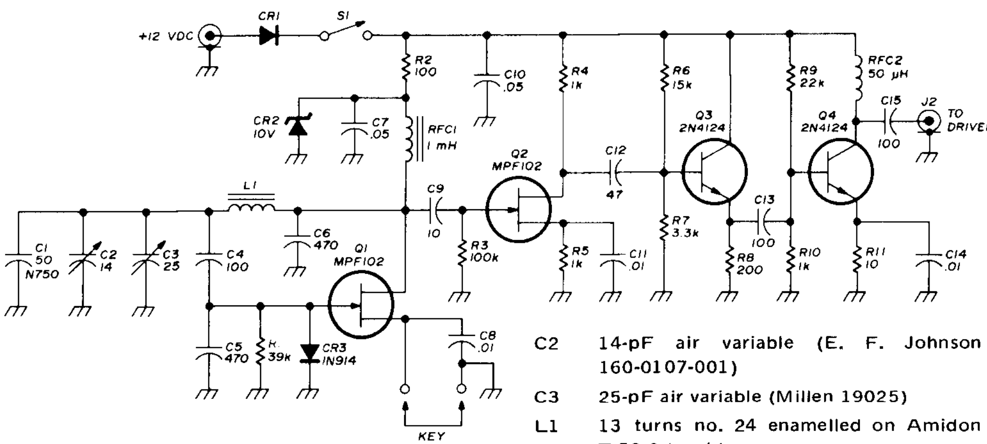
vfo

The schematic of the vfo is given in **fig. 1**. Except for the values of the oscillator's frequency-determining components and the addition of a buffer-amplifier stage, the circuit is the same as that used in my 80- and 40-meter versions.[1] The two buffer stages and the buffer-amplifier provide excellent isolation of the oscillator from the driver and final. There is only a slight frequency pull by the final amplifier even though the transmitter operates straight through. Frequency pull is so slight that there is no need to compensate for it when zero-beating a signal with the driver and final turned off.

The feedback inductor, L1, consists of 13 turns of number-24 enamelled wire, wound tight and evenly spaced on an Amidon T-50-2 toroid core. To assure short leads, the inductor is mounted on a perforated board on a standoff insulator between the tuning capacitor, C3, and the front panel of the Minibox. The leads are connected to soldering lugs. The negative temperature coefficient ceramic capacitor, C1, is mounted on variable capacitor C3.

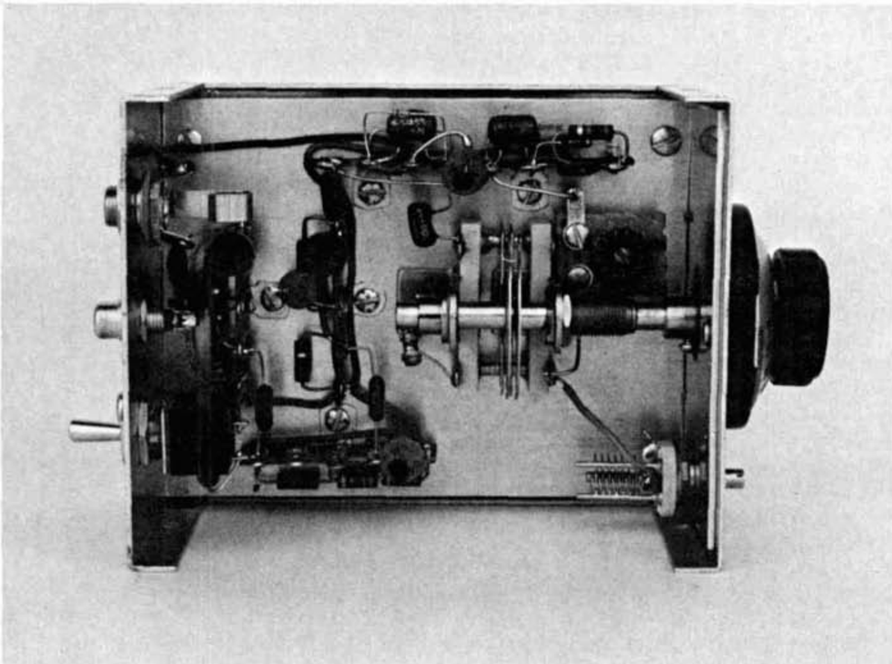
A clamping diode connected between the gate of Q1 and chassis ground has been added to the oscillator circuit. Wes Hayward, W7ZOI, when observing the gate voltage of an MPF102 on a high-frequency oscilloscope, found that, without the diode, the gate potential increased to the power supply potential on positive peaks.[2] The diode limits positive excursions to +0.7 volt, thus preventing conduction in the gate of the jfet and possible instability of the oscillator stage. Any small silicon diode such as a 1N914 will do the job.

All components of the oscillator stage except the inductor, variable capacitors and the negative temperature coefficient ceramic capacitor, C1, are mounted on two terminal strips: a 5-tie-point strip and a 3-tie-point strip. The center tie point is grounded in each case. The two terminal strips are mounted in line as a single 8-tie-point unit. This represents a constructional change from the earlier 80- and 40-meter versions and permits easy assembly of the components on the two strips as a unit before attaching them to the chassis. One simple way to do this is to bolt the terminal strips to a narrow



C1	50-pF, N750 temperature coefficient ceramic
C2	14-pF air variable (E. F. Johnson 160-0107-001)
C3	25-pF air variable (Millen 19025)
L1	13 turns no. 24 enamelled on Amidon T-50-2 toroid core
RFC1	1-mH rf choke (Millen J300-1000)
RFC2	50-µH rf choke (Millen 34300-50)

fig. 1. Circuit for the solid-state 20-meter vfo which tunes from 14.0 to 14.2 MHz is based on the very stable Vackar design. Protective diode CR1 is any silicon rectifier.



Layout of the vfo and buffer stages. Inductor L1 is mounted on the perf board just above the main tuning shaft. Transistors and other components are mounted on terminal strips. Key jack, on-off switch and output connector are mounted on rear panel, to the left.

piece of aluminum just as they would be mounted on the chassis. This unit can then be held in a vice while you attach the components.

To strengthen the Minibox, an aluminum chassis is built into it. The ends of the chassis are cut from half inch right angle aluminum stock available in most hardware stores. The aluminum angle strips are attached to the front and back panels at the base of the box and an aluminum plate is bolted to them. This chassis can be removed from the Minibox after fitting for easy assembly of the terminal strips, the tuning capacitor and the inductor.

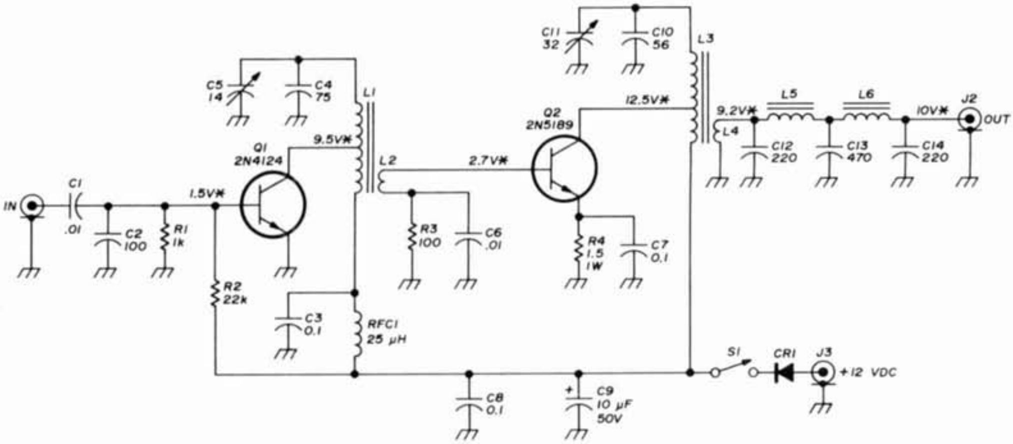
A piece of aluminum is used to reinforce the front panel. It is held in position in the box against the front panel by the machine screws that fasten the chassis in place and by the dial and the trimmer capacitor, C2. Be sure that this piece, as well as the chassis, are cut just small enough to permit the cover to slip on the box. The extra reinforcement may be omitted, however, if the additional strength and rigidity are not desired.

A two-inch vernier dial is used to adjust the main tuning capacitor. It is a common item found in many radio stores but can also be ordered by mail. Holes for the dial can be marked for drilling only after the chassis has been fitted and the main tuning capacitor is in place. A key jack, two phono jacks and an on-off switch are mounted on the rear panel. One phono jack is for battery or power supply leads, the other for rf output.

driver and final

A Motorola 2N4124 transistor is used to drive an RCA 2N5189 in the final amplifier to 2 watts output (see **fig. 2**). Both transistors are inexpensive, priced in the 70 to 80 cent range. The 2N5189 is a very efficient transistor but it seems to have been overlooked by amateurs.

The 2N4124 driver operates as a class-B amplifier with its base biased to approximately collector-current cutoff. With no signal applied to the base, col-



C5	14-pF air variable (E. F. Johnson 160-0107-001)
C11	32-pF air variable (E. F. Johnson 160-0130-001)
L1	16 turns no. 24 enamelled on Amidon T-50-2 toroid core, tapped 6 turns from B+ end
L2	2 turns small insulated wire wound over B+ end of L1
L3	16 turns no. 20 enamelled on Amidon T-68-2 toroid core, tapped 3 turns from B+ end
L4	3 turns small insulated wire wound over B+ end of L3
L5	11 turns no. 20 enamelled on Amidon
L6	T-50-2 core
RFC1	25-μH rf choke (Millen J300-25)

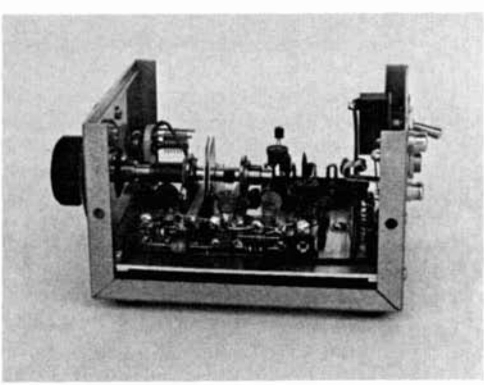
fig. 2. Solid-state driver and final provides 2 watts output on twenty meters. The voltages marked with an asterisk are the rms rf voltages at those points. Protective diode CR1 is any silicon rectifier.

lector current is near zero, thus minimizing current drain during key-up condition. As a class-B amplifier, the stage is easier to drive than if operated class-C. A capacitor divider consisting of C1 and C2 provides the correct impedance match between the vfo and the base of the 2N4124.

A fixed-tuned tank circuit is used in the driver stage. Once the tank is correctly adjusted for maximum output in the center of the CW portion of the band, no further adjustment is required. The primary, L1, of the tank coil is tapped 6 turns from the B+ end to assure a proper low-impedance match for the collector of the 2N4124. Two capacitors are used to tune L1, a 75-pF dipped mica and a miniature variable, C5, which is mounted on the front panel.

The secondary of the driver tank, L2, consists of two turns of small insulated wire wound over L1 at the B+ end. This provides a low-impedance match to the base of the 2N5189. A 100-ohm resistor in series with L2 and chassis ground helps protect the 2N5189 from damage.

The tank circuit of the 2N5189 final is similar to that of the driver stage. The collector is tapped to L3 at 3 turns from the B+ end. The variable capacitor, C11, is mounted on the front panel directly above C5. A double-pi network consisting



The vfo and buffer stages are built into a 3x4x5-inch Minibox. Extra chassis stiffness is provided by the 1/16-inch aluminum subchassis and panel stiffeners.

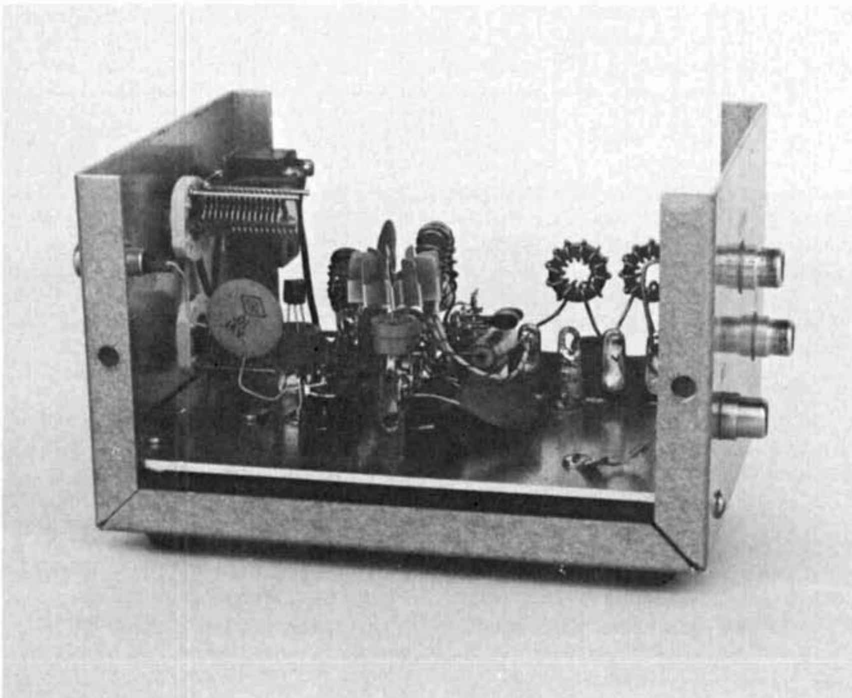
of L5, L6, C12, C13 and C14 is used in the output to assure good harmonic attenuation.

construction

Construction of the driver-final unit is similar to that of the vfo. An aluminum chassis is cut and fitted inside the Minibox using 1/2-inch right-angle stock and a piece of flat aluminum cut slightly smaller than the bottom of the box to permit the cover to slide in place. The chassis is then removed and all the components except the variable capacitors are mounted on it out of the box.

Components of each stage are mounted on two terminal strips – one 5- and one 3-tie-point strip – arranged as an in-line unit. The driver stage is near the front. The double-pi network is mounted at the back on a 4-tie-point terminal strip. This arrangement permits orderly assembly of components and short leads.

The Minibox should now be drilled for the on-off switch, input jack and the two miniature variable capacitors on the front panel and the output jack and the B+ jack on the back. You may also want to include an antenna jack for connection to the receiver if one antenna is used for both transmitting and receiving. This jack should be connected to the transmitter output through a 68-pF capacitor, with reversed silicon diodes to ground on the receiver side of the capacitor to protect the receiver against damage.



Like the vfo, the driver and 2-watt output stage are built into a 3x4x5-inch Minibox. Transistor Q2, a 2N5189, is provided with a clip-on heat sink.

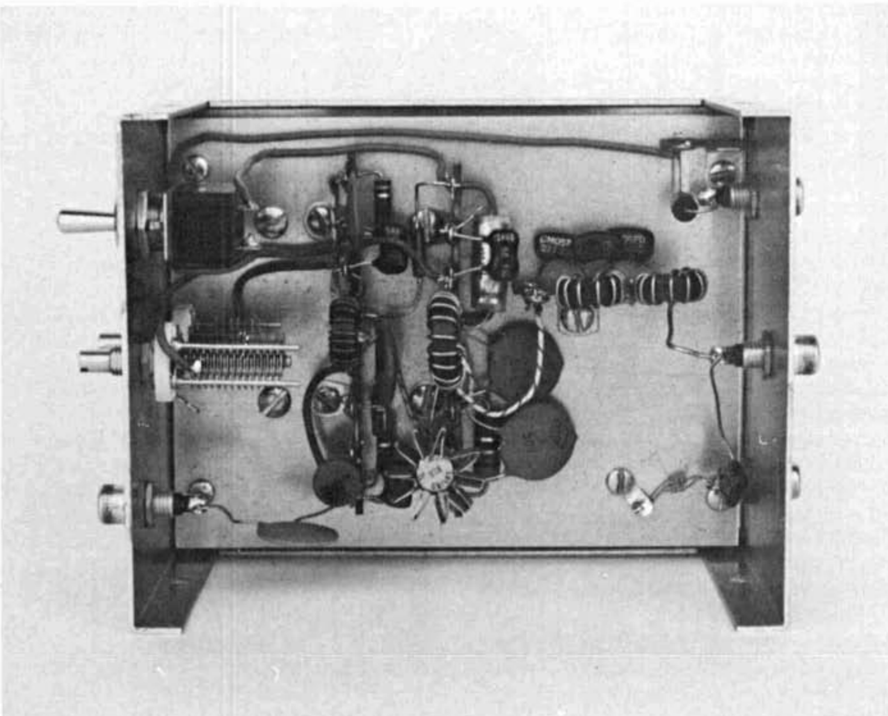
testing

Check both units for possible shorts to the chassis before connecting them to a power supply (and before inserting the transistors, if you use sockets). If the units pass the short test – B+ to chassis – then insert the transistors and connect the output of the vfo to the input of the driver-final using a short, shielded jumper cable. Output of the final should be attached to a 50-ohm dummy load; a two-watt resistor works well.

The units require 12 volts dc. I use a separate 12-volt lantern battery for each unit so as to contribute toward good voltage regulation for the vfo.

with your multimeter. If you have a sensitive reflected power and swr bridge meter, check for output. Adjust the front panel trimmer capacitors for maximum power output at the midpoint of the CW portion of the band.

This completes the adjustments and



Layout of the driver and power output stages. All components are mounted on terminal strips. Tank circuits L1/L2 and L3/L4 are mounted just to the right of the air variable. Inductors L5 and L6 are mounted on another terminal trip near the rear panel.

Now, turn on the power supply switch for the vfo and, while keying, listen for the signal in your receiver. When you have found the signal, set the dial for full capacitance of the tuning capacitor and use the air trimmer to set the low band-edge at 14 MHz.

Next, turn on the driver-final unit and again key the vfo. You should now hear a stronger signal. If you have an rf probe and a vtvm, you can check for rf voltages at points given in the schematic. If you do not have an rf probe, rectify the rf output with a germanium diode and capacitor and read the rectified voltage

checking. The transmitter is ready to be put on the air. An antenna tuner is strongly recommended for matching the transmitter to the transmission line and for additional attenuation of harmonics. The extra resonant circuit provided by the antenna tuner also helps in receiving.

references

1. C. E. Galbreath, W3QBO, "A VFO for Solid-State Transmitters," *ham radio*, August, 1970, page 36.

2. Wes Hayward, W7ZOI, "A Second Generation MOSFET Receiver," *QST*, December, 1970, page 12.

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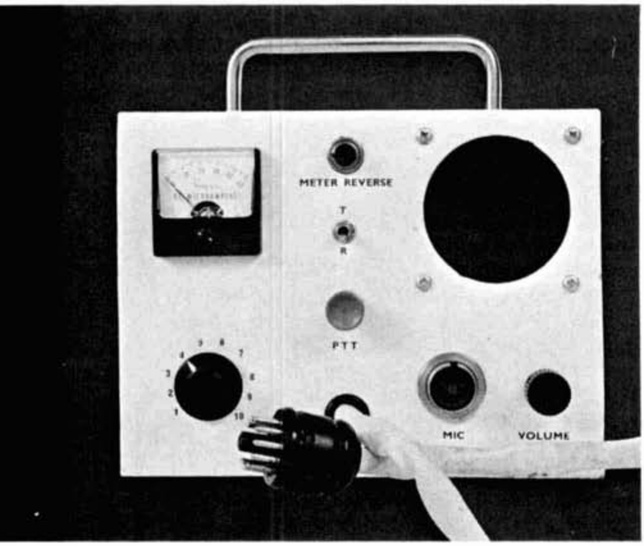


test set

for Motorola radios

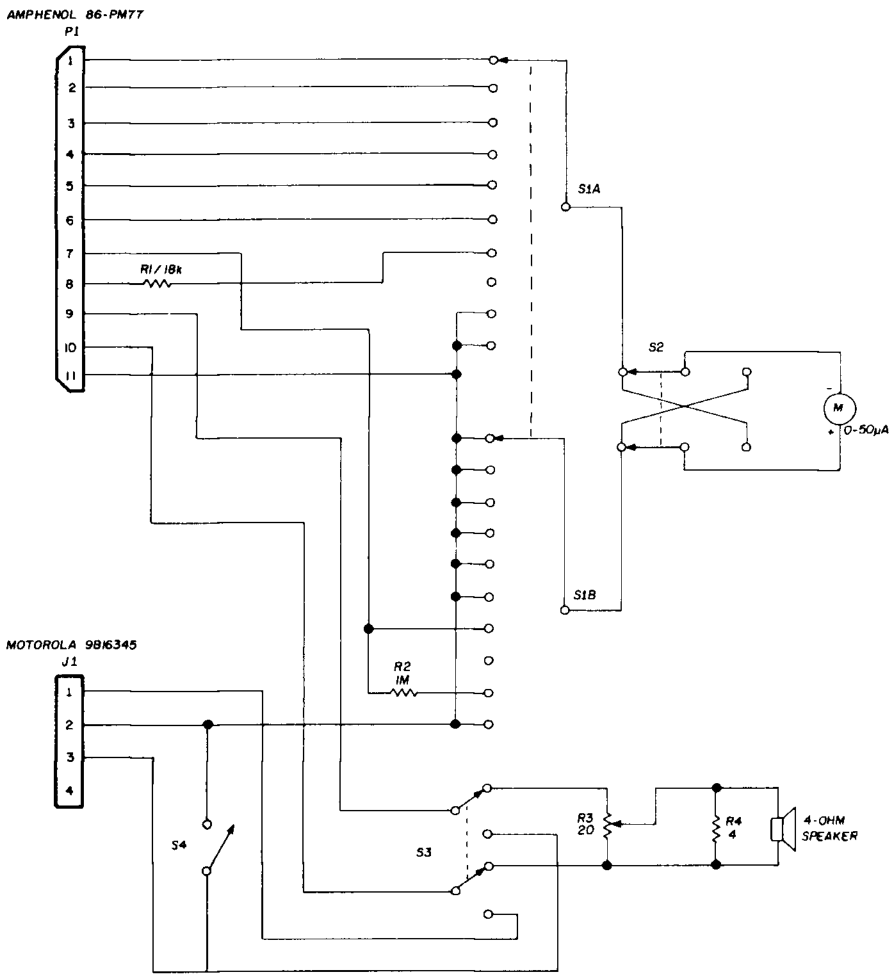
Complete construction details for a test set for most Motorola business-band fm radios

David L. Marshall, KØBKD, Albuquerque, New Mexico

Today is the day of vhf-fm communications. Although there are many amateur-band fm transceivers in use, retired business-band fm sets still hold a sizable portion of the market. This is due to their increased reliability resulting from their ruggedized construction suited to a mobile environment. Another reason is their reduced cost.

One minor disadvantage with old business-band radios is that they usually require specialized test equipment to keep them operating. The average amateur has no way to maintain his own equipment and has to rely on a friend in the two-way radio business. This is fine until the radio quits and test equipment is not available for one reason or another. In my opinion, half the fun of ham radio is maintaining your own equipment.

In this article I will describe a simple but comprehensive test-set for Motorola fm radios. Although you can buy a readymade test-set from Motorola, they cost about $250, which is a bit expensive



M1	50 μA meter (Simpson 1212)	S1	2 pole, 10-position, non-shorting wafer switch (Centralab PA-1005)
R1	18k, ½ watt, 5%	S2	snap-action, momentary contact dpdt switch (Alco MSPE-206R)
R2	1 megohm, ½ watt, 5%	S3	miniature dpdt toggle switch (Alco MST-205N)
R3	20-ohm potentiometer, 5 watt	S4	spst momentary contact switch
R4	4 ohms, 2 watt		

fig. 1. Circuit for the test set for Motorola vhf-fm radios. Adapter cable for Motrac sets is shown in fig. 2.

for a piece of equipment that isn't (hopefully) used too often.

Fortunately, many years ago Motorola showed great insight by designing their radios to be compatible with a universal test-set. Today, the same test-set will align 80Ds, 140BYs, Twin-Vs, T-Powers, utility base stations, Motracs, Motrans, Mocoms and Micors. There are many different models included under these different generic names so it is obvious that a test-set is a most handy piece of equipment to have, regardless of what Motorola radio you may have. The few exceptions are the H23XXX hand-carry, the HT-200, the HT-220, the PT-200, PT-300, etc. Due to space limitations, these sets do not have meter sockets and are aligned with a volt-ohm-meter.

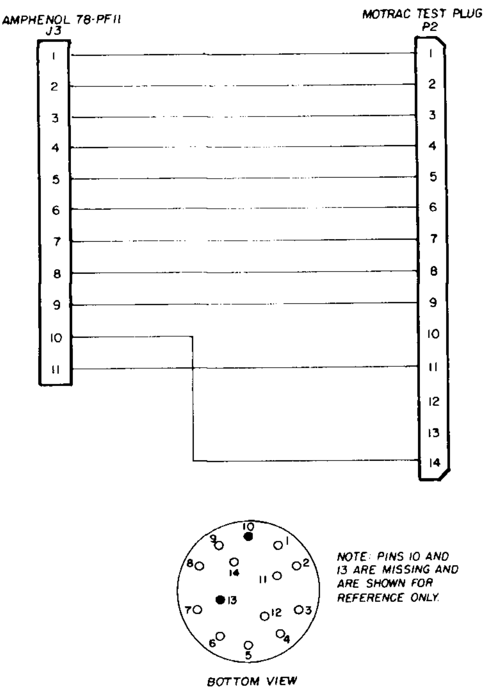


fig. 2. Motrac adapter cable for use with the test set shown in fig. 1. In the bottom view of the Motrac test plug, solid dots are pins, open dots are holes.

test-set

The basic test instrument for checking Motorola radios is a 50-μA meter. By switching in suitable multipliers, a voltmeter can be built which will monitor all test points. There are two additions to this basic meter which make it even more valuable. One is bringing the receiver audio out to a speaker built into the test-set. This facilitates working on trunk-mounted radios where the normal speaker is some distance away. In the commercial version there is also an audio voltmeter used for making signal-to-noise measurements. For reasons of simplicity this feature was not included in the unit described here. Another useful addition to the basic test-set is the provision for transmitter push-to-talk and a microphone connector. These items are extremely useful when you are working on a transmitter with a remotely located microphone.

Although there is nothing critical about the construction of the test-set, there are a few considerations which are worthy of mention. From experience I have found that a cable about three feet long is optimum. To align Motrac, Motran, Mocom and Micor radios an adapter is necessary. This converts the normal eleven-pin test-set plug to a plug similar to a nine-pin tube except that it has four pins in the center. EECO logic modules have this kind of plug, or the adapter can be purchased from Motorola (TKN6025A, price $9.50). If you do not anticipate working on any of these radios this adapter is not needed as the earlier radios used an eleven-pin socket which is compatible with the test-set plug.

Another area of flexibility is in the meter reversing switch. It can be incorporated into the selector switch so that position +4 is normal and position −4 is reverse. Since the reversing switch is used only for discriminator alignment it is recommended that the meter reversing switch be a separate pushbutton switch so that it is not necessary to switch through the reverse position every time. It is also recommended, for reliable operation, that the switch be a good quality, snap action microswitch. Quality switches are a must since changes in contact resistence, encountered with inexpensive switches, can change meter readings.

The selector switch should be a non-shorting type, also of good quality. I happened to use a shorting-type switch because I had it on hand; however, when using a shorting-type switch, if the unit is switched while measuring PA plate current, the power supply fuse will blow because of the momentary short circuit.

The fm test-set will be only as good as the meter, so a good quality meter movement is a must. I recommend using a Simpson meter since Motorola uses this type in their test-set so the difference in internal resistances should be small. Also, the largest available meter should be used to get the maximum needle swing – some of the indications are quite small.

The microphone connector is a special connector similar to an Amphenol 91-PC4F. Occasionally these can be found on a junked Motorola control head. If you can't locate an old one, connectors can be ordered from Motorola (9B16345, price $.95).

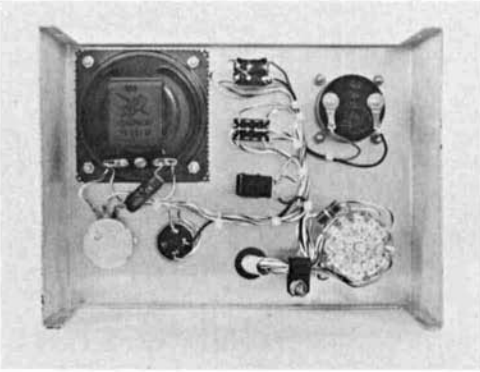
I built my test-set into a Bud CU-3009A Minibox, 3½ x 6 x 8 inches. This size is large enough to allow for modifications such as a larger meter, yet is small enough to be easily portable. The unit has seen a great deal of service in mobile maintenance and working on repeaters. It has always been easily portable.

motorola alignment

For those of you who are not fortunate enough to own a manual on your particular radio, receiver alignment is very straight-forward. However, because of the different number of multiplier stages used in the various frequency ranges, Motorola has chosen to base the alignment on the power amplifier stage. The PA plate is always position seven. The PA drive is position six, the last multiplier is position five, etc. In a low-band radio the switch will go back only to position four. A uhf radio will go back to position one.

Receiver alignment is started by aligning the discriminator for a zero output with an on-frequency signal. There are several methods in use, but generally the primary is peaked with the secondary detuned and then the secondary is tuned for a zero voltage output. It is best to use a 455-kHz signal rather than relying on a received signal. When this is done the 455-kHz i-f stages are aligned again using a 455-kHz signal. The signal level should be kept below the point of saturation on meter position one. If the receiver is badly detuned, position two may be used for a more sensitive indication.

The 12-MHz i-f is aligned by coupling a signal at 12 MHz to the first mixer. The high i-f is then aligned again using position one. At this point an on-channel signal should be copiable, even though weakly. Using this on-channel signal, the rf amplifier should be aligned. When this is done the signal should be further reduced and, using position two, the complete alignment should be touched up for the best sensitivity.

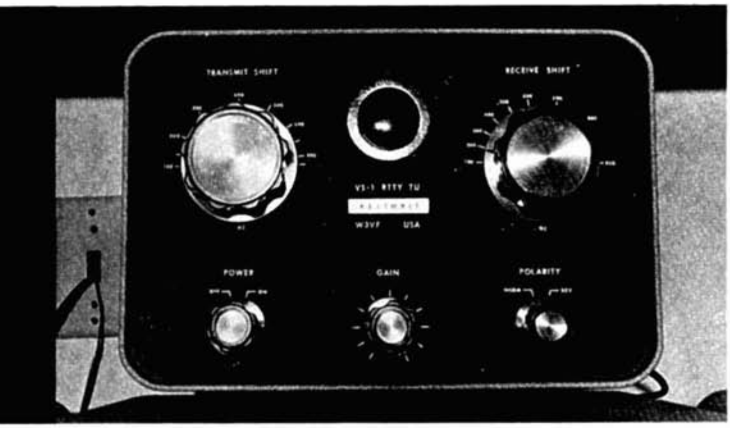


All components for the Motorola test set are mounted behind the front panel.

The final step in the alignment procedure is to adjust the received frequency. This is done by adjusting the first oscillator trimmer for a zero-voltage indication at position four.

Transmitter alignment is simply a matter of tuning each stage for maximum meter indication with the exception of position seven. Position seven is adjusted for a dip in the PA plate current as in any class-C stage. If the loading and coupling controls are adjusted for maximum and then backed off to about 80% the output power will be reasonably close to what Motorola recommends for that particular radio. The final step is to adjust the transmit frequency. I use a 30-MHz frequency counter coupled to the output of the phase modulator. It is important not to load the oscillator as this will change the frequency when the counter is removed. By using a one-second time-base on the counter, you will obtain accuracy that is entirely adequate for amateur vhf operation.

ham radio



variable-shift RTTY terminal unit

And now the VS-1 — with continuously variable filter tuning for improved response in noise and interference

Keith Sueker, W3VF, 110 Garlow Drive, Pittsburgh, Pennsylvania

Phase 1 of my RTTY experience began with the gift of a Model 26 Teletype machine from Mac McKinley, W3OB. During my first attempt to print something off the air, a simple single-tone detector and vacuum-tube keyer were used.[1] This system worked but with poor results in noise and interference. At this point I began to learn something about TU problems and their solutions.

Phase 2 was introduced by a version of Irv Hoff's ST-5,[5] which was modified to use tone pairs of 1000/1170 Hz and 1000/1850 Hz for narrow and wide shift respectively. These frequencies were chosen to allow the sharp CW selectivity in my SB-300 to be used for narrow shift without cluttering up the receiver with an odd-frequency BFO crystal for RTTY. This system unit worked well enough to allow some 20 meter DX.

Phase 3 began with some serious considerations of how to build a better TU in terms of noise and interference response. Advanced design units, such as the ST-6[6] with switched discriminator filters, do an excellent job on narrow shift but have an excessively broad response for marginal receiving conditions on wide shift. The phase-locked loop units[7] have the same problem, since they are subject to interference through the whole lockup range of the vco. A solution is to use sharply tuned filters, in or ahead of the discriminator, which will pass only the individual tones of interest and their sidebands. When fixed-tuned filters are used, however, one runs into the problem of various amateur shifts and the many commercial stations that use shifts other than 170 and 850 Hz.

The ST-6 discriminator filters at 170-Hz shift are about as narrow as one might wish to use. The 850-Hz shift filters, however, are broad enough to provide a linear discriminator characteristic over 850 Hz. If we sharpen these filters, we can reduce the interference bandwidth, but we still must cope with the various intermediate shifts in use. Why not make one of the filters with a continuously adjustable center frequency? Then we can tune any shift.

the VS-1 TU

The filters are the key to the whole project. If one filter were continuously tunable, what kind of filter and tuning system could be used? Active filters using RC networks are popular, but high Q requires critical feedback adjustment and stability is questionable with simple techniques. Toroidal LC filters using 88-mH inductors provide excellent Q, but how does one tune such a device? To accommodate the shift range between 170 and 850 Hz, one filter must be varied nearly 700 Hz — a task clearly impossible for an ordinary variable capacitor. Also, resistive loading would have to be continuously variable to keep the noise bandwidth matched to the fixed filter.

The solution to these and other problems turned out to be relatively simple: both audio tones were increased to higher frequencies so that one of the filters could be tuned with a variable capacitor. Some calculations showed that a frequency of about 11 kHz would allow a sufficient tuning range with a 365-pF capacitor. At 11 kHz, the tuning range would be less than ten percent of center frequency, providing good noise balance without compensation.

By going to a high frequency, the absolute bandwidth would be increased by some four times over that obtainable

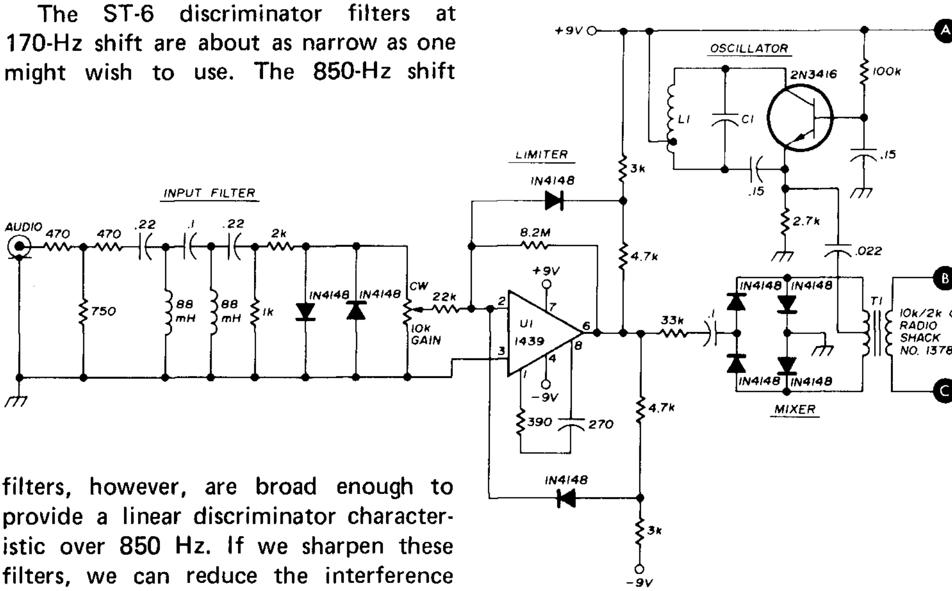


fig. 1. Input circuit and mixer for the RTTY terminal unit. For values of L1 and C1, see text.

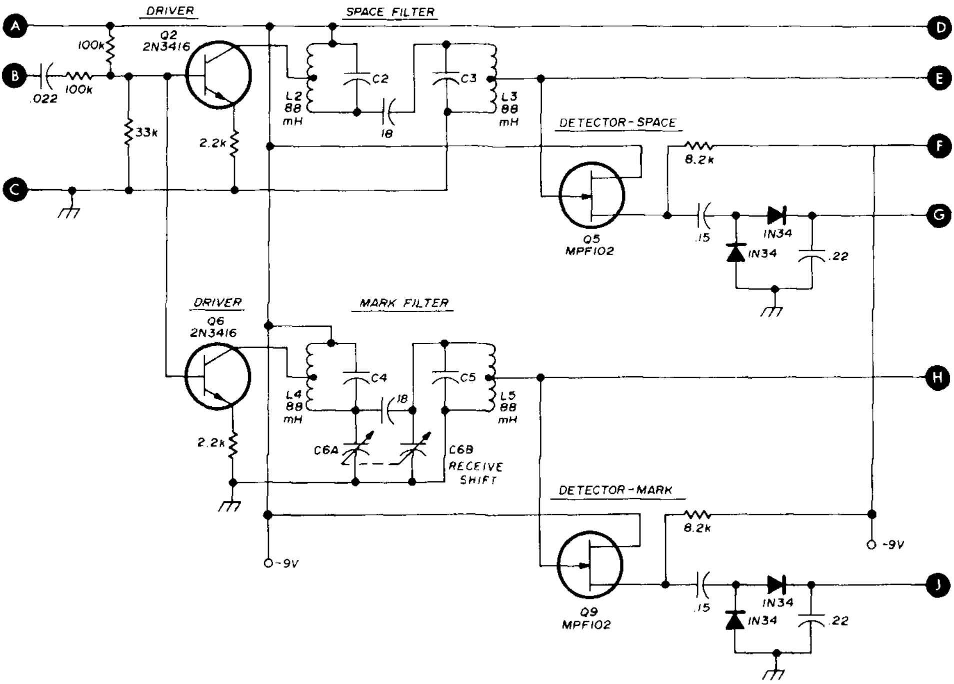


fig. 2. Mark - space filters and detectors for the RTTY terminal unit. For value of tuning capacitors, see text.

at the usual audio tones. This problem was resolved by using two filter sections with loose coupling and a dual-section capacitor for tuning. The selectivity turned out to be ideal.

input circuit and mixer

Whether using fm or limiterless a-m detection, it is important to restrict incoming audio to the frequencies of interest. Most modern receivers have i-f filters with sufficient high-frequency cutoff. However, the low-frequency response should extend no lower than the bottom of the i-f passband. Hence, a high-pass filter is used on the VS-1 input to reduce hum and noise (**fig. 1**). Toroidal 88-mH inductors are used, and the filter is designed for a low-frequency cutoff of 1000 Hz. The input pad allows satisfactory operation from any receiver output impedance. Measured attenuation is 1 dB at 1000 Hz, 13 dB at 700 Hz, 29 dB at 500 Hz, 38 dB at 300 Hz, and more than 40 dB for lower frequencies.

Op-amp U1 and the associated circuit form a high-gain stage, which can be operated as a linear amplifier or a symmetrical clipper (limiter) by varying the input gain control. The gain control could be replaced by a switch as in the ST-6; but a control provides more audio range.

Once the audio tones have been amplified and/or clipped, they are heterodyned *to the filter frequencies by the local* oscillator and balanced mixer. The oscillator is a Hartley circuit with a tap on L1, which is made by unwinding 50 turns from an 88-mH toroid, attaching a lead, and rewinding the turns in the same direction. Capacitor C1 tunes the oscillator to the desired injection frequency. (More about this later.) Use of an 88-mH toroid ensures that the oscillator and filters will track over a wide temperature range. The balanced modulator is conventional. It uses unmatched diodes and is coupled to succeeding stages through an inexpensive transistor interstage transformer.

filters

Transistors Q2 and Q6 are filter drivers (**fig. 2**). The high collector output impedance, coupled with feed at the center tap of the toroid, results in negligible filter loading. Each filter consists of two LC sections using 88-mH toroids with loose coupling provided by the 18-pF capacitors. Space filters L2,C2 and L3,C3 are fixed-tuned to the heterodyned space frequency. Mark filters L4,C4 and L5,C5 are identical to the space filters except for the dual variable tuning capacitor, C6, which tunes the filters to the heterodyned mark frequencies.

frequency selection

Up to this point discussion of the frequencies has been rather vague. This TU is basically a superhet receiver with dual i-f channels, one of which can be tuned. A nominal i-f must be used that can be varied by the frequency difference between maximum shift (usually 850 Hz) and minimum shift (usually 170 Hz) with a ganged capacitor of reasonable size. For a shift difference of 680 Hz and a dual 365-pF broadcast capacitor, the minimum possible filter frequencies are about 11,000 Hz and 11,680 Hz for the tunable filter. If the other filter is set at either 10,830 Hz (11,000-170 Hz) or 11,850 Hz (11,680 + 170 Hz), the entire range of shifts from 170 Hz to 850 Hz can be tuned with the variable capacitor.

space and mark. Common practice in amateur RTTY is to shift frequency down for space. If the receiver sharp filter is to be used, the lower of the two audio frequencies must fall at the lower end of the sharp passband – about 950 Hz in the Heath SB series. Narrow shift then has

table 1. VS-1 frequency relationships.

	frequency (Hz)	
	170 shift	**850 shift**
audio space	950	950
audio mark	1120	1800
oscillator	9880	9880
space filter	10,830	10,830
mark filter	11,000	11,680

space at 950 Hz and mark at 1120 Hz, while wide shift has space at 950 Hz and mark at 1800 Hz for reception on USB. If LSB is used, the frequencies will be inverted. Since the space frequency is fixed for USB reception, the mark filter should be tunable.

We have defined 11,000 Hz to 11,680 Hz for mark frequencies and 10,830 Hz or 11,850 Hz for space. At this point I chose 10,830 Hz so that (a) 170 Hz would fall at the maximum (counterclockwise) setting of C6, and (b) the dial could be calibrated clockwise in increasing shift frequency. The local oscillator must be tuned to the heterodyned space frequency of 10,830 Hz minus the audio space frequency of 950 Hz for a resultant

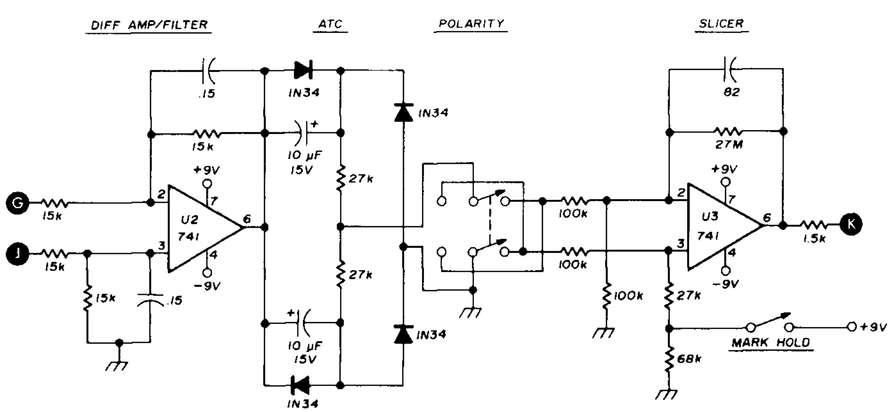


fig. 3. Schematic of the automatic threshold corrector (ATC) and slicer.

of 9880 Hz. **Table 1** should make things a little less confusing.

oscillator tuning. The frequency conversion process has been described in some detail, because oscillator tuning must match the receiver bandpass characteristics, and some adjustments may be needed. Also, the oscillator can be moved in frequency to match the more usual tones of 2125 and 2975 Hz for 850-Hz shift if desired. An understanding of the conversion process is necessary to customize this portion of the VS-1. Oscillator tuning can be by means of a front panel control if several different audio frequencies must be used. The filter frequencies may also be increased if a wider range of shifts must be accommodated, but selectivity will suffer.

detectors, ATC, and slicer

Source followers Q5 and Q9 provide stiff drive sources for the detectors while presenting negligible load to the filters (**fig. 3**). Mark and space detectors are identical voltage doublers using germanium diodes for low voltage drop. The outputs are combined in differential amplifier U2, which also provides some filtering. Filter time constants allow transmission rates to 100 wpm. ATC circuitry following the amplifier is from the ST-6, as is the polarity reversing switch. The slicer, U3, is another differential stage, which further squares the keying pulses and adds more filtering. It also has a mark hold input to lock up the keyer for transmitting. In my equipment, this switch is located next to the machine keyboard, and another pole is used to actuate the transmitter. Three positions are used: "Print," "Mark-hold," and "Transmit."

keyer and FSK

Transistor Q10 (**fig. 4**) drives the parallel-connected magnet circuit at the 60-mA level. The 20-ohm resistor in the emitter return provides a signal to the FSK circuit to detect mark current when the keyboard is actuated for transmitting. The mark signal is sufficient to drive U4, the FSK stage, sharply to zero from its "on" biased state. On space signals, it snaps back on to full output.

FSK control is simple but effective. The positive output voltage of U4 is clipped at +6.2 V nominal by the zener, and the feedback diode limits negative excursions to a negligible value. Thus, we have established two regulated voltages: +6.2 V for space and zero volts for mark. The two divider pots set the end points for the shift pot, so that a continuously variable positive space voltage can be supplied to the transmitter vfo. Note that the shift voltages are not affected by loop current or contact resistance in the keyboard.

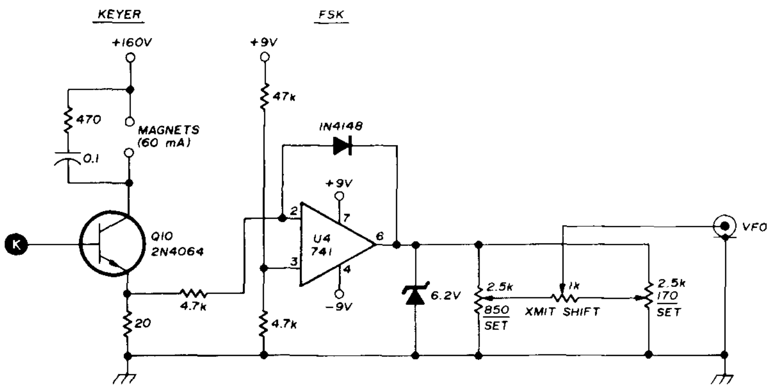


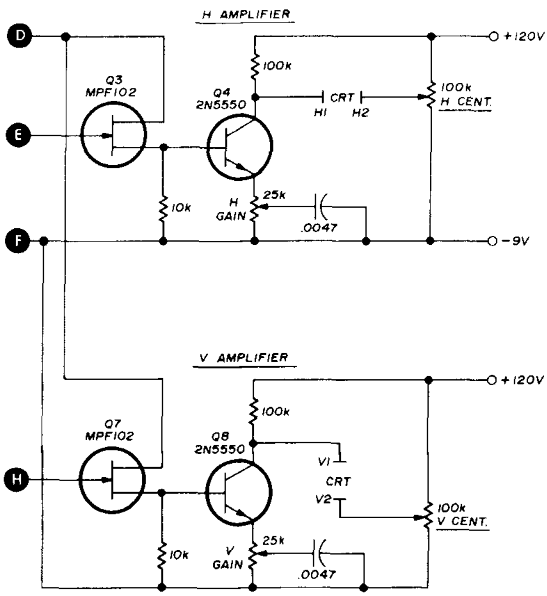
fig. 4. Circuit for the keyer and frequency-shift keyer used in the VS-1 RTTY terminal unit.

Any of the usual transmitter FSK schemes can be used as long as the voltage requirement does not exceed 6.2 V. Mark and space can be inverted by simply inverting the inputs to U4. Output polarity can be reversed by reversing the feedback and zener diodes.

Keying turned out to be simple on my SB-401. The shift pot is connected across the vfo sideband selector diode. The diode is supplied from a high-resistance circuit, and the low output resistance of the shift pot shorts out the normal bias and sets a new level in "transmit." The actual shift depends slightly on vfo tuning, but it is essentially constant over the usual RTTY frequencies, so the shift pot can be calibrated directly in shift frequency from 170 to 850 Hz. The control shown in the photos is a standard pot equipped with a homebrew stop at 180 degrees.

tuning indicator

AC outputs from the filters are isolated by source followers Q3 and Q7, which drive deflection amplifiers Q4 and Q8 (see **fig. 5**). Separate isolators are used, because the waveforms at the sources of Q5 and Q6 are affected by detector loading. The 2N5550 transistors will deliver about 100 V p-p, which is enough to drive most small CR tubes. Both deflection channels are direct coupled to eliminate trace jumping from noise or overload. The variable bypass circuits in the emitters provide gain control independent of positioning. My VS-1 uses a 1EP1 CRT, but only because I found one at a hamfest for five bucks.



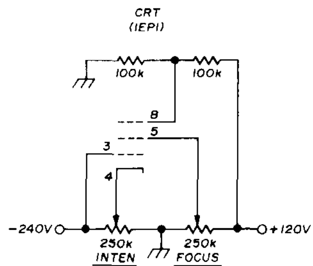


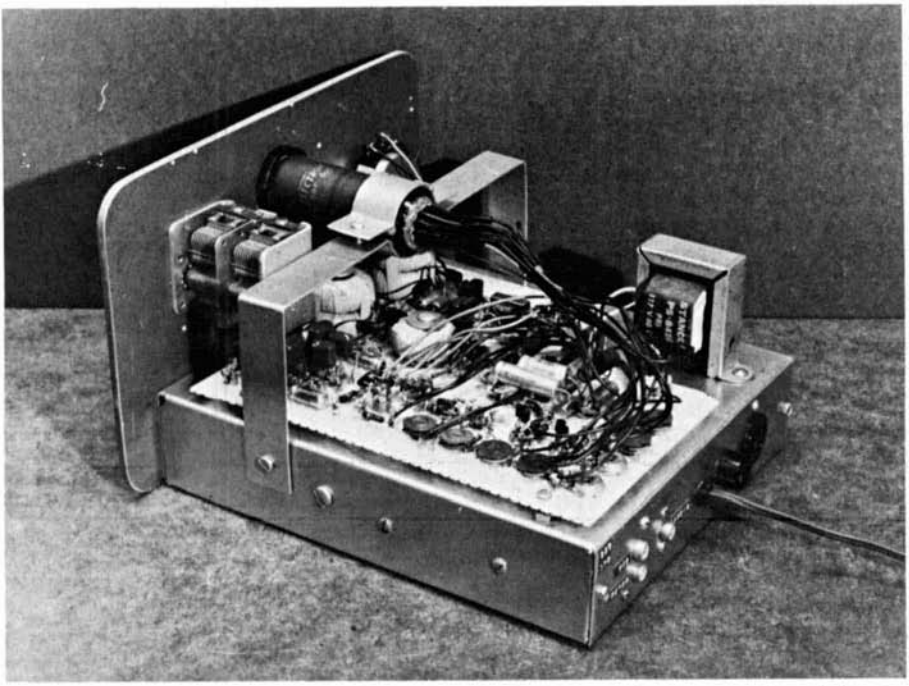
fig. 5. Tuning indicator used with the 1EP1 chathode-ray tube used in the author's unit.

These tubes are too expensive to buy new. Any of the small oscilloscope CR tubes can be used with equal success.

power supplies

Power for the dc loop and low-level stages is derived from a Stancor PS8416 transformer (**fig. 6**). Although rated for only 25 mA on the dc output, the lack of appreciable loading on the 6.3-V winding keeps transformer temperature reasonable. The 2-μF capacitor provides a higher inrush current on mark signals and allows the average mark hold current to be reduced to 50 mA without degrading print quality.

Low voltages are generated by a pair of voltage doublers and 9.1-V zener diodes. There is nothing mysterious about this arrangement; any low-ripple power supply would work as well. A pair of 6.3-V filament transformers provide an isolated supply for the CRT heater and dc supplies for the intensity, focus, and deflection circuits. Here again, any standard configuration[8] could be used. However, the loop supply should not be used for accelerating or deflection voltage supplies, because the regulation causes trace problems.



Construction of the VS-1 RTTY terminal unit. All circuits except power supply and keyer are mounted on a section of perforated circuit board. Power supply components are mounted under the chassis.

construction and components

Few of the parts in this unit are at all critical, and the mechanical layout is not too important. Common-sense precautions should be taken to isolate filter input and output circuits, but no special bypassing or decoupling is required for stability. Leads to the mark tuning capacitor should be reasonably isolated to minimize coupling around the 18-pF capacitor. If a CRT tuning indicator is used, the tube should be kept away from power transformers.

Except for U1, which handles 10+ kHz signals, all op-amps can be garden-variety 741 types. These ICs are inexpensive and are exceptionally immune to damage. Dual units are available for about $1.00 in TO-type packages from a number of *ham radio* advertisers. Type 709 op-amps could be used, but they are more subject to latchup, require compensation, and will not tolerate output shorts.

Transistors and fets are also garden-variety types except for the high-voltage deflection and keyer units. The fets hang between the power-supply rails, so they should have voltage ratings of at least 20 V. Nearly any npn transistor can be used in place of the 2N3416s. Silicon diodes are low-leakage computer types, and the 1N4148 is suggested for those who must buy them. Power diodes can be any silicon types rated at least 0.5 A and of voltage ratings listed on the schematic.

Germanium diodes can be any rf type but should be checked for high leakage. Capacitor types are unimportant except for C1 through C6, which must be low leakage types such as mylar, polystyrene, or polycarbonate. Trim capacitors can be these types or mica. The total capacitance required at C1 through C5 is 0.002 to 0.003 μF.

Once the filters have been tuned, the oscillator coil should be tuned to roughly 9880 Hz. Next, run a steady signal through the receiver and adjust tuning until two incoming frequencies, 170 Hz apart, are centered in the narrow passband. Trim the local oscillator frequency in the VS-1 to place the lower audio frequency at the peak of the space filter.

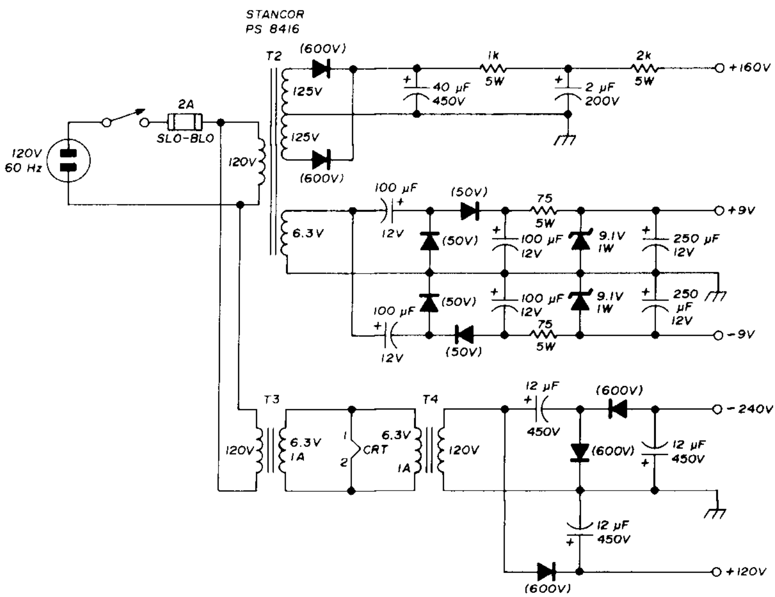


fig. 6. Power supply for the complete VS-1 RTTY terminal unit. Required PIV of diodes is indicated in parenthesis.

adjustments

After construction and wiring are completed, the usual dc voltage checks should be made. Assuming things look normal, the filters may then be tuned. A precise frequency source is not needed, since any errors in alignment will be compensated by the oscillator alignment.

Set the alignment generator to 10,830 Hz, add temporary 82-pF capacitors across L4 and L5, set C6 to maximum, and tune all sections for maximum output from Q5 or Q9. The unit may be fine tuned either by capacitor selection and trimming or by pruning turns. Remove padders, and filter alignment is complete. The variable capacitance of C6 provides the required frequency offsets.

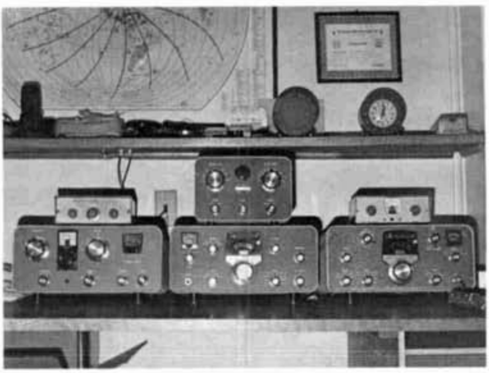
It should now be possible to adjust C6 to place both space and mark signals at the peaks of their respective filters for any shift frequency from 170 Hz to 850 Hz. If a number of different receivers or different audio tones are to be used, a variable tuning capacitor can be used. A 730-pF unit will provide enough range.

modifications

Motor control and automatic mark hold circuits have been deliberately omitted from this unit, but they undoubtedly will be required by some operators. The tail-end circuitry of the VS-1 is sufficiently similar to the ST-6 so that the same circuit techniques are directly applicable. The power supply may have to be beefed up to handle the additional load, however.

Some type of CRT display is definitely recommended, since the filters are much too sharp for hit-or-miss tuning. It may be possible to use one of the dual magic eye tubes, but it's doubtful if meter schemes are feasible.

The FSK system shown is by no means the only one that will work. Any of the standard techniques can be used for relay operation directly from the keyboard.



Station at W3VF. The design of the VS-1 is similar to the Heathkit transmitter and receiver used by the author.

For those interested in a hybrid scheme, the 741 at U4 will drive a high-sensitivity reed relay directly with a 12-Vdc coil. This will allow the VS-1 to be used with existing FSK installations. About 10-mA of current is available.

operation

Reception with the VS-1 requires accurate tuning and a reasonably stable receiver. The receiver is tuned to peak the space signal on the appropriate CRT axis, and the shift tuning is then varied to peak the mark signal. Selectivity may be a little scary at first, but only a poor receiver or a drifting transmitter will cause problems. The variations in shift are rather interesting to watch. Advantage should always be taken of the highest degree of receiver selectivity that will pass both space and mark signals.

performance

The VS-1 performs well. Filter selectivity is sharp enough to present a clean cross pattern on 170-Hz shift, and the print holds up until signals disappear into the noise. The ATC circuit works well at machine speed but can't cope with keyboard sending under poor conditions with severe signal fading, particularly for loss of the mark signal. A manual offset on the slicer would be required for this condition. Performance in interference is very good, and the improvement in print from using sharp receiver selectivity on 170-Hz shift is striking.

Despite the many words written about limiter/discriminator versus limiterless two-tone reception, the VS-1 shows little difference in print quality as the input gain is varied. Even under poor signal conditions, the limiter can be kept saturated with no increase in print errors. This condition is probably due to the narrow filters, but I'll leave that for others to comment on. The unit is a pleasure to operate, and it now occupies a permanent spot in my station. I hope this article serves to stimulate further work toward improvements.

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ham radio

More Details? CHECK—OFF Page 110

medium current polarity inverter

Complete construction details for a polarity converter which will provide a negative voltage supply from a positive source

J.R. Laughlin, San Jacinto College, Box 200, Pasadena, Texas 77505

Now you can go mobile with that favorite circuit that has been "grounded" for lack of negative power supply. An increasing number of semiconductor circuits and devices are appearing that require both a positive and a negative power supply for proper operation. For ac-powered installations this is no problem. However, in the automobile or for applications where only storage batteries are available a real problem can develop in supplying the needed negative power.

Often the designer can resort to trick circuitry in order to operate from a positive source. More often, especially in the case of ready made circuits, nothing within reason can be done. The polarity-inverter shown in **fig. 1** offers a convenient solution to this problem by supplying a negative potential with power enough to suit most applications. This circuit operates with great efficiency and is virtually noiseless due to its high frequency of operation and lack of the usual power transformer. Filtering is a snap at this high frequency, requiring only nominal values of capacitance to do the job.

The output voltage is well regulated and is adjustable, and the 2-ampere maximum provides up to 22 watts of negative power from your auto battery.

circuit operation

A block diagram of the polarity inverter is shown in **fig. 1.** The square-wave generator is a stable unijunction transistor oscillator which drives a transistor (Q2) to produce the square wave. Charge time for C2 is controlled by the value of R1 (see **fig. 2**). Discharge time for this capacitor is controlled by R2.

While C2 is charging, Q2 is saturated; during the discharge time of C2, Q2 is turned off. By proper choice of values for R1 and R2, the output of Q2 will be a symmetrical square wave. Transistor Q3 buffers this output and furnishes driving power for Q4 and Q7. Transistor Q4 inverts the signal from Q3 and supplies base drive to Q5, an emitter follower, which has the capability of driving Q6 into complete saturation.

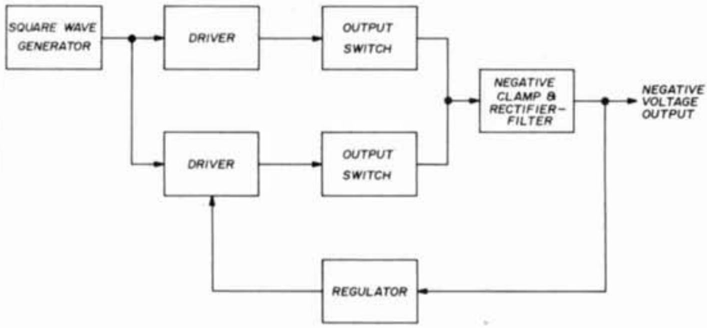
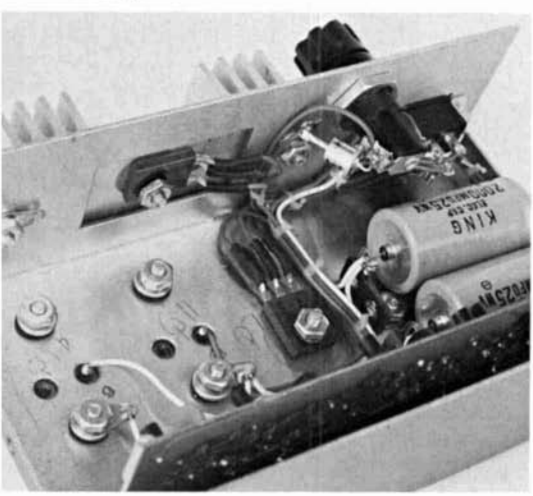


fig. 1. Block diagram of the polarity inverter. With input from an automobile battery, this unit will furnish up to 200 mA at -13 volts, and up to 2 amps at -10 volts.

Power transistor Q6 (2N3055) is mounted on heatsink through an opening in the enclosure. Transistor Q7 (2N2955) uses chassis for heat sinking.

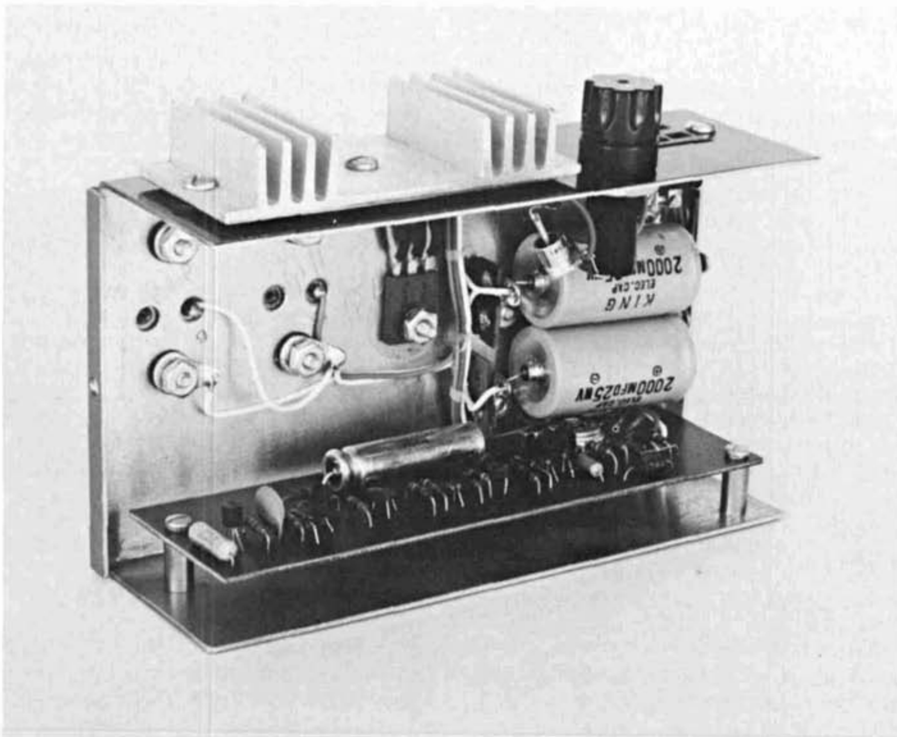


Transistors Q7 and Q6 are connected to form a complementary-symmetry output stage. These output switches are connected in a manner that allows them to be operated in a saturated mode and are driven hard to insure low V_{sat} when they are turned on. This minimizes voltage losses across the transistors and permits the highest possible output voltage.

The circuit configuration of transistors Q3, Q4 and Q5 insures that only one of the two output transistors can be turned on at a time. This arrangement is similar to the standard audio amplifier output circuits found in many stereo sets.

As Q7 and Q6 are alternately switched on and off, a square wave that alternates from ground to nearly battery potential is applied to the positive end of C3. Transistors Q10 and Q11 are connected as diode rectifiers. These are germanium transistors and their diode junctions can handle a lot of current with low voltage drop. Their low voltage drop prevents loss of output voltage that would have resulted if silicon diodes had been used.

Transistor Q11 works in conjunction with C3 to clamp the square wave output of Q6 and Q7 negatively. Thus, at the junction of C3 and Q11 and Q10 a square wave is present that varies from approxi-



Construction of the solid-state polarity inverter. Majority of components are mounted on printed circuit board (fig. 3) which is installed on half-inch spacers. Heatsink is for power transistor Q6. Transistors Q10 and Q11 are mounted on top of enclosure.

mately ground potential to a negative value almost equal to the positive value of the battery voltage. This negative-going square wave is rectified by Q10 and filtered by C4.

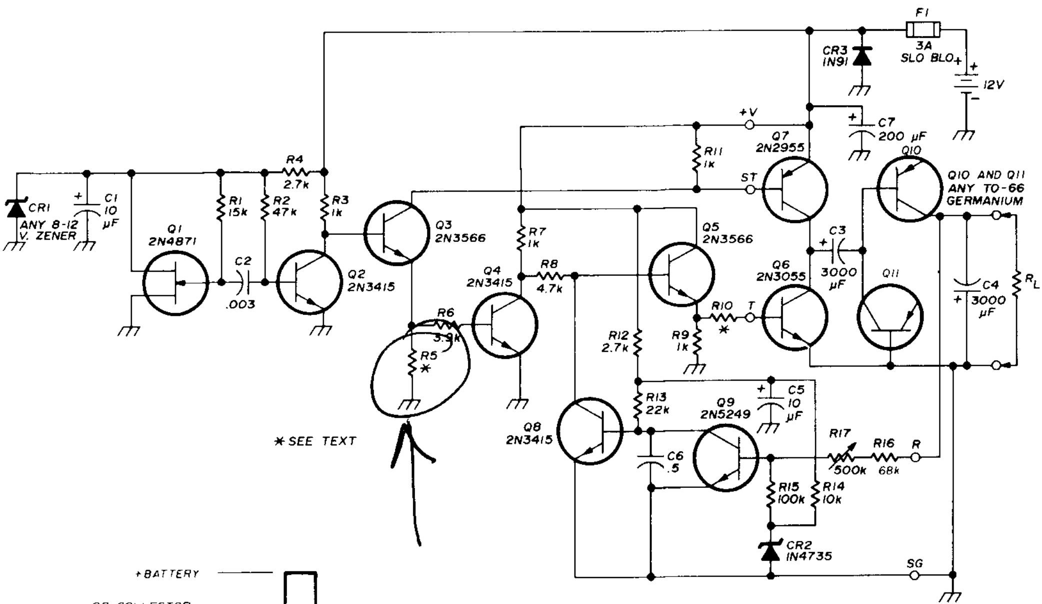
The value of negative voltage across C4 cannot be quite equal to the value of battery voltage because the combined voltage drops across Q6, Q7, and Q10 and Q11 combine to reduce the total output. Naturally, these losses increase with heavier output currents. **Table 1** shows available output voltage versus output current.

table 1. Maximum available output voltage vs output current.

output current (amps)	output voltage (-Vdc)
0	14.0
0.2	13.0
0.4	12.7
1.0	11.4
1.5	10.9
2.0	10.0

Regulation of the output voltage is performed by a transistor feedback loop consisting of Q9 and Q8. A zener diode, CR2, provides a stable reference voltage with which to compare the negative output. This comparison is done by Q9. If the negative output voltage attempts to increase, the conduction of Q9 will decrease. The positive-going collector of this transistor will increase the conduction of Q8, thereby preventing the base drive to Q5 from reaching its full positive value.

Reduction of this drive prevents Q6 from being turned on all the way during half of the cycle, resulting in a lower output voltage to the negative clamp circuit and rectifier. Thus, the output voltage is maintained automatically at a level which depends on the relative values of R15 and the series combination of potentiometer R17 and R16. If R16 is



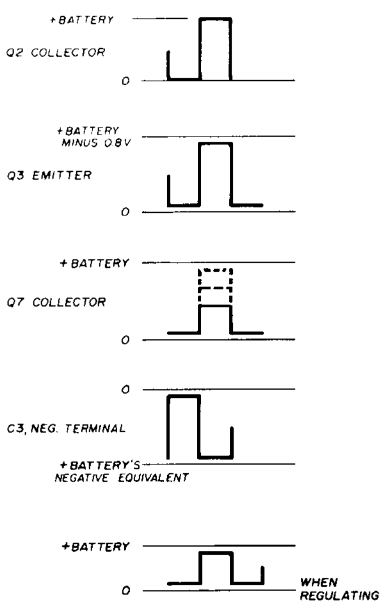


fig. 2. Schematic diagram of the polarity inverter and important waveforms. See text for value of R5.

reduced in value, regulation will occur at a lower output voltage. **Table 2** shows approximate output voltage levels for various values of R16 and R17.

Efficiency depends on the difference between the battery supply current and the output current. For example, with a 2-ampere output current the input supply current from the battery will be approximately 2.2 amps. This additional 200 mA of battery current represents a loss of efficiency. Most of this 200 mA is current drive to the bases of Q6 and Q7 through their base-current limiting resistors, R10 and R11. This amount of base drive is not necessary if the inverter is to operate with loads lighter than 2 amps. These resistors may be increased as shown below for the reduced output currents.

maximum output current (amps)	R10 and R11 (ohms)
2	100
1	240
0.5	330
0.25	680

Increasing these resistor values will substantially increase the operating efficiency for lower operating output currents.

construction

It is suggested that you use the circuit board shown in **fig. 3.** This can be mounted into an aluminum box along with the other components. Heat sinking is not critical for any of the transistors except Q6. This transistor can dissipate quite a lot of power when delivering heavy output current at low output voltage level.

Diode CR3 is not a necessary part of the circuit but was included after the original unit was completed and connected to the battery in reverse! This diode will cause the fuse to open up instead of should there be a wiring error. An ammeter in the battery line should indicate approximately 200 mA with the unit turned on and unloaded. This current will be less if higher values of R10 and R11 are used, as discussed earlier.

It is a good proceedure to start at transistor Q1 with an oscilloscope and check each waveform as shown on the schematic. Verify the amplitude of each waveform. If all waveforms are normal, connect a 5-ohm power resistor across the

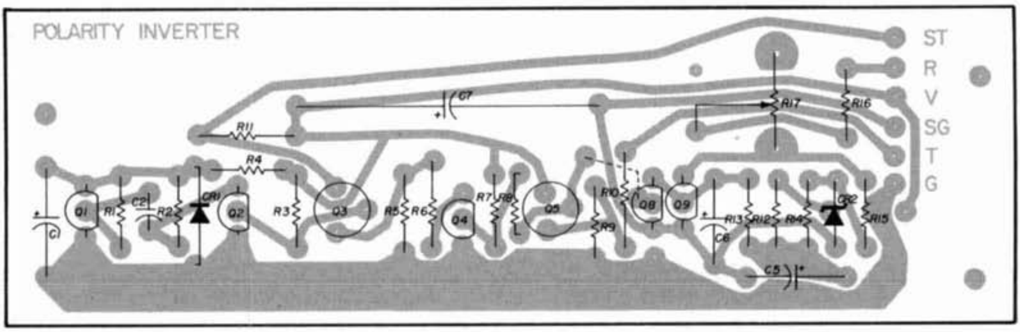


fig. 3. Full-size printed-circuit layout for the polarity inverter.

one or more output transistors if this should happen again. Transistors Q10, Q11 Q6 and Q7 are mounted to the aluminum chassis at widely spaced intervals to facilitate heat dissipation.

checkout

Be certain the fuse is in place before connecting the unit to a battery. Do not bypass the fuse during checkout. It may save a relatively expensive transistor

table 2. Output voltage vs total resistance of R16 and R17 (battery maintained at +14.2 Vdc).

R16 + R17 (ohms)	output voltage (-Vdc)
220k	10
180k	8
160k	7
135k	6
110k	5
96k	4
70k	3

output terminals. With a 5-ohm load, an output voltage of ten volts represents full loading of the inverter. Of course, if you plan to operate at lower output current levels, a higher value resistor may be used for loading the output during the checkout.

Rotation of potentiometer R17 should produce a variation in the output voltage. Alternately connecting and disconnecting the load resistor will verify the regulator action. It should be noted that the maximum regulated output voltage cannot exceed that voltage shown in **table 1** for the various values of peak output currents shown.

For example, if your audio power amplifier draws from 0.05 to 1.5 amps on peaks this inverter can supply 10.9 volts of negative regulated dc. If your requirements are from zero to 0.4 amps, the inverter can deliver −12.7 volts dc.

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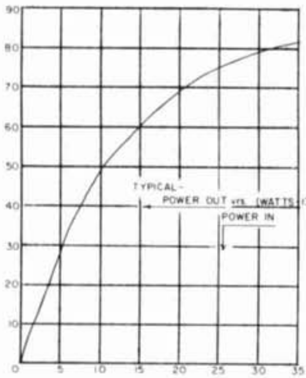

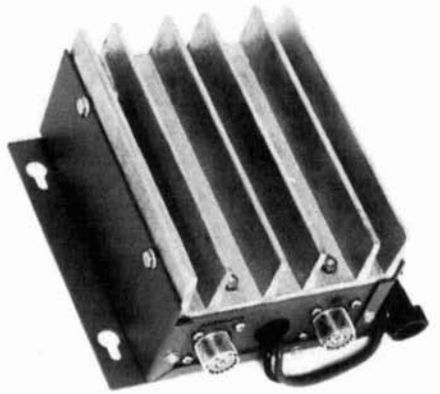

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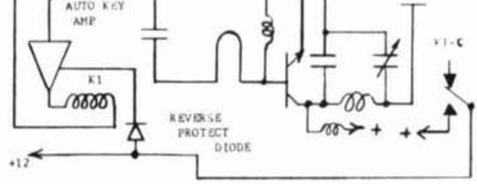

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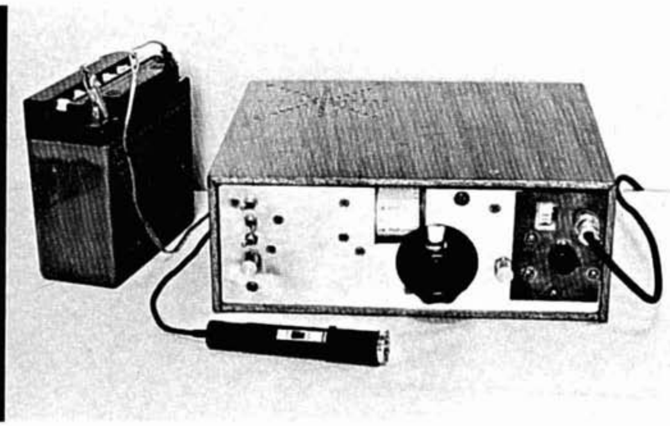


single-band ssb transceiver

using the LM373 communications IC

How to use the versatile LM373 and several other ICs to build a compact ssb transceiver for 14 MHz

Charles H. Hill, W5BAA, 2510 Beatty, Houston, Texas 77023

About two years ago a new products announcement in *ham radio* described a linear communications IC, the National Semiconductor LM373. Although I have found the LM373 to be the most versatile IC for the communications field on the market, I have not seen any articles in the amateur publications which have given the LM373 the praise which I feel it deserves.

One of the strongest assets concerning the device is the very comprehensive data which the National Semiconductor Corporation supplies on request. With this data you can use the IC in a variety of signal processing roles with a minimum of brain work. In order to sing the praises of the device, I will describe a 20-meter QRP ssb transceiver which I have built around the LM373. Although sufficient information is provided in this article so that you may homebrew your own version, I would strongly suggest that you obtain the data sheets on the IC from National Semiconductor in order that you may fully appreciate the versatility of the device.*

communications IC

Fig. 1 is a functional outline of the LM373. National bills the device as an a-m/fm/ssb i-f strip; however, it is used for a host of other functions including dsb generators and receiver frontends. The package includes an agc-controlled gain stage, the output of which may be used to drive a crystal, mechanical or LC filter; a fixed-gain stage, which may be driven by one of the forementioned filters; a balanced mixer driven by the fixed gain stage and an aqc generator which is matched to the agc controlled stage. In addition to the access points for the filter terminations, access is provided for nulling both the signal and local oscillator ports of the balanced mixer.

*National Semiconductor Corporation, 2975 San Ysidro Way, Santa Clara, California 95051.

transceiver

In the transceiver a common i-f strip is built using U1, an LM373 (see **fig. 2**). By switching the input signal applied to pin 2 from the output of the receiver frontend to the output of the dsb generator, the local oscillator signal at pin 6 from the 9-MHz bfo to the 5-MHz vfo, and the output at pin 7 from the audio amp to the 14-MHz filter, one LM373 IC acts as both a receiver i-f strip with a built-in product detector and agc system, and transmitter filter and hf mixer with a built-in agc controlled speech compression. Of course, a 9-MHz ssb filter, FL1, placed between pins 9 and 4 provides the necessary filtering. All of the switching is handled by diode signal swtiches.

In addition to i-f duties, the LM373 is also used as a dsb generator. In this case the output of a dynamic microphone is fed directly to pin 4 of U2, the fixed gain

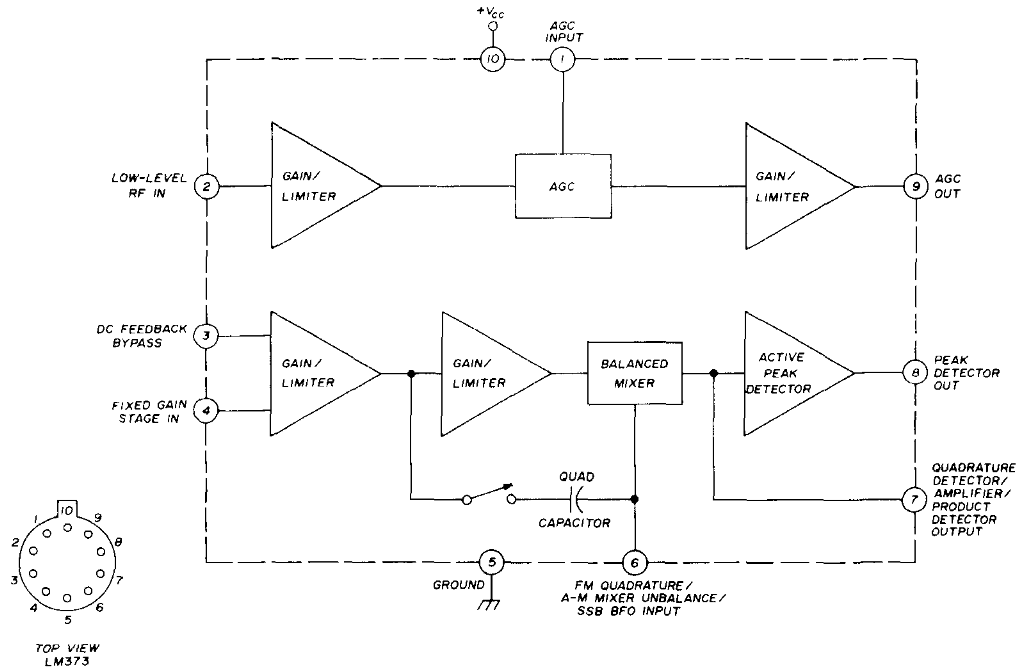
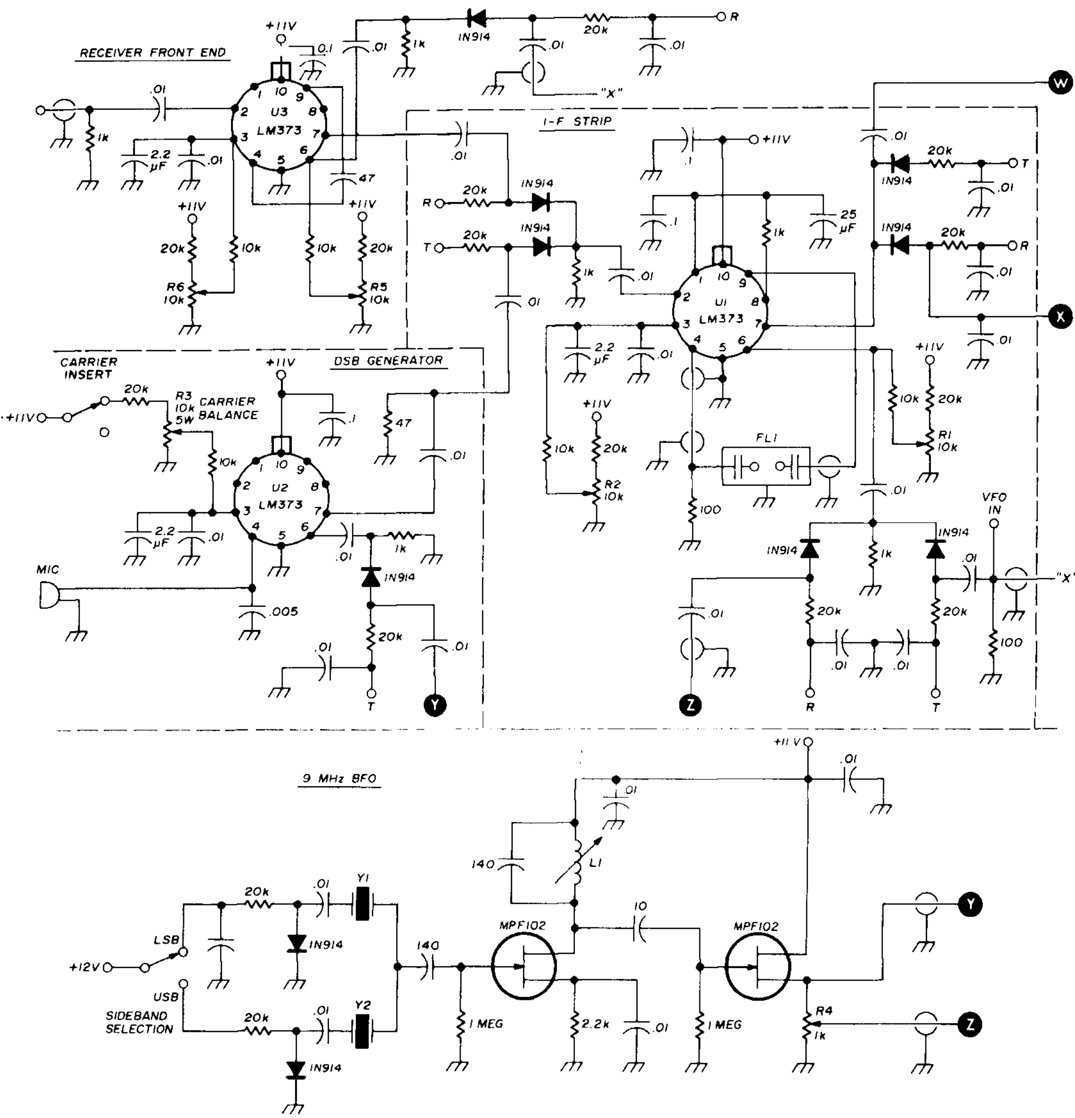


fig. 1. Block diagram of the versatile National Semiconductor LM373 communications IC. This IC can be used in a number of communications circuits, as illustrated in the 14-MHz transceiver circuit shown in fig. 3.

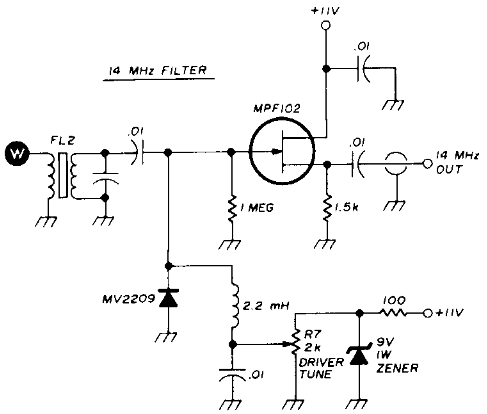


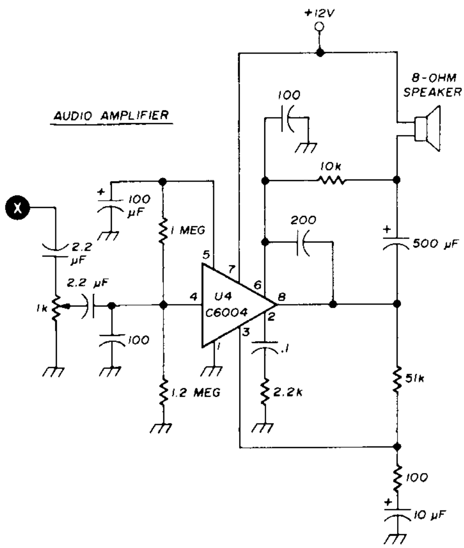
FL1	9-MHz crystal filter (KVG)
FL2	Primary, 3 turns no. 26; tuned winding, 26 turns no. 26 on Amidon T-25 SF toroid core
L1	24 turns no. 26, spacewound to cover entire length of Miller 4300-2 slug-tuned, ceramic coil form
Y1,Y2	Upper and lower sideband crystals for 9-MHz crystal filter (KVG)

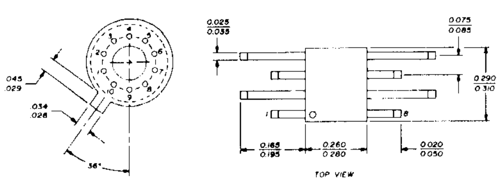
fig. 2. Circuit diagram for the complete 14-MHz CW/ssb transceiver based on the National Semiconductor LM373 communications IC.

stage input. The 9-MHz bfo signal is switched to pin 6 during transmit, and the dsb output is taken at pin 7.

In the transceiver an LM373 is also used as a receiver front end. The input signal is applied to pin 2 of U3, and a 47-pF capacitor couples the output of the agc controlled gain stage (pin 9) to the input of the fixed gain stage (pin 4). The 5-MHz vfo is switched to pin 6 during







receive, and the frontend output is taken from pin 7. All of these LM373 circuits are taken directly from the National Semiconductor application notes.

Audio amplification is provided by the Motorola C6004, U4, which is capable of producing a 1-watt output into an 8-ohm speaker without any transformers. The circuit shown is taken directly from Motorola application notes.

The 14-MHz ssb signal from the i-f strip is separated from the 4-MHz product by the 14-MHz filter, FL2. This resonant circuit is tuned by a Motorola Epicap, MV2209. This allows the driver tune control, a variable resistor, to be remotely mounted on the front panel. An fet drain follower provides a low impedance output from the tuned circuit to the input of the linear amp. The first stage of the linear amplifier is an RCA CA3028A IC in a differential amplifier configuration which drives two 2N2102 emitter followers. These, in turn, drive two 2N2102 transistors in push pull. The final consists of two 2N3553 transistors in class B push pull. The output is transferred by transformer T2 through a coax relay to the 14-MHz input/output filter. This filter serves as both an output filter for the transmitter and an input filter for the receiver.

construction

The transceiver is built on a single piece of copper clad epoxy board with the 5-MHz vfo and linear amplifier mounted separately. The copper is left on to provide the necessary ground plane. Holes are drilled with a no. 60 drill to allow component leads to pass through the board. Ground connections are made directly by soldering to the copper foil. Leads above ground are isolated by reaming away the copper around the holes. The component leads are then hard wired on the non-copper side of the board. All capacitors are rated for 15 volts dc and all resistors are 1/4 watt. All diodes except the Epicap are 1N914 switching diodes. The ssb filter and matching crystals are manufactured by KVG. The only shielding required is around the 9-MHz bfo, around the vfo and between stages of the linear amplifier.

The layout I used is shown in the photograph. An exact layout is not provided here because the actual components I used may be unavailable to you, or may vary somewhat in size. To produce a

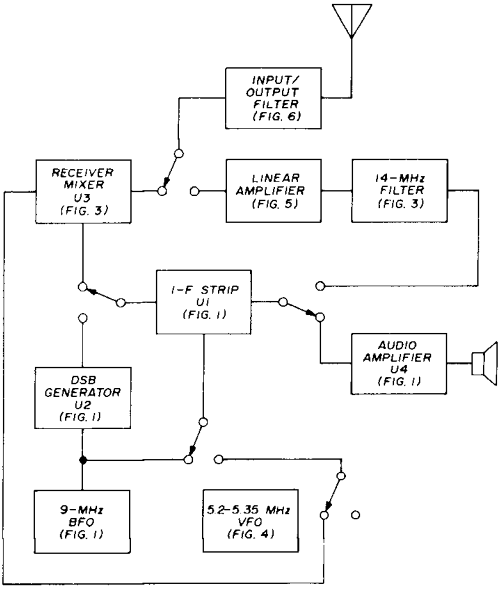


fig. 3. Block diagram of the 14-MHz transceiver in the receive mode. Complete diagram of the transceiver is given in fig. 2.

drilling template, I used ten-squares-per-inch grid paper on which to lay out the components to actual size. It is wise to make all leads as short as possible, and to use shielded lead where indicated in the diagrams. A rigid mechanical frame will insure adequate frequency stability.

tuneup

Since I have a limited amount of test equipment (5-MHz oscilloscope, a Q meter and a crystal calibrator) I had to use a boot-strap method to align the unit. First, check the audio amplifier by touching pin 4 of the C6004 IC with a metal screw driver. The 60-Hz pickup should drive the audio amplifier to good audio output.

Next, the 9-MHz crystal bfo may be brought to life by monitoring its output with an oscilloscope (I am able to see the signal on my 5 MHz scope) or listening to the audio output for a rush of noise while adjusting inductor L1. The i-f strip may be aligned by switching the carrier insert switch to the carrier on position, the unit to transmit, and observing the output of U1 at pin 7 with the scope while adjusting R1 (the signal port null) for a maximum signal output.

Also, it will be necessary to adjust the local-oscillator port null, R2, but first it is necessary to null the carrier from the dsb generator. This is done by switching the carrier insert switch to carrier off and adjusting R3 for a null as observed at pin 7 of U1. Resistor R4 must be adjusted to a threshold point where the null is minimum. Now the i-f local oscillator port may be nulled by adjusting R2 for a null as observed at pin 7 of U1.

Since the signal port adjustment will interact with the local oscillator port null, R1, R2 and R3 must be adjusted in sequence several times to achieve maximum signal output at pin 7 with the carrier on, and for minimum signal with the carrier off.

At this point the i-f strip and the dsb generator are aligned. The receiver front-end may be aligned with or without a signal generator. To align the circuit with

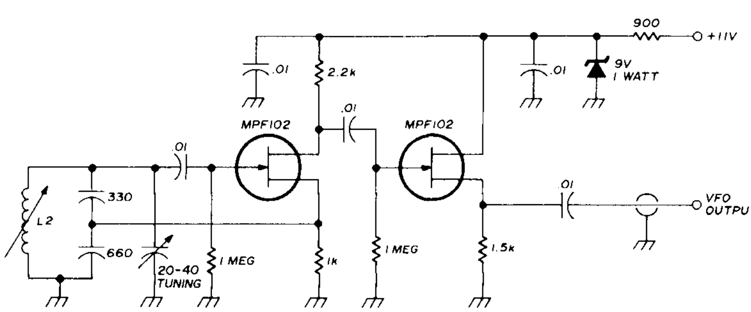
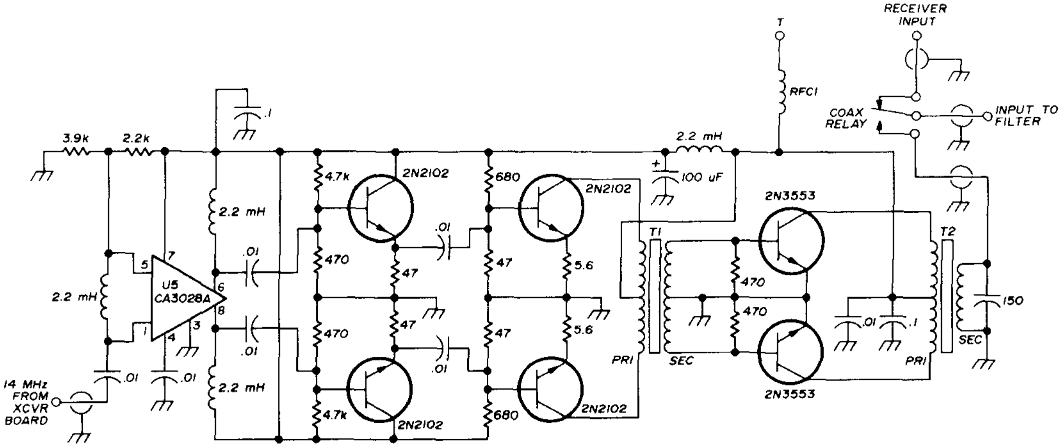


fig. 4. Vfo for the 14-MHz transceiver shown in fig. 3 covers the range from 5.2 to 5.35 MHz. All fixed capacitors are silver-mica units. Inductor L2 is 35 turns no. 26, scramble wound over one-half the length of a Miller 4300-2 ceramic coil form.



RFC1	25 turns no. 16 on Amidon T-50 SF toroid core
T1	Primary, 20 bifilar turns no. 26; secondary, 6 bifilar turns no. 26 on Amidon T-37 SF toroid core
T2	Primary, 18 bifilar turns no. 26; secondary, 26 turns on Amidon T-68 SF toroid core

fig. 5. Solid-state linear amplifier provides up to 5 watts output at 14 MHz.

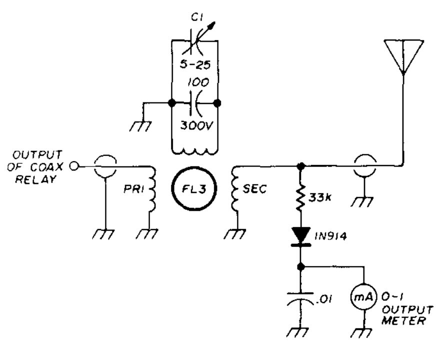
a signal generator, feed the 14-MHz generator output into the antenna terminal and tune L2 until a signal is heard. Next, adjust the signal port null control, R5, for *a maximum signal as observed* at pin 7 of U1. Also, check for peak tuning of the input/output filter by adjusting C1 for maximum signal at pin 7. If a signal generator is not available, attach an antenna during daylight hours and tune L2 until 20-meter signals can be heard, then adjust R5 and C1 as described above. *Now the local oscillator port may* be nulled by removing the antenna or signal generator and adjusting R6 for minimum signal at pin 9 of U1. This concludes the receiver frontend alignment.

The linear is a wideband amplifier and requires no alignment. However, if trouble is experienced, bypass the 14-MHz *filter and* feed the signal from pin 7 of U1 directly into the linear amplifier. This signal will contain a 4-MHz component which can be detected by the scope, thus facilitating normal amplifier troubleshooting procedures. To peak the signals into and out of the linear amplifier, adjust R7, the 14-MHz filter tuning, and C1, the input/output filter tuning, for a maximum signal as indicated on the output meter. This should be done with a 50-ohm antenna or a 50-ohm dummy load attached to the antenna terminal, and the carrier insert switch in the carrier *on* position.

When you have reached this point you can calibrate the transceiver by whatever means available and try it out on the air.

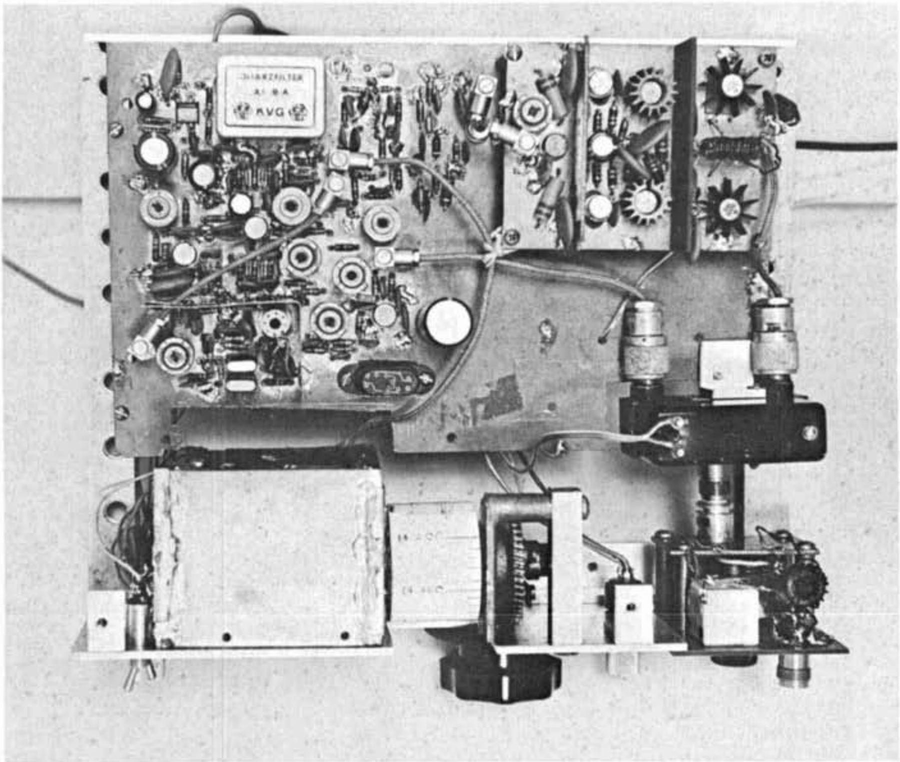
performance

With only 5 watts PEP output a little operator skill is required to communicate.



FL3	Primary, 3 turns no. 16; secondary, 2 turns no. 16; tuned winding, 16 turns no. 16, all on Amidon T-50 SF toroid core

fig. 6. Input/output filter for the 14-MHz CW/ssb transceiver.



Layout of the 14-MHz transceiver, showing the location of the major components.

However, I never cease to be thrilled to announce that I am using QRP after receiving a good signal report. Always get a signal report before telling the other fellow that you are QRP – by some strange phenomenon my signal always goes down after announcing my power level. I must admit that several unsolicited compliments of the audio quality have been made which I attribute to the agc action of U1 during transmit.

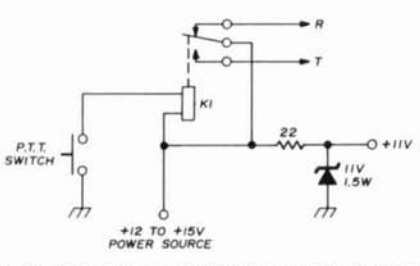


fig. 7. Dc power switching for receive/transmit control on the 14-MHz transceiver.

refinements

If you want to add 75-meter operation, a 4-MHz filter could be switched into the signal path between U1 and the linear amplifier, and a 4-MHz input/output filter switched in between the coax relay and the antenna. Also, a trimmer capacitor would have to be switched into the vfo tank circuit to obtain proper frequency coverage.

The fet oscillator circuits could be replaced by the National LM375 IC, which is a linear IC designed for oscillator/buffer duties. If more power output is desired, a linear amplifier built around the Motorola 2N6367 or 2N6368 might be considered. These transistors are silicon npn devices designed especially for hf ssb service. Motorola provides very good information on the use of these devices from 4 to 30 MHz.

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More Details? CHECK—OFF Page 110

single-frequency repeaters for vhf fm

A discussion of single-frequency fm repeaters and how they might be put to use on the vhf amateur bands

George Allen, W2FPP, 4059 Bay Park Drive, Liverpool, New York 13088

A repeater, by definition, is a device used to automatically relay radio, television, telephone or telegraph signals and is used to extend communications over a range not possible with direct communications. The type of repeaters which most people are familiar with are radio or television repeaters which receive on one frequency and transmit on another (**fig. 1A**). This type of repeater is very common and is the type of repeater which most of the electronics industry has been devoting its time to. However, there is another type of repeater about which very little has been written. This is the single-frequency repeater – a repeater which transmits and receives on the same frequency. This article presents a discussion of the single-frequency repeater.

single-frequency repeater

The single-frequency repeater is broken down into two classes – the delayed type and the simultaneous type. The *delayed* repeater (**fig. 2**) can operate either on a single channel or use separate frequencies for input and output. This type of repeater records communications on the input channel and re-transmits the recorded communications on the output frequency at a later time. The received signal is not transmitted simultaneously with reception but is *delayed* for later transmission. This type of repeater is useful for data or teletype transmission but is not suitable for voice operation where the receiving station is expected to make an immediate reply.

The *simultaneous* single-frequency repeater uses the same frequency for input and output and appears to transmit the received signal during the period of reception. Comparatively little work has been done on this type of repeater in recent years, and a review of the *Applied Science & Technology Index* back through 1968 shows no references to papers or articles devoted to this type of repeater. Although I have heard rumors that both General Electric and Motorola have built and operated single-frequency repeaters, I haven't been able to find any details.

The simultaneous single-frequency repeater can be further broken down into two sub classes — continuous and time division. The continuous single-frequency repeater (**fig. 3**) transmits and receives continuously on the same frequency. To make this type of repeater work, the input and output must have separate

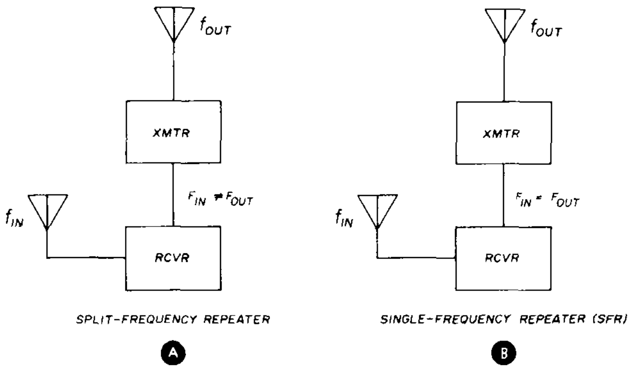


fig. 1. Two basic types of repeaters. The conventional split-frequency repeater shown in (A) transmits on one frequency and receives on another. The single-frequency repeater in (B) transmits and receives on the same frequency.

antennas which are completely isolated. Total isolation between input and output must be achieved so that any signal from the transmitter must be below the receiver noise level. From a practical standpoint in the vhf region this is next to impossible to achieve using a single repeater site. Even if this degree of isolation could be achieved, other problems would make this type of repeater impractical. For one thing, transmitted signals would be reflected back from the geographical terrain and any objects in the vicinity of the repeater, so it would be nearly impossible to keep the transmitted signal from getting back into the receiver.

A more practical approach would be to use a time-division single-frequency repeater as shown in **fig. 4.** This type of repeater is similar to the delayed type in that the input is delayed slightly before transmission. The transmitter and receiver operate alternately, but at such a fast rate that they appear to be operating continuously and simultaneously.

Fig. 5A shows a simple plot of the operating period for the time-division SFR. Note that in a given time interval a portion is used for receiving and a portion is used for transmitting. During the receive interval the received information is saved, either by using a delay line or some type of recording device, such as a tape

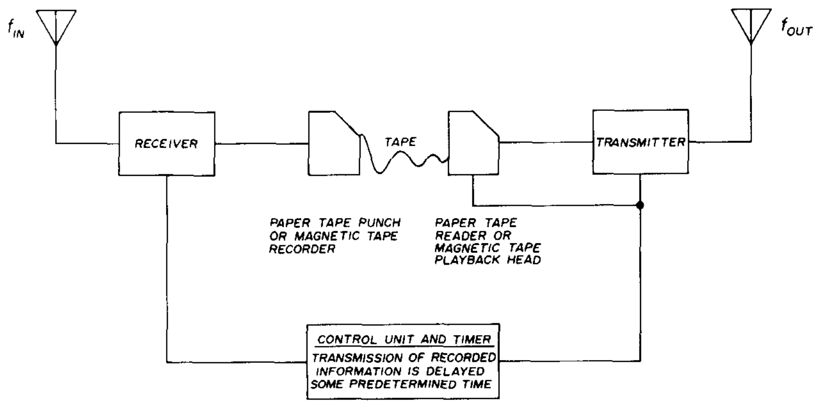


fig. 2. Basic layout of the delayed single-frequency repeater. Input and output frequencies are the same.

recorder. During the transmit interval, the recorded or delayed information is retransmitted.

At this point several problems are introduced into the time-division SFR. First, receivers and transmitters do not turn on and off simultaneously. There is always turn-on and turn-off lag. Thus, after the receiving interval is finished it is necessary to wait for some time period t_{RS} to give the receiver a chance to turn off. This time interval will depend on the type of the receiver and the muting and blanking circuits involved. In regard to the transmitter, it is necessary to wait some time interval t_{TS} before turning on the receiver to prevent the transmitter output from feeding back into the receiver. This is shown in **fig. 5B.**

Another problem arises due to the fact that the receiver is not receiving continuously. There is a significant time period for which information is not received, but lost. This is a sampling mode of operation so the signal to be repeated is not repeated in its entirety, but only samples of the communication are repeated. This problem of lost information can

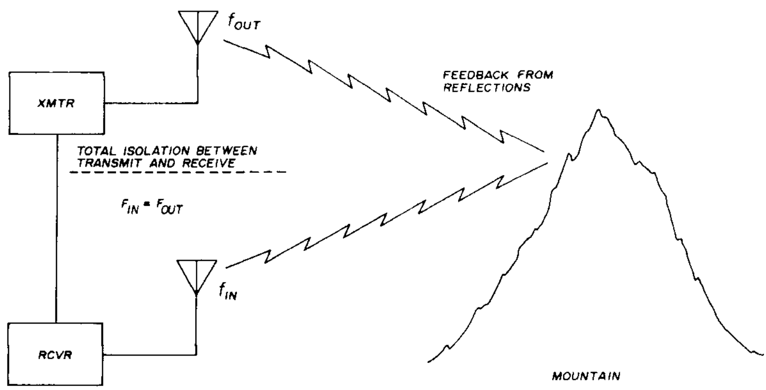


fig. 3. Continuous single-frequency repeater is not practical because any reflections from the transmitter are fed back into the receiver.

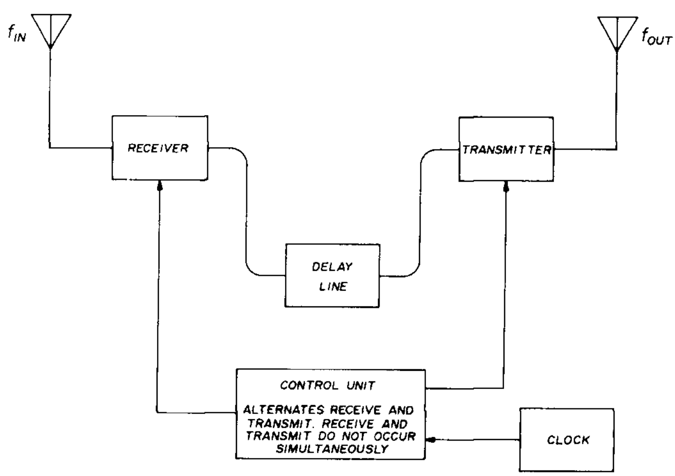


fig. 4. Time-division single-frequency repeater alternates transmission and reception. The repeater appears to be on continuously, but is not.

be solved by choosing the proper sampling interval.

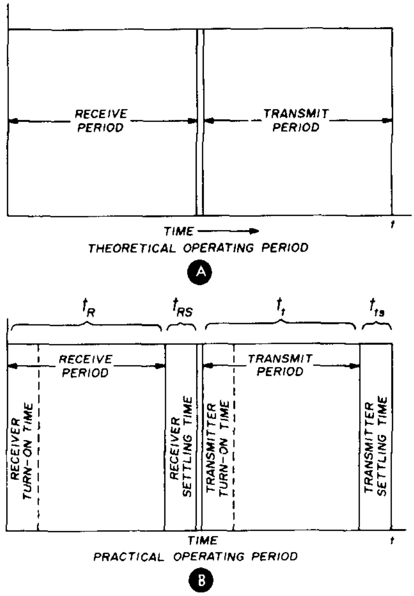
The telephone companies have been using similar sampling techniques for overseas cables, so it is possible to select a sampling interval which will provide minimum loss of intelligibility. This interval, which is rapid enough to retain intelligibility, must be slow enough so that the receive-transmit times are greater than the settling times. If the settling time becomes too large as compared to receive-transmit time, then too much information will be lost and intelligibility will be impared.

Another factor in the choice of time interval is the availability of delay lines. The delay line or recording media must delay the input for a period equal to the receive time plus the receive settling time.

mobile considerations

When attempting to receive a transmission from a time-division SFR a couple of things will be evident. For one, as shown in **fig. 6**, a noticeable buzz or noise could be apparent at the mobile receiver due to the fact that the transmitter is not transmitting continuously but is being on-off modulated in a pulsed mode. This pulsed modulation will appear as a buzz at the mobile receiver and may possibly override the information being repeated. Some means must be used to eliminate this undesired noise. One possible solution would be to use standard noise blanking techniques. Perhaps a more practical approach would be to use a type of synchronized, muted detector. This type of detector would be similar to the detector used to detect synchronous data transmissions. It would have a time interval sensing circuit that would use the first few time intervals to determine the length of interval that the repeater is using. After this interval has been determined, the detector would anticipate future time intervals and in turn, mute the received audio during non-transmit periods. This would eliminate the received pulse noise.

In regard to the mobile transmitter, it appears that no changes would have to be made. In cases where the received mobile signal is stronger from the other mobile than from the repeater, the synchronized, muting detector would capture the mobile signal and operation would be as if there were no repeater. It might be



t (OPERATING PERIOD) $= t_R + t_{RS} + t_t + t_{ts}$

TIME INFORMATION LOST $= t_{RS} + t_t + t_{ts}$

t_R = RECEIVE SAMPLING PERIOD

fig. 5. Operating period of the time-division single-frequency repeater.

mentioned that with this type of system, de-sensing would not occur when mobiles that were using the repeater were close to each other.

While all of the above considerations may appear to present a horrendous picture of impracticality for this type of repeater, you must recognize that each of the concepts are practical from a state-of-the-art standpoint. The only concept not proven is the sum total of all the techniques working together.

experimental single-frequency repeaters

A single-frequency repeater can be

built for experimental purposes provided that some assumptions are made. These preclude any exotic circuitry and emphasize simple compromise approaches. The first assumption is that a short timing cycle would present difficulty due to the complexity of high-speed switching circuitry. Thus, a long timing cycle must be used. Second, delay lines for storage of information from receive to transmit may not be available, so simple means such as tape loop delay lines must be used. Third, commercially available receivers and transmitters must be used to simplify repeater construction and permit use of existing transmit/receive circuitry. Finally, the information to be repeated must be slow in nature so that a signal loss of seconds would not result in a loss of information. It is also assumed that due to the long operating period, noise in the mobile receiver would not be a factor. Thus, this experimental repeater would be used basically to repeat frequency modulated telemetry or on-off tone signals of long duration.

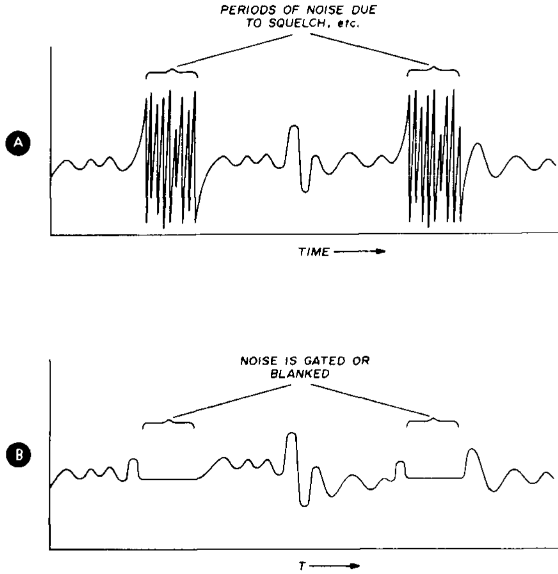


fig. 6. With a time-division single-frequency repeater the operator of a base or mobile station would hear a buzzing sound due to the periods between transmission as shown in (A). This could be solved with noise blanking or gating techniques as discussed in the text (B).

practical single-frequency repeaters

The single-frequency repeater shown in **fig. 8** is designed to repeat tone-frequency modulated telemetry. The basic unit consists of a transceiver such as a Standard 826MA, a clock which generates 0.1-second pulses, a slow speed tape recorder with separate playback head, and a simple control unit. The control unit is set up to count pulses and control the repeater in a manner such that the receiver receives for 9 pulses, waits for a 1 pulse settling period, the transmitter transmits for 9 pulses and waits for a 1 pulse settling period. The record head is connected to the receiver and the playback head is connected to the transmitter. The recorder is operated at the slowest possible speed. Note that it may be necessary to move the playback head or provide another playback head at a distance far enough from the record head to provide an approximate one-half second delay.

The single-frequency repeater shown

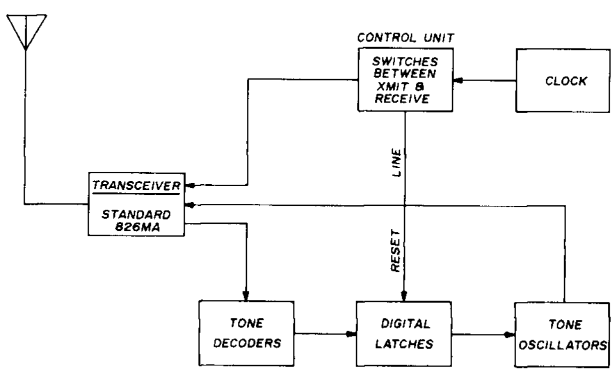


fig. 7. This practical single-frequency repeater is similar to the system shown in fig. 8 except that the tape recorder is replaced by a series of tone filters, tone oscillators and digital latches (see text).

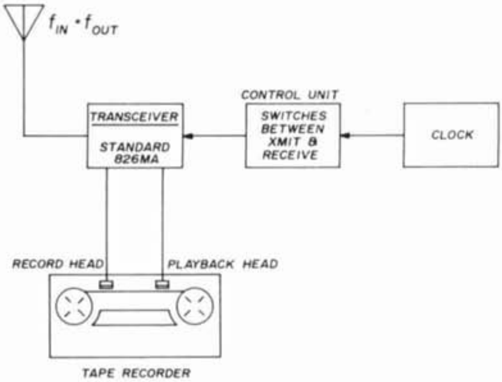


fig. 8. Practical single-frequency repeater. The time delay between record and playback is equal to one-half the operating period.

in **fig. 7** is the same as SFR shown in **fig. 8** except that it is designed to repeat on-off signaling. The tape recorder delay line is replaced by a series of tone filters, tone oscillators and digital latches. The tone filters determine which tones have been received and set the appropriate digital latches. When transmit time comes, the digital latches cause the proper tones to be re-transmitted. At the end of the transmit period, all latches are reset and the system is ready to receive.

Note that any transceiver may be used in either of the single-frequency repeaters. However, the Standard 826MA was chosen because of its commercial quality and reliability.

summary

The purpose of this article is to present a simple discussion of the single-frequency repeater and to stimulate discussion and experimentation on the subject. It is hoped that enough ideas have been presented to kindle an interest in other experimenters.

In regard to future work, I am presently working on the two practical SFRs as discussed in this article. I have filed a license application with the FCC for this type of repeater and will start on-the-air tests as soon as the license arrives. I expect to report the results of these tests within the next year.

ham radio

open-wire
impedance-matching baluns

How to design and use transmission line sections for impedance-matching baluns

R. W. Johnson, W6MUR, Post Box 803, Ben Lomond, California 95005

Greater emphasis on the low-frequency bands as the sunspot cycle decreases makes it desirable to more carefully consider the use of wire antennas. In two previous articles on this general subject I have described a two-band matching system using open-wire lines[1] and a means of using lumped capacitances instead of open stubs in a double-stub matching system.[2] In this article I will discuss two convenient wire-line baluns that will also serve as impedance transformers.

Baluns made of coaxial cable, or toroidal transformers, are limited in their impedance transforming capability because you cannot conveniently change the Z_o of the coax or the turns on the transformer. The balun must be used in addition to the matching system. It is relatively easy, however, to make a balun using open-wire lines that will give a much wider choice of matching ratio, and thus combine the matching system and the balun into one unit to do both jobs at once.

In working with transmission lines it is helpful to use the pi equivalent circuit which is given in **fig. 1.** While this may not be physically realizable using lumped constants, except at one frequency, it is nevertheless a valid mathematical model which helps us to more easily visualize what is going on with transmission lines than if we try to work exclusively with the transmission line equations or graphical devices such as the Smith Chart. **Fig. 1B** shows the true balanced configuration; if you are unconcerned with balance to ground, simply make the top series reactance $jZ_o \text{Sin}\ \theta$ and eliminate the bottom one, as in **fig. 1C.**

Looking at **fig. 1** you can see that a transmission line without any connected reactances is always symmetrical; that is, the shunt arms are equal to each other. There is only one symmetrical pi network that will exhibit transformer action between two pure resistances, and that is the one in which each reactance is equal to $\sqrt{R_{in} R}$, and where the series and shunt reactances are of opposite sign. Thus, in **fig. 1** you must have $\sin\theta = 1/(\tan \theta/2)$. With the aid of trigonometric identities it is easy to show that this requires that θ

be some odd multiple of 90°; e.g., a quarterwave line. The transformation is simply $R_{in} = Z_o^2/R$ for this special case.

In other cases, transformer action between two pure resistances demands that there be some added reactance to one or the other of the shunt arms, so that the network becomes asymmetrical. This is why a stub or lumped reactance of some sort must be added to the general line to achieve a match, although there have been arrangements using an off-resonant antenna as the added reactance on the load side.

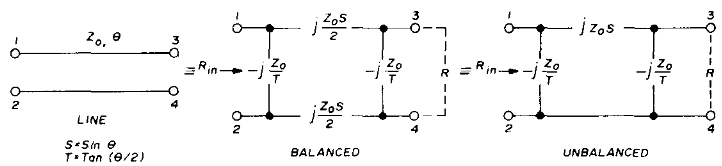


fig. 1. Pi equivalents of the transmission line.

lattice networks

It is also desirable to consider the *lattice* network as an impedance transformer. The lattice is shown in **fig. 2** in both conventional and bridge forms. Again, for the symmetrical network there is only one configuration where transformer action takes place between two pure resistances, that being where the shunt and series arms of the lattice network are pure reactances, equal in magnitude and opposite in phase, $X_1 = -X_2 = X$ in **fig. 2**. The transformation again is simply $R_{in} = X^2/R$ where X is the reactance of each of the lattice arms.

The important thing about the lattice network is that with this particular configuration, any one terminal can be grounded without disturbing the balance to ground. Thus, a coaxial cable can be attached to terminals 1 and 2 and a balanced antenna to terminals 3 and 4; conversely, a grounded antenna may be connected to terminals 3 and 4 and an open-wire line connected to terminals 1 and 2. Note also that with the load disconnected, the input resistance is zero since the bridge is essentially two series-resonant circuits in parallel. This facilitates tuning the network in cases where the reactances are made adjustable.

Fig. 3 shows a lattice network made up of transmission line sections. A shorted line is shown for the inductive arm, and an open line for the capacitive arms, but this is not the only configuration that can be used; it is merely the one which results in the least total amount of line. Its bandwidth characteristics are also fairly good. The reactance of the shorted line is jZ_oT and that of the open line is $-jZ_o(1/T_1)$, where T and T_1 are the tangents of the angles θ and θ_1, respectively. For the symmetrical lattice matching section it is required that $X = \sqrt{R_{in}R} = Z_oT$, and also that $TT_1 = 1$, from which it can be shown that the total line length $\theta + \theta_1$ must be 90°, or a quarter wavelength. The tap point is given by

$$\tan\theta = \frac{\sqrt{R_{in}R}}{Z_o} \quad \text{and} \quad \theta_1 = 90 - \theta$$

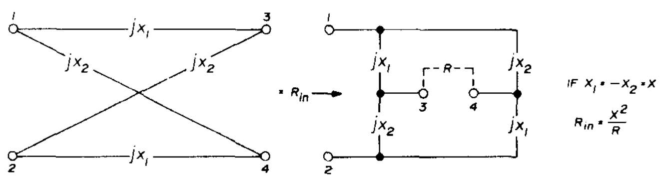


fig. 2. Reactive lattice network. In this configuration any one terminal can be grounded without disturbing the balance to ground. The transmission-line balun shown in fig. 3 is based on this equivalent circuit.

For the greatest bandwidth, tan θ must be small rather than large. That is, the tap point must be near the shorted end of the total line in each case. This dictates that Z_o must be as large as convenient. In the special case where $R_{in} = Z_o$, where the same line impedance is used for matching as in the transmission line to be matched, tan $\theta = \sqrt{R/Z_o}$.

The curves of **fig. 4** give the electrical length of the tap point for some typical Z_o values and antenna resistances, assuming you want to match to 50-ohm coax. The *physical* length, of course, depends on the frequency and v_p, the velocity of propagation on the line (typically 0.82 for twin lead, 0.66 for coax, and 0.96 or so for open-wire feeders). In feet, it is 2.73 $(v_p/f)\theta$ where f is in MHz and θ is in degrees.

Fig. 5 shows one configuration as connected to and suspended from a balanced antenna. For a center-fed full-wave antenna of R = 4000 ohms and 300-ohm twin-lead for the balun, $\theta = 56.15°$. The tap point is thus at 62.4% of the total length of the quarter-wave twin lead, measured from the shorted end on each line.

re-entrant crossed line

The re-entrant crossed line is another form of transmission line impedance-matching balun that may be convenient in some cases, especially where you may want to feed a grounded vertical array using open-wire line. In this case the total length of line required turns out to be a full wavelength. The line and its equiva-

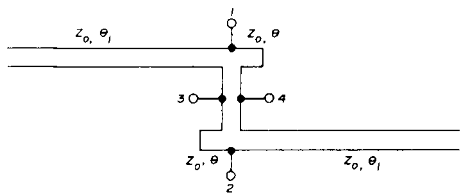


fig. 3. Open-wire transmission-line balun using the lattice network. Antenna is connected to terminals 1 and 2; transmission line is connected to terminals 3 and 4. Tap point for terminals 3 and 4 for matching 50-ohm coaxial cable is given in fig.. 4.

lent circuit is shown in **fig. 6**; note the *crossed* connections.

It turns out that for the case of interest, namely, a symmetrical lattice with arms equal and of opposite sign, the impedances across terminals 1-2 and 3-4 in the equivalent circuit of **fig. 6** must be anti-resonant. That is, $T = -T_1$, where in this case the Ts are tangents of *half* the angle. You are left with the lattice itself in **fig. 6**, and for $S = -S_1$ simultaneously with $T = -T_1$ it is necessary that $\theta = 360 - \theta_1$. The result is

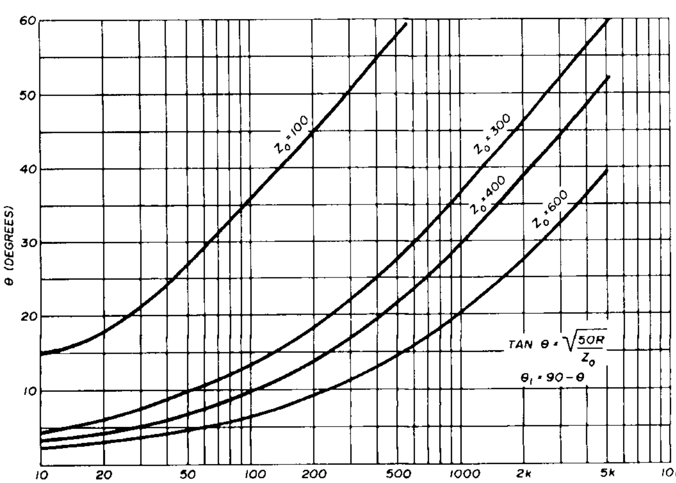


fig. 4. Balun tap point for matching 50-ohm coaxial cable to the open-wire balun shown in fig. 3.

$$\sin\theta = \frac{2\sqrt{R_{in}R}}{Z_o}$$

$$\theta_1 = 360 - \theta \text{ in degrees}$$

For this case Z_o must be greater than $\sqrt{R_{in}R}$ for a real result since $\sin\theta$ cannot be larger than unity. This lattice can also be used as a balun — any one terminal of the network can be grounded without disturbing the balance to ground.

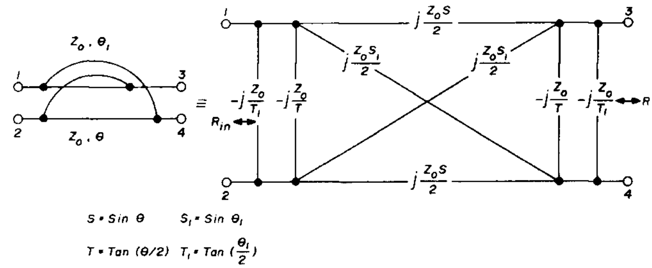


fig. 6. Crossed re-entrant line impedance-matching balun discussed in the text.

If it is assumed that you are matching to 50-ohm coaxial line, then for various characteristic impedance values for the re-entrant balun, you can match load resistances, R, up to the values shown below:

Z_o	R_{max}
300	450
460	1058
600	1800
800	3200
1000	5000

The maximum bandwidth will be achieved when the lattice arm reactance variation with frequency is the least. Since each arm is proportional to $\sin\theta$, it follows that when the variation of $\sin\theta$ is the least, broadest bandwidth will be attained. This occurs when $\sin\theta$ is near unity, which means you ought to choose a value for Z_o somewhere near (but always greater than) $2\sqrt{R_{in}R}$.

For $Z_o = R_{in}$, $\sin\theta = 2\sqrt{R/Z_o}$ and Zo can be no smaller than 4R. This matching balun is not too well suited, therefore, to very high impedance antennas.

For an antenna resistance of, say, 400 ohms, to match a 50-ohm coax using 300-ohm twin lead for the balun requires $\theta = 70.5^\circ$ and $\theta_1 = 289.5^\circ$. The equivalent lattice arms are $(300 \sin 70.5^\circ)/2 = 141.4$ ohms and $(300 \sin 289.5^\circ)/2 = -141.4$ ohms. Transformation is $(141.4^2)/400 = 50$ ohms. The reactances across the input and output terminals are (300)/(tan 70.5/2) = 424.3 ohms and (300)/(tan 289.5/2) = - 424.3 ohms, and are in anti-resonance.

As a final example, suppose you wanted to feed a grounded vertical antenna having a base resistance of 36 ohms, using 300-ohm twin-lead, and also wanted to use twin-lead for the balun matching section. $\theta = 43.85^\circ$ and $\theta_1 = 316.15^\circ$ for this case, so for 7.2-MHz operation you would need 13.64- and 98.32-feet of twin-lead for the two lines (remember to cross the connection).

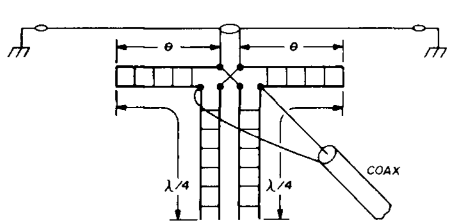


fig. 5. Typical open-wire impedance-matching balun based on the simplified circuit in fig. 3. In this case the balun is suspended from the center of a balanced antenna.

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2. R. W. Johnson, W6MUR, "Transmission Line Matching Using Two Fixed Capacitors," *ham radio*, September, 1973, page 58.

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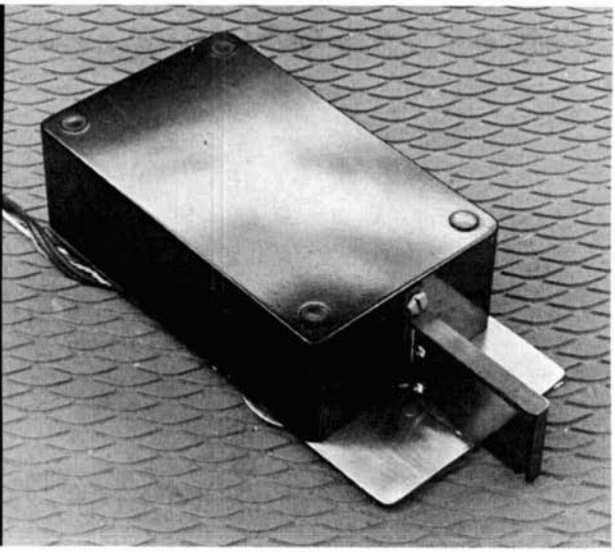


compact electronic keyer package

Complete electronic keyer, ac power supply and squeeze paddle in one compact package

Gene Brizendine, W4ATE, Huntsville, Alabama 35803

As the panarama of amateur equipment design flows on, transceivers and compact solid-state power supplies have rendered operation from your boat, car or camper increasingly attractive. Strangely enough, in this enlightened age of integrated circuits, the electronic keyer has hardly kept pace, and it generally comes in two packages, or one bulky one.

Early labor-saving electronic keyers required sizeable power supplies and heat-radiating tubes and required considerable operating desk space. The keyer described in this article emphasizes compact packaging, rather than electronic features. Briefly, what is probably the most advanced squeeze paddle circuit, including 16 ICs, power supply and monitor, is housed in a package the size of a conventional bug. The circuit is essentially that used in the Pickering keyer.[1]

construction

Printed-circuit board is used extensively because of its rigidity, shielding and ease of working. A sheet of PC board forms the base, which supports all components. An inverted bakelite instrument case, approximately 6x3x2 inches, forms the cover. The cover is shielded with a sheet of heavy aluminum kitchen foil, cemented to all inside surfaces and extending over the edges, where it is clamped against the base in final assembly.

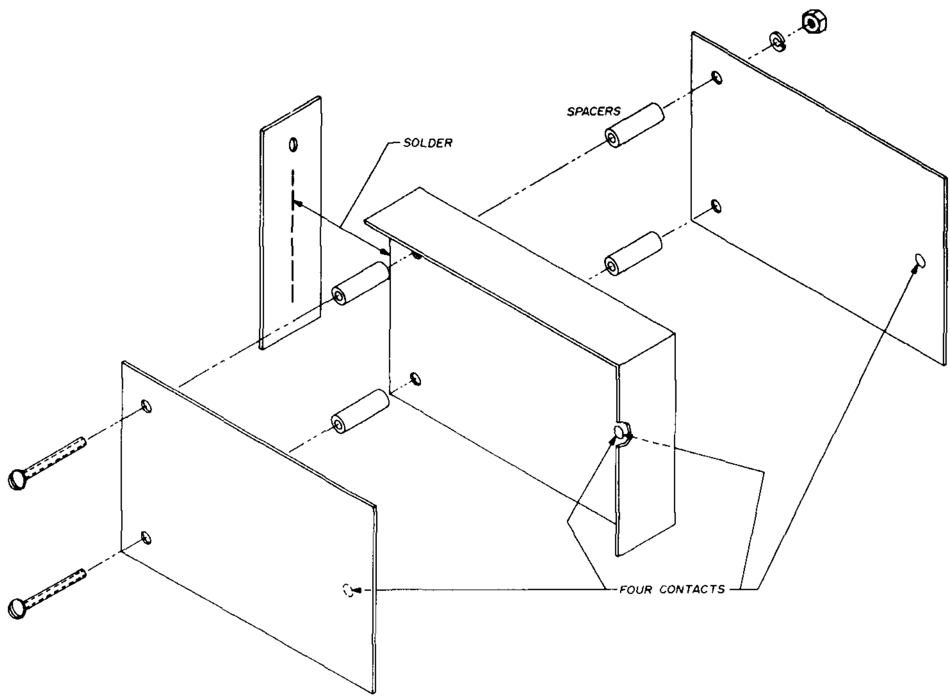
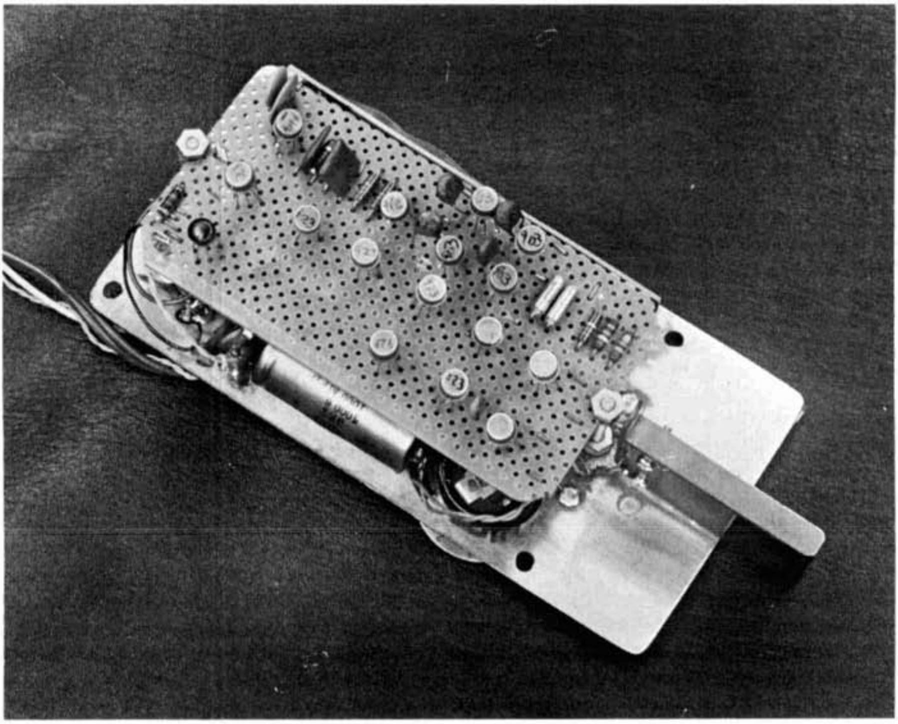


fig. 1. Squeeze paddle details. The middle double-sided PC board carries the stationary contacts. Single-sided PC board is used for the dot and dash paddles.

The RTL logic was wired in when the keyer was designed several years ago. However, an even more compact design is now possible with DIP devices and printed-circuit techniques.

To simplify the IC wiring, perforated board is used, with holes punched in a 45-degree pattern. Eight leads of each IC are inserted into holes forming a square, with pin 8 (+) at the topmost corner and the ground lead, pin 4 (−) at the bottom corner. Thus, IC power wiring is completed by joining all top corners with a positive voltage bus and connecting the bottom corners to ground. This method quickly connects 32 leads and also allows easy identification of the remaining pins.

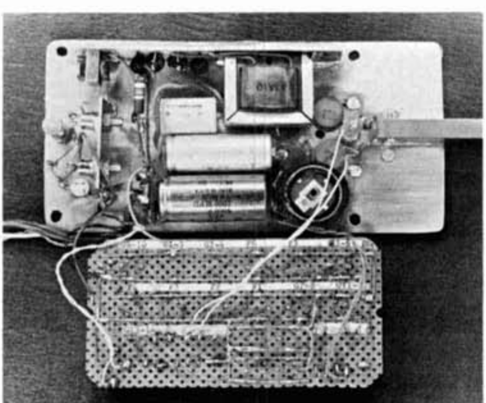
The 3-volt IC power supply components are mounted on vertical strips of PC material which are soldered to the base. A pocket transistor radio output transformer serves as a compact power transformer. After completing the IC board and power supply wiring the board is secured to the base with two studs.



Keyer with the cover removed showing the ICs and discrete components mounted on perforated circuit board. The speed control is located at the bottom left side of the keyer (foreground).

squeeze paddle

The squeeze paddle assembly is also constructed of PC board as shown in **fig. 1.** The double-sided stationary portion is made very rigid by soldering strips along the front, top and rear edges. The bottom edge and the rear strip are soldered to the base. The rear strip is also secured to the cover with one screw.



Interior of the electronic keyer showing the arrangement of the power-supply components and the speed control.

Coin-silver contacts are soldered at the front edge of the boards to provide maximum travel when pressure is applied at mid-paddle. Some experimenting with spacers between paddles may be necessary to satisfy your particular preference. If audible clicking is objectionable, the inner surfaces may be covered with sound-absorbing felt or flocking spray.

The light, compact package shown may be attached on the operating desk with double-sided masking tape. At my station it has given four years of carefree operating pleasure.

reference

1. T. Pickering, W1CFW, "The Micro-Ultimatic," *73,* June, 1966, page 6.

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More Details? CHECK—OFF Page 110

calculating gain vs height of DX antennas

Using a simple graph to predict the gain of your DX antenna if you raise its height above the ground

Gerd H. Schrick, WB8IFM, 4741 Harlou Drive, Dayton, Ohio 45432

It is well known that the higher your antenna is, the better it is for working DX. Although some amateurs talk about optimum antenna height, it is generally accepted that the higher the better. The question is, how much better?

Recently, I replaced an 80-foot tower with a 100-foot tower; both towers were available temporarily so I could make some comparison measurements. With the same antennas mounted on both towers, any gain differences had to be attributed to the difference in height. A gain difference of 2 dB was consistently measured, whether I was working local stations or DX. However, comparison measurements with DX stations were more difficult because of fading and other long-distance communications problems.

theory

Any horizontal antenna works in conjunction with the ground and the mirror-image concept can be applied. The formula for radiated power density, P_d vs the radiation angle, α, is

$$P_d = \sin^2 (h^\circ \sin \alpha) \qquad (1)$$

where h^o is antenna height in degrees of wavelength (1 wavelength = 360 degrees).

The gain difference, in dB, for two different heights, h_1 and h_2, is:

$$\Delta dB = 10 \log \frac{P_{d2}}{P_{d1}} = 20 \log \frac{\sin (h_2 \sin a)}{\sin (h_1 \sin a)}$$

For DX work a very small angle of radiation, a, a so-called grazing angle close to zero degree, can be used. Therefore, sin a is a very small number, much less than 1. The factor $\sin^2$ becomes very small when the actual antenna height is in the order of a few wavelengths. This results in the following approximation:

$$dB_{DX} = 10 \log \frac{h_2^2}{h_1} = 20 \log \frac{h_2}{h_1} \quad (3)$$

This is a surprisingly simple formula. Since the wavelength factor has dropped out, it is no longer necessary to express antenna height in electrical degrees — absolute feet or meters will do. Doubling your present antenna height will give you a maximum gain of one S-unit (6 dB). This is shown in **fig. 1**.

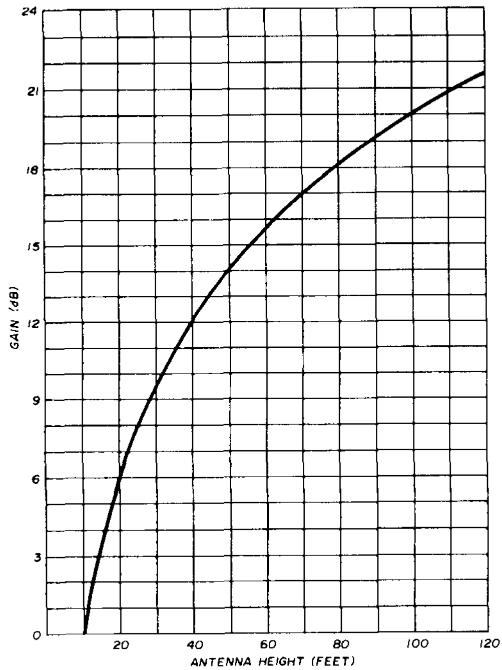

fig. 1. Antenna gain in dB for DX obtained by increasing height of an antenna above ground (angle of radiation less than 3 degrees). Referenced to zero dB gain at 10-feet antenna height.

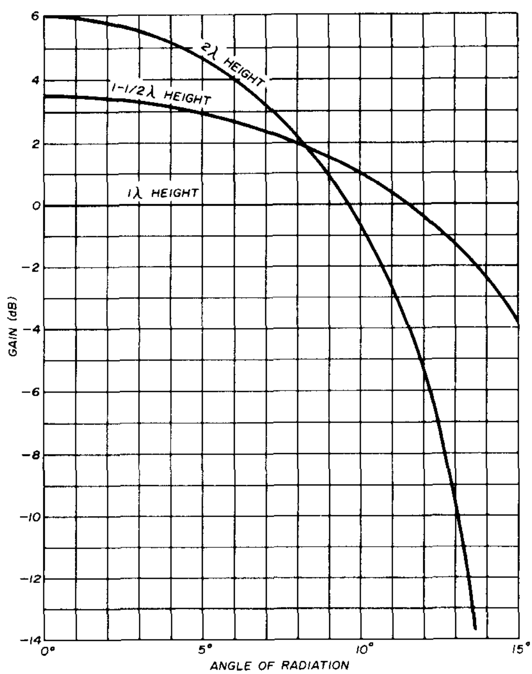

fig. 2. Power density vs angle of radiation for antennas 1½- and 2-wavelengths above ground, as compared to an antenna 1-wavelength above ground.

example

In my case, where I increased my tower height from 80 to 100 feet, the expected power gain is predicted by **eq. 3:**

$$dB = 20 \log \frac{100}{80} = 20 \log 1.25 = 1.9 \text{ dB}$$

This is in very good agreement with the practical measurements described previously.

Fig. 2 shows the difference in power density in dB versus the angle of radiation for horizontal antennas 1-, 1½-, and 2-wavelengths above ground. This graph is based on the use of the exact formula given in **eq. 2.** It can be seen that the approximation in **fig. 1** is quite accurate for radiation angles less than about 3 degrees. Certainly, it is not possible to obtain higher gain than that implied by **fig. 2.**

ham radio

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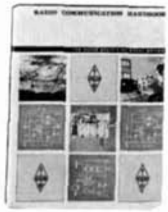

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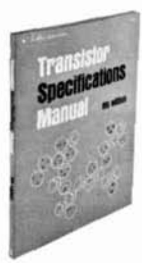

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antenna and control-link calculations for repeater licensing

How to calculate the antenna patterns and control links required for your new FCC repeater license

P. J. Ferrell, W7PUG, 6021 South 119th Street, Seattle, Washington 98178

The new FCC repeater regulations are admittedly a tough act to follow. In particular, the antenna pattern and control-link power justification seem to be difficult to obtain. These items may be calculated directly rather than measured, since either method is acceptable to the FCC. No attempt will be made to derive or justify the formulas employed, but references for each are included wherein they are most adequately treated. Without attempting to ruffle any feathers, I feel that anything tending to raise the amateur's technological level above that of appliance operation and CB can't be all bad.

the antenna pattern

The first problem is to produce an "antenna pattern" and the supporting data to satisfy 97.41 (f)(3,6 and 7) of the new repeater regulations. Any repeater group possessing an antenna range or instrumented helicopter can pass immediately to control-link calculations on page 60. Antenna patterns can be measured of course, but great pains must be taken to guard against reflections, to be certain that only the far-zone pattern is sensed and that the instrumentation is in calibration. All in all, it would require an effort far beyond the meager resources of our group.

A calculated pattern approaches the actual radiation pattern only when the antenna is well clear of obstructions. This condition tends to be met only at vhf and above. Calculating the radiation pattern for an arbitrary current distribution is an extremely messy proposition, but if the antenna can be characterized as a linear array with equal current elements, considerable simplification can be obtained using pattern multiplication. Simply stated, you can separately calculate the patterns due to length, to width and to the individual elements and obtain the overall pattern by multiplying them all together, *provided* the elements and their currents are all the same. Fortunately, the two most popular types of base-station antennas favored for repeater use meet this requirement.

Type 1, the J-pole or vari-loop antenna consists of dipole type elements mounted on a metal pole, separated about one wavelength and fed with a corporate feed insuring a nearly uniform and inphase current distribution.

Type 2 is typified by the Communications Products "Stationmaster" and the Prodelin "Big Stick" antennas. Home-built antennas of this type have been described in the amateur literature.[1] Some amateurs have "improved" them by specifying foam-dielectric coaxial cable, thereby insuring a non-uniform current distribution and a fair sized reduction in antenna performance. These antennas consist of 8 or 9 pieces of alternating series connected coax, each an exact half-wavelength internal electrical length. Special end elements are on the ends, and all elements are spaced about 0.3 wavelength apart and can be considered to have a shortened dipole pattern.

The vertical (E-plane) pattern of the antennas consists of an element part and an array part. We shall assume a *cosine* voltage pattern for the individual element. This is exact for a short dipole, and is a good approximation to a half-wave dipole. The normalized array part is given by:[2]

$$E = \frac{\sin(N\psi/2)}{N \sin(\psi/2)}$$

where $\psi = 2\pi D_e \cos \emptyset$,
N = number of colinear elements
D_e = interelement spacing in wavelengths
$\emptyset$ = elevation angle measured from broadside

The overall normalized E-plane voltage pattern is

$$E = \frac{(\cos \emptyset)\ \sin(N\pi D_e \cos \emptyset)}{N \sin(\pi D_e \cos \emptyset)}$$

You can see that this pattern depends only on the values of N (number of elements) and D_e (spacing between any two elements).

The horizontal (H-plane) pattern of antenna type 2 is omnidirectional and that of antenna type 1 can be adjusted to be omnidirectional. It can also be adjusted to give either a figure-8 or a cardioid pattern. These last two cases are difficult to handle because of the reflections occurring due to the support pole.

A fair approximation may be obtained by replacing the reflecting pole with an *image* antenna fed with an equal current, leading in phase by twice the spacing between the element and the support pole. A typical value for element spacing is about 4 inches or 0.1 wavelength. The normalized voltage pattern in this case is given by:[3]

$$E = \cos\ [2\ \pi\ D_s(\cos\theta - 1)]$$

where D_s = element spacing from the pole in wavelengths
θ = azimuth angle

Note that if we set $D_s = 0$, then $E = 1$ for all azimuth angles, which is just the omnidirectional case.

Having obtained both E- and H-plane normalized voltage patterns, all that remains is to calculate the gain associated with each and multiply them (add dB) to obtain the overall gain.

Gain is calculated by adding up the total power (integrating) at all angles, recalling that power is proportional to voltage squared. Gain is defined as the ratio of maximum to average (isotropic) power. The FCC regulations specify a half-wave reference dipole rather than the more common mythical isotropic antenna, so a factor of 1.64 (2.2 dB) is tossed in to meet that requirement. This is due to a half-wave dipole having a gain of 2.2 dB over an isotropic antenna.

Even this greatly simplified calculation is likely to prove tedious, and so the job was subcontracted to a digital computer. The *Tymshare Superfortran* program used is listed along with sample printouts in the appendix. The two examples are for a 4-element J-pole set for a cardioid pattern, and a 9-element (Stationmaster) antenna. Both E- and H-plane patterns are symmetric about the point of maximum gain (0°) so that only half of each need be printed out.

A polar plot of antenna gain in dB, or field voltage for each pattern can be constructed from the printout. If no reflecting mount is present, answer zero when asked "offset from reflecting mount, wavelengths." In this case, no H-plane pattern is printed out. However, it would be wise to construct a circular plot entitled "azimuth pattern" for submission along with the E-plane pattern.

the control link

I now wish to demonstrate compliance with 97.67(b) of the FCC regulations for a repeater control link. A calculation of this sort is commonly referred to as a "power budget." The handiest procedure is to use decibels exclusively in the magic formula:[4]

$$A = 36.6 + 20\log F + 20\log D + M + L_t - G_t + L_r - G_r$$

where A = worst case loss in dB between transmitter and receiver
F = frequency in MHz
D = separation between antennas in statute miles
M = fading margin in dB (typically 20)
L = transmission line losses in dB
G = antenna gains above an isotropic reference in dB

Clearly, antennas, transmission lines and the propagation path can be all be represented together as an attenuator of A dB.

A certain calculable amount of signal power is required at the receiver input to insure satisfactory link performance, and the necessary transmitter power output is just A dB above that amount. The required receiver input signal power may be calculated from:

$$R = -204 + 10\log B + NF + CNR$$

where R = required input signal power in dBW (dB below 1 watt)
B = receiver i-f bandwidth in Hz
NF = receiver noise figure in dB
CNR = required i-f carrier-to-noise ratio in dB to obtain satisfactory receiver performance. A typical value for an fm receiver is 12 dB

Finally, the required transmitter power is

T = R + A, T in dB relative to 1 watt (dBW)

The following numerical example was submitted to the FCC with the Seattle Repeater Group's application for a repeater license. At 450 MHz, for the given 25-mile path, loss between isotropic antennas is given by 36.6 + 20 Log F + 20 log D.

118 dB	median path loss, Green Mountain to Seattle
2 dB	repeater uhf receiving antenna gain (dipole)
4 dB	repeater site transmission line loss
9 dB	control site antenna gain
5 dB	control site transmission line loss
116 dB	median loss between units
20 dB	fading and diffraction loss margin
136 dB	worst case path loss between transmitter and receiver

-159 dBW	thermal noise power in 32 kHz i-f bandwidth (KT = -204 dBW/Hz)
15 dB	receiver noise figure
15 dB	required SNR at detector input for reliable tone transmission
-129 dBW	signal power required at the uhf control receiver

The necessary control transmitter power is 136 - 129 = +7 dBW. Converting to watts, 7 dBW = 5 watts.

references

1. K. W. Sessions, K6MVH, "Colinear Gain Antenna for VHF/UHF Repeaters," *73,* July, 1971, page 42.
2. John D. Kraus, *Antennas,* McGraw-Hill, New York, 1961, page 78, equation 4-52.
3. op. cit., page 292, equation 11-55.
4. *Reference Data for Radio Engineers,* Howard W. Sams & Co., Inc., Indianapolis, 1968, page 26-19.

ham radio

More Details? CHECK—OFF Page 110

the ham notebook

RTTY line length indicator

When using radio teletype it is my practice to try to type some of my reply while printing the transmission of my contact. To do this I use a model 15 typing unit which prints the incoming signal while I am using the keyboard to punch a tape on my model 14 reperforator. As I cannot see what I am typing or tell when I reach the end of a line with this system, I have designed a system to count the words or spaces in the usual teletype line.

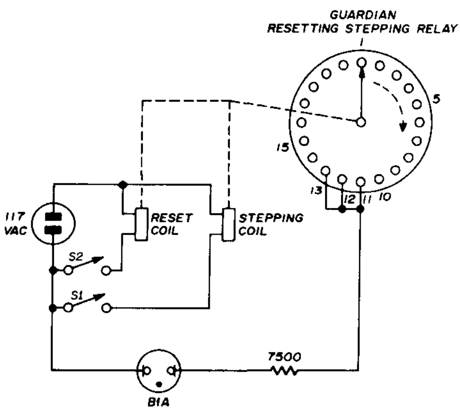


fig. 1. Circuit for the RTTY line length indicator uses a Guardian resetting-type stepping relay. Microswitch S1 is mounted under the space bar. Microswitch S2 is located under the carriage-return key.

The key to counting the number of words in a line is a stepping relay. I use a Guardian resetting type relay. Two microswitches are used to actuate the relay coils. These are mounted under the keyboard of the typing unit and are so positioned that when the space bar is pressed it also strikes the microswitch lever. This switch then closes the circuit to the advancing coil on the stepping relay, advancing it one step.

The other microswitch is positioned under the carriage return key. When this key is pressed this switch actuates the circuit to the relay reset coil. The relay is then reset and is again ready to count the spaces in the next line.

The usual practice is to have ten words per line. I connected my stepping relay to make contact on the 10th, 11th, 12th and 13th steps. When this contact is reached, the alarm circuit is energized. A bell, buzzer, light or other signal can be used. As this relay is a 115-volt ac unit, I use a GE B1A neon bulb in a red panel mounting. This bulb must be used in series with a 7500-ohm resistor. The light is mounted next to the keys and the frame of the typing unit.

After ten spaces or steps the light comes on and stays on for the next three steps. If advanced any further it will go out and you will know you have gone beyond the usual line length. You don't have to worry about over advancing the relay. It goes to the limit and that's all. The stepping relay should be mounted in an insulated box for quiet operation. I also shock mounted mine.

Hal Dressel, W2UVF

ssb filters

Recently, a short item on sideband location was published in *ham radio*, based upon information published in the instruction manuals for the Swan 500 and the Collin S-Line. There was mention of the need to maintain an adequate separation from a band edge. There are several other aspects of filter performance, some of which are easily demonstrated in a receiver or transceiver. By making these tests, a phone operator may have a better understanding of just where his sideband lies, and where he may produce or receive interference.

Without an antenna, or on a dead band, turn on the calibrator and set the dial to zero-beat with the signal. Then, turn the marker to that setting. Now, swing the dial and note the S-meter reading for the maximum signal strength off to one side. This maximum may be adjustable by the preselector tuning control, so that the bandwidth can be noted for different signal levels.

Turn the dial away from zero-beat until the beatnote has dropped to one of several convenient levels, and finally disappears in the noise. Make a note of these dial settings. Then change to the other sideband and repeat the test.

In the case of my Collins S-Line, the normal 2.1-kHz filter (for receiving only, it has also the 1.5 and 3.1 kHz filters) drops down or out about as follows:

Peak signal	LSB	USB
S7	+3.5 kHz	-3.5 kHz
S9 + 10 dB	+4.2 kHz	-4.2 kHz
S9 + 20 dB	+4.4 kHz	-4.4 kHz
S9 + 20 to S3	+3 kHz	-3 kHz

The size of an S-unit must not be taken as 6 dB. It is about 4½-dB on the Collins 75S3A meter at S9, and less than 3dB around S3. The marked range of 20 dB for S3 to S9 is about correct on my meter, indicating an average of 3.3-dB per S-unit. It takes only 10 dB to increase the meter from 40 to 60 dB as marked on the scale, but 20 dB above that.

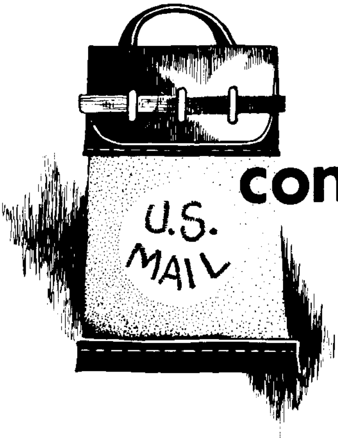
The amount of interference, produced or received, varies with the power and the conditions. If a station talks locally or under good skip conditions with an S9+20 dB signal, he should reduce power as soon as he hears that he is above S9, or suspects it from the strength of the station being worked. Minimum power is required by the Communication Act, Section 324. Otherwise, there may be considerable unnecessary "splatter" due to the filter skirts or equipment problems. In addition, there may be off-frequency emissions greater than those indicated by the filter skirts demonstrated by the above tests. Furthermore, the transmitter can have a somewhat different filter performance than the receiver.

FCC regulations, Section 97.63, says ". . . sideband frequencies . . . shall be confined within the authorized amateur band." It doesn't specify any number of decibels down. This means that the amount of attenuation that is acceptable, depends upon power and propagation conditions, not just -20 dB or -60dB, or any other particular figure. Some standards have been set — BBC Johore, in discussing their harmonic on 14240 kHz, indicated that ITU requires harmonics to be down 60 dB.

The test demonstrates that you must not operate as close to a band edge as previously suggested, even when you have the equipment to comply with the requirement that the frequency of the emissions shall be measured by independent means, and of an accuracy sufficient to assure operation within the band.

The test also shows that interference may be expected from other stations in the range given in the above table, depending upon his filter performance; and that you will produce interference on the *opposite* side by at least the amount shown, plus any additional hash produced above the filter curve. If you hear the calibrator on the "undesired" side of the zero beat, then you may expect to receive some interference there, and produce some on the opposite side when transmitting, because of the lack of pure *single* sideband operation.

Bill Conklin, K6KA



comments

frequency scaler

Dear HR:

Response to the construction article for the simple frequency scaler appearing in *ham radio* for September, 1972, has been most gratifying. That article stirred up the interest of a large number of hams and for many of them it was the impetus for their first-time experience with frequency counters as well as with scalers. To the serious experimenter the article presented no problems. Some difficulties cropped up for the less knowledgeable, however. One of these was that their high-sensitivity, high-impedance counter might indicate twice or even three times the frequency which was fed to the scaler.

Although the solution to this type of problem is really quite simple, a wide variety of schemes have been tried by amateurs in an attempt to solve this problem. One of these schemes is the one which W6MGI described in *ham notebook* of the February, 1973, issue (page 57). For his particular cable and for its particular length, it worked for him. His L-section approach does not provide a general solution, however. The characteristics of his filter will change with every foot of cable he adds or removes (50-ohm cable, for example, has about 28.5 pF per foot).

Any ECL or other very high-speed pulse circuit requires detailed attention to feedback and matching problems. Use of a ground plane which provides a good, low-impedance ground current return path is one essential. A well regulated power supply is another. These can be and are provided within the scaler itself.

The matching problem can best be met outside of the scaler. Experience with and during development of the scalers manufactured by Belmont Spectrum Research clearly demonstrates that for interconnects of more than about 5 inches (some of this length is within the scaler) the matching problem can only be generally successfully met by providing an adequate line termination. This, of course, cannot be accomplished within the scaler. It is therefore standard practice for manufacturers to caution users that an interconnecting cable (scaler to counter) must be terminated in its characteristic impedance if waveform distortion cannot be tolerated. Whether it is called distortion, ripple, ringing or reflections, is immaterial; the steep multiple wave fronts of ECL or other high-speed pulse systems are what trigger the counter and disconcert the casual scaler user. So – the really simple solution to the matching problem is to terminate the line.

One means of providing a proper termination is to connect a resistor, equal to the characteristic impedance of the line you are using, across the input connector inside the counter. Another means, which is preferred because its use is more flexible and does not entail any change to the counter, is to use a "termination adaptor" at the counter end of the line. This method will provide a proper termination when the scaler is used and when unplugged will permit retention of the counter's regular input for use over its normal range. A BNC tee-connector (UG-274/U) at the counter, with the appropriate resistor connected to one arm of the tee, will do very nicely.

These tees are often available through surplus sources at very low cost.

An even better arrangement is to use the 50-ohm termination adaptor made by Tektronix (part no. 001-0049-01). This adaptor, originally intended for use with Tektronix scopes, is highly recommended. It currently sells for $10.

Now, as to overall results. My Belmont Spectrum Research scaler will drive up to 18 feet of either 50- or 100-ohm coaxial cable without substantial waveform degradation when any one of these terminating arrangements are used.

F. Everett Emerson, W6PBC
Belmont Spectrum Research
Belmont, California

passive sideband generator

Dear HR:

Although I have been running my passive sideband generator at the signal levels indicated in the article,* (about 3 volts peak rf and 0.3 volt peak audio at the input ports), the audio level should be several volts. I forgot about the extra loss in the resistive output branch of the audio phase shift/network.

An appropriate audio signal level can be deduced as follows. Suppose the peak rf input is 3 volts. After going through the rf phase-shift networks this is reduced to 2.1 volts. A good guess at the modulator resistance is 1000 to 1500 ohms (500 for the balance pot, 200 or so for the diode and 300 to 800 for the load seen through the output rf transformer).

If we pick 1250 ohms, the rf current through a diode is about 1.7 mA. The modulating signal should be small compared to the carrier, ten percent being a usual limit, so the audio peaks should approach, but stay under, 0.17 mA.

The resistance of the audio phase-shift network output branch, including the modulator impedance, has been adjusted to 3900 ohms. Hence, the audio voltage at an output of the phase shift network should be about 0.66 volt. As noted in Van Heddegem's article, the voltage across the 3-ohm input resistor will then be 1.32 volt, making the total peak input voltage 6.4 volts (about 1.4 watts). This is an upper limit. A third to half of this voltage is adequate while still being ten times what I have been using.

Incidentally, with this higher level of modulation, the carrier balance will be much less critical since a voltage gain of ten or so is thus moved from the sideband amplifier to the audio section.

Worthie Doyle, W7CMJ
Port Orchard, Washington

*W. Doyle, W7CMJ, "Phasing-Type SSB Generator," *ham radio,* April, 1973, page 22.

Dear HR:

Cheers to author Doyle for his fine article on the phasing-type ssb generator in the April issue. Some three or four years ago I built up a solid-state phasing exciter (not exactly passive like Doyle's) and I used a tired 6L6 linear amplifier. I worked Albuquerque, New Mexico, from Libertyville, Illinois, and was heard in Guantanamo Bay, Cuba, on 14 MHz.

For checking the Doyle passive audio phase-shift network I recommend use of an oscilloscope having separate X and Y axis inputs. Leave the *carrier input* of Doyle's **fig. 1** disconnected for this test. Disable the sweep of the oscilloscope, and adjust both channels for equal gain. Hook the X input to one side of the phased audio output; the Y input to the opposite side. Talk into the microphone (or use a signal generator) and observe the pattern on the scope from 300 to 3,000 Hz. It should remain a nice circle, changing in diameter with signal level. If not, make adjustments to the circuit until you get the best circle.

Connecting the phase-shift network ground terminal to the oscilloscope chassis ground should cause no change to the geometry. This same method may be used to check the ninety-degree rf phasing as well, but probe lead length and amplifier phase differences inside the oscilloscope can give misleading information.

Paul Schmidt, W9IDP
Libertyville, Illinois

new products

two-meter fm transceiver

A complete new line of American-made 2-meter fm transceivers for the amateur has been introduced by General Aviation Electronics, Inc. (GENAVE). The new offering includes the GTX-10, the GTX-200, and the previously introduced and very popular GTX-2.

The new GTX-10 fm transceiver is a full 10-channel, 10-watt output unit and retails for $199.95. The new GTX-200 features independent selection of 10 transmit and 10 receive frequencies, offers 30 watts nominal output power and retails for $259.95. The well-received GTX-2 provides 10 push-button channels with backlighting for night operation, 30 watts nominal output power and retails for $249.95.

The radios are manufactured in the same U.S. Government inspected facilities where precision aircraft instruments are fabricated, under the same watchful quality control procedures.

Internally, all radios are equipped with netting trimmers for each transmit crystal. All use standard, readily available, American-made semiconductors. High selectivity 8-pole second i-f filters are incorporated in the design of all three units, and rf output stages are vswr protected.

The GTX line is engineered for use with available tone encoders and autopatch service, as well as simultaneous operation on MARS frequencies. Externally, all three radios have multi-position switches which include a low power (one-watt) setting for longtime low-power drain operation, and indexed volume and squelch controls.

Each GTX transceiver comes complete with a quick-disconnect power cable, SO-239 antenna connector, mobile mounting bracket and sturdy ceramic plug-in microphone. A 146.94-MHz communications channel is also included. The remaining plug-in crystals are available at $6.50 each for installation at the factory or by the owner.

The new GTX-10 features superlative cross-mod performance, and is easily cross-wired for duplex crystal operation. The circuit board is laid out so that conversion to 30-watt output can be accomplished quite easily.

The new GTX-200, with independent selection of 10 transmit and 10 receive frequencies, offers 100 possible channel combinations. A switch for lock-in of preselected frequency pairs allows simple one-knob operation when desired. High sensitivity is assured by incorporation of a dual-gate mosfet in the receiver frontend. An external speaker jack is provided on the rear panel.

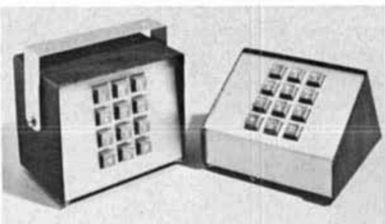
The popular GTX-2 has been refined and updated for superior sensitivity with the same dual-gate mosfet in the receiver frontend as is found in the GTX-200.

The new radios are founded on the technology and know-how derived from Genave's experience as a leading manufacturer of a full line of navigation and

More Details? CHECK—OFF Page 110

communications radiotelephones for the marine industry. For more information, write to General Aviation Electronics, Inc., 4141 Kingman Drive, Indianapolis, Indiana 46227, or use *check-off* on page 110.

touch-tone encoder enclosures

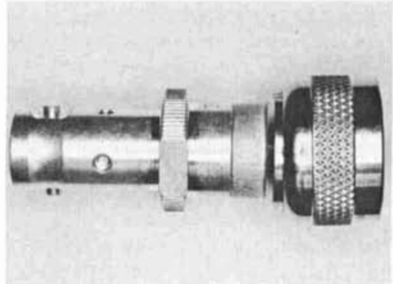


There has been a tremendous increase in the use of Touch-Tone encoders for use in autopatch and other special control systems in fm repeaters. Now two compact enclosures are available for Touch-Tone pads – one for base stations and one for mobile use. The Touch-Tone pad is held properly in position by pre-mounted internal brackets.

The top and sides of the enclosure are covered with walnut-grained vinyl. The satin anodized aluminum face is die punched to accept standard 12-button Touch-Tone pads such as those manufactured by Western Electric, Stromberg-Carlson, ITT-Kellogg, etc. (Automatic Electric Touch-Tone pads will not fit because they use non-standard spacing between the buttons.)

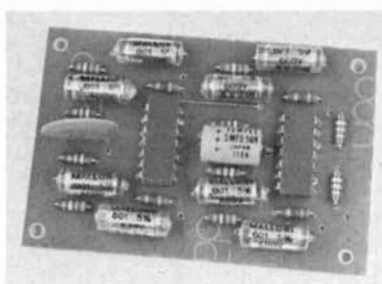
The mobile mount "M" Touch-Tone enclosure has an anodized pivoting gimbal bracket which provides multi-position mounting under the instrument panel, as well as top-of-equipment mounting on four rubber feet. The base station "B" enclosure holds the Touch-Tone pad at a convenient 30-degree angle. There is ample room inside both enclosures for transmitter keying circuitry. Either model is available for $6.25 including postage from the Detroit Area Repeater Team, Post Office Box 201, Clawson, Michigan 48017. For more information, use *check-off* on page 110.

connector adapters



A new series of adapters is now available which will interconnect any combination of cables and connectors having BNC, RCA phono, microphone (Amphenol 5/8-27 type) or type-F (CATV type) terminations. These adapters are intended for use in test equipment, and in audio, video and rf applications. The adapters are priced at $4.95 each, postpaid in the United States. Order from Adapters Unlimited, Post Office Box 48822, Los Angeles, California 90048. For more information, use *check-off* on page 110.

cw filter



The new audio CW filter available from MFJ Enterprises offers three degrees of switch-selectable selectivity in the same filter. The three bandwidths are 180, 110 and 80 Hz. In the 80-Hz position, response is down 60 dB one octave away (one-half and twice the center frequency of 750 Hz).

There is no insertion loss when this audio filter is switched into the line. Also, with most narrow-band audio filters ringing can make copying impossible. In the new CWF-2 CW filter ringing is nearly

More Details? CHECK—OFF Page 110

eliminated by the technique of cascading four low-Q stages. This provides very narrow bandwidth and extremely high skirt rejection without audible ringing.

The CWF-2 filter offers very low output impedance and a very high input impedance. This means that, unlike some other filters, no impedance matching is required for optimum performance. Loads greater than 500 ohms produce no distortion and loads less than 500 ohms, such as an 8-ohm speaker, produce some distortion which does not affect copying.s,

To use the CWF-2 you simply plug it into the phone jack of your receiver or connect it to the speaker terminals. It can also be installed between audio stages in your receiver. The circuit consists of four IC operation amplifiers in an active-filter design which eliminates all inductors and reduces the unit's size to a mere 2x3-inch printed-circuit board.

The CWF-2 CW audio filter is available in kit form for $9.95, or completely wired, tested and guaranteed for $12.95. For more information, write to MFJ Enterprises, Post Box 494, Mississippi State, Mississippi 39762, or use *check-off* on page 110.

two-meter collinear mobile antennas

Extremely low radiation angle, 5.2-dB gain over a 1/4-wave ground plane, low swr and wide bandwidth are features of Hustler CG-Series Super Gain two-meter collinear mobile antennas from New-Tronics Corporation.

Model CGT-144 is complete system including collinear antenna with stainless steel radiating sections, 180° swivel ball, heavy duty trunk lip mount for easy "no holes" installation on side or edge of trunk lip, and 17-foot RG-58/U coax with factory-attached connectors. Power rated at 200 watts fm, the completely operational CGT-144 has a swr of 1.1:1 (typical) at resonance and a swr within 1.5:1 over its 6-MHz bandwidth of 143-149 MHz. Overall length is 86 inches.

The Hustler Model CG-144 consists of the 84-inch collinear antenna with 3/8-24 threaded base to fit standard mobile ball mounts. It has the same electrical characteristics as the Model CGT-144. For complete specifications contact New-Tronics Corporation, 15800 Commerce Park Drive, Brook Park, Ohio 44142, or use *check-off* on page 110.

base command

The PACE Communications division of Pathcom, Inc. today announced a new special purpose instrument called *Base Command* designed for sophisticated control of base station operation on the amateur two-meter band. Designed specifically to keep constant surveillance on the performance of your base station, the P5407 *Base Command* is placed in the transmission line between your transceiver and antenna to monitor and control the functions of your transmitter.

Antenna installation efficiency is measured by checking the standing-wave ratio. The transmitter power is measured on one of three scales; the 5-watt and 50-watt levels terminate in an internal dummy load. The 500-watt scale samples power while it's going through the line to the antenna.

The modulation capability of the transmitter is measured on the reference meter. The audio quality of your transmitted signal may be monitored continuously. The unit also provides a visual "on the air" indication. Performance of your hand-held or mobile sets can also be checked with the built-in field strength meter. A television interference filter is also built into the P5407.

More Details? CHECK–OFF Page 110

This is where the name, *Base Command,* was designated since this versatile instrument not only monitors but controls the base-station transmitter. For more details, write to PACE Communications, Box 306, Harbor City, California 90710 or use *check-off* on page 110.

phase-locked loop handbook

The complete story of the phase-locked loop is told in a free 76-page paperback entitled *Signetics Linear – Phase-Locked Loops Applications Book,* recently published by Signetics Corporation, a subsidiary of Corning Glass Works. The book is a companion to the larger *Linear Specifications Handbook* which is also available, although it must be requested separately, according to Jack Mattis, manager of consumer product marketing in the company's linear department.

Phase-locked loops are a new class of monolithic integrated circuits developed by the Signetics research and development department in 1969 and marketed by the firm during the following year. They are based on frequency feedback technology which dates back 40 years. A phase-locked loop is basically an electronic servo loop consisting of a phase detector, a low-pass filter and a voltage-controlled oscillator. The controlled oscillator phase enables the PLL to lock or synchronize with an incoming signal.

In addition to the dash of history given in the book's introduction, other sections provide a short glossary and descriptions of the phase-locked loop principle and PLL "building blocks." Major sections include explanations of general loop setup and tradeoffs, PLL measurement techniques, monolithic phase-locked loops, expanding loop capability, and specific applications.

Some of the more interesting passages contain information on how the user of the PLL can apply the circuit to his own projects. As a functional building block, the phase-locked loop is suitable

for a wide variety of frequency-related applications. These generally fall into one or more of the following categories: fm demodulation, frequency synthesis, frequency synchronization, signal conditioning, and a-m demodulation. Each category is covered by a section in the book.

A number of construction projects are suggested as a means of proving the feasibility of using the phase-locked loop circuit in specific applications. The book provides information on building an fm i-f amplifier and demodulator, a phase-locked a-m receiver, an i-f stage with agc and a-m/fm detection, a translation loop for precise fm i-f generation for tv, a phase-locked FSK demodulator and many others. For a free copy of the handbook, write to Signetics PLL Handbook, Signetics Corporation, 811 East Arques Avenue, Sunnyvale, California 94086.

integrated-circuit fm detector

A unique method of fm detection by a new technique of linear gating is featured in the new Signetics ULN2111 monolithic integrated circuit. This linear device comprises a three-stage limiter and a balanced product detector.

Applications for the ULN2111 device include tv sound channels, fm receivers, automatic frequency control systems and communication receivers. An outstanding feature of the ULN2111 is that only one, simple, low-cost, single-winding coil is required for tuning. Consequently, only one screwdriver adjustment is required to tune a detector circuit which uses the ULN2111.

The frequency range of the ULN2111 extends from 5 kHz to 50 MHz. Outputs of 0.6 V with a total distortion of less than 1% and a limiting threshold voltage of 400- μV rms are typical. Another feature is a voltage gain of 60 dB. When ordered in small quantities, plastic in-line packages are priced at $1.50 each.

For more information, write to Signetics Corporation, 811 East Arques Avenue, Sunnyvale, California 94086.

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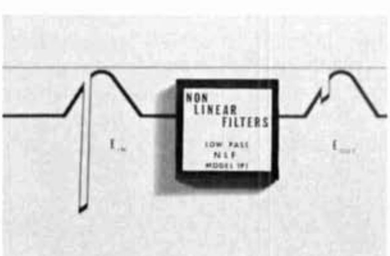
from cb to ham beginner

Whether you are a CBer or an SWL – or anyone else who wants to get started in amateur radio – this book answers your questions. It is an easy-to-read book that explains how to select and tune a communications receiver, listen in on the ham bands, acquire technical knowledge by building and experimenting, learn to send and receive code, build or buy an amateur transmitter, erect an effective antenna, and put an amateur station on the air.

In addition, the author, J.A. Stanley, presents a *fun* way of obtaining your Novice license. Instead of spending countless hours reading dry theory, simple tests are provided that use low-cost, readily available parts. A transmitter project is even included, so you can get on the air as soon as you get your license.

The book is filled with photos and drawings to simplify the subject matter. You need no technical background other than that obtained from operating a CB rig to understand and enjoy this book. It's an invaluable reference source that tells you how to become a Novice-class radio amateur. All the latest FCC Novice rule and frequency changes are included. 144 pages, softbound. $4.25 from Comtec Books, Greenville, New Hampshire 03048.

non-linear filters



Non Linear Filters has introduced a non-linear low-pass filter module that attenuates frequencies above the corner frequency while introducing no phase shift, either above or below the corner frequency. This filter is significantly more effective than linear filters for eliminating noise spikes. For example, high amplitude noise spikes occurring on fm carriers are highly attenuated before limiting. Since the non-linear low-pass filter introduces no phase delay distortion, it is useful where non-sinusoidal waveshapes are encountered, such as in wideband digital communications systems (square waves), spectroscopy (Gaussian-shaped peaks) and narrow-band telephone-line digital communications.s,

The non-linear low-pass filter is effective in systems that have undesirable commutating spikes such as samplers, encoder and time-shared demultiplexers. The filter can also be used to filter wide-band white noise without introducing any phase shift in the signal of interest.

The corner frequency of the non-linear low-pass filter module is set by the user by adding two external capacitors to the unit. The corner frequency can be set anywhere in the range from dc to 10 kHz. For more information and applications data, write to Non Linear Filters, Box 338, Trumbull, Connecticut 06611, or use *check-off* on page 110.

semiconductor replacement manual

A comprehensive 52-page Semiconductor Replacement Manual has just been released by the Sprague Products Company. Containing over 30,000 OEM part numbers, listed alpha-numerically, which can be replaced by Sprague's new line of 82 popular semiconductor devices, this manual also includes performance characteristics, outline drawings and pertinent parameters for the entire Sprague line.

Included in the new Sprague semiconductor replacement line are 42 small-signal and power transistors, 5 field-effect transistors, 6 silicon rectifiers, 14 linear integrated circuits and 15 LED devices.

These semiconductor devices provide exact replacements for components found in home/mobile entertainment and communications equipment. In addition,

More Details? CHECK—OFF Page 110

the LED devices (which include visible light sources, infrared light sources, alpha-numeric displays and opto-electronic photo transistors) offer hobbyists new options and designs in electronic experimentation.

Copies of Semiconductor Replacement Manual, K-500, may be obtained without charge from Sprague distributors, by writing to Sprague Products Company, Marshall Street, North Adams, Massachusetts 01247, or by using *check-off* on page 110.

course in radio fundamentals

The new fifth edition of *A Course In Radio Fundamentals* is a completely-rewritten version of this long-time favorite, now in its 30th year. Modernizing the text, plus the introduction of much new material to increase the scope, has almost doubled the previous size of the book. If you are introducing someone new to amateur radio, this is a good book to start him off with.

Unlike the preceding printings, which used *The Radio Amateur's Handbook* as a text, the present volume is a complete and independent study manual. Paced at an intermediate technical level, the treatment is quantitative to the extent permitted by restricting mathematics to simple algebra.

The twenty-six chapters fall into two categories: electrical and radio circuit fundamentals and basic principles of electronic amplifying devices. Each chapter ends with a collection of questions and problems for testing the reader's understanding of the subject discussed. Answers to problems are given at the end of the book. The study text is followed by ten experiments using inexpensive, readily-available equipment. Detailed experimental data are contained in each.

The new soft-cover edition has 184 pages and contains well over 200 illustrations, including circuits and graphs. Priced at $2.00 from your local electronics supplier, or order from Comtec Books, Greenville, New Hampshire 03048.

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For a limited time only, SWAN ELECTRONICS and participating dealers are offering the best-buys-of-the-year in amateur radio . . . *complete* mobile or home stations . . . at very low "special package" prices. Act now. Be sure orders are placed early to assure early delivery – these special offers will expire December 25, 1973.

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2 Meter FM 144 Channels $359.95

Independent switching of 12 transmit and 12 receive channels gives you up to 144 possible channel combinations for your communications pleasure. Now you can move off the crowded frequencies, effectively eliminating unwanted QRM. Eight crystals are included for the most popular frequencies.

The FM-1210A is the only 2 meter transceiver providing a crystal oven for superior stability in the coldest of weather conditions. Transmitter is fully solid-state. DC power cord is included for mobile operation and the heavy-duty pedestal type AC power supply is perfect for home station applications. Mobile mounting bracket and dynamic microphone is supplied.

The FM-1210A transceiver may be purchased without the AC power supply at just $319 for mobile installation off any standard 12V DC system.

More Details? CHECK–OFF Page 110

Specials from Swan Electronics

Compact Quarters? Need Space? Try the 300B Cygnet *de novo* with built-in speaker and AC power supply

Yes! A complete amateur radio station expertly engineered into this newest generation of portable Cygnet SSB/CW transceivers. It's lightweight, less than 25 pounds. An ideal traveling companion for Hams on the move. Take it on vacation – operate from motel room, hunting cabin, boat or car. Connect an AC power source, plug in your microphone and antenna – you're on the air!

With 5 bands and 300 watts P.E.P. input, the Cygnet *de novo* has all the control and power necessary to work the world. A CW sidetone monitor is provided along with capability for CW semi-break-in with an optional VOX unit. Requires plug-in DC converter for 12V DC mobile operation

300B Cygnet *de novo* . . $499.95
VX-2, VOX unit $35.95
SWAN 14-A, converter . . $44.95

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Like its big brother, the SS-200, this monobander needs no transmitter tuning and is infinitely protected from VSWR damage. Front-end overload, distortion and cross-modulation is virtually eliminated.

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Includes: SWAN MB-40 or MB-80 transceiver, appropriate antenna coil, 6 foot whip and 36 inch antenna base section, SWAN BMT mount, SWAN 404 microphone, and all necessary mounting brackets, coax and connectors.

THERE'S MORE
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SS-200 (200 watts P.E.P.). $779.00
PS-10 (115V AC power supply) $89.00
PS-20 (115V AC power supply) $139.00

SWAN MARK 6B — Linear Amplifier. 2,000 watts P.E.P. input. Compatible with SWAN 250C. 50 to 54 MHz. Complete with power supply. $679.95

SWAN VHF-150 — 2-Meter Amplifier. 150 watts P.E.P. input. Will operate class "B" or "C". Rugged self-contained power supply $299.95

SWAN 117-XC — 117V AC power supply with speaker, phone jack and indicator light. Ready to plug in and operate $109.95

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SWAN FP-1 — Hybrid Phone Patch. Separate receiver and transmitter gain controls. Exact voice reproduction . . $48.95

SWAN 55C — 5-Band Remote Control Mobile Antenna. Power rated at 1000 watts P.E.P. A SWAN exclusive . . . $129.00

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GOLDEN SWAN 1040-V — Trap Vertical Antenna. 10, 15, 20 and 40 meters, PR @ 2000 W. $69.95
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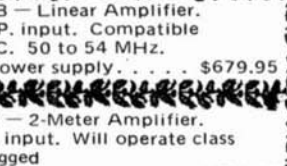
2-METER MOBILE FM STATION FEATURING THE NEW FM-2XA TRANSCEIVER

#3 PACKAGE



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More Details? CHECK—OFF Page 110

MISC.

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6 Meter AM Transmitter with VFO
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See Page 112 for MORE from Barry

 More Details? CHECK—OFF Page 110

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7448	1.15
7450	.35
7451	.35
7453	.35
7454	.35
7460	.35
7472	.50
7473	.65
7474	.65
7476	.70
7480	.65
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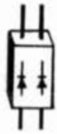

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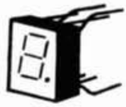

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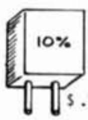

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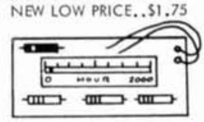

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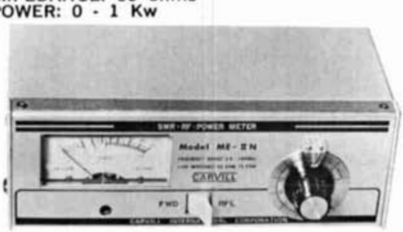

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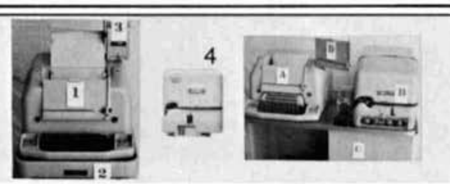

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SPEAKER/PATCH

FP-2
AC POWER SUPPLY

The SP101-P Landliner provides phone patch operation as well as speaker. Front panel: Patch switch, meter, TX and RX gain controls. Rear apron: Receiver 4 ohm output, receiver 600 ohm output, monitor null switch, balance control, line jack, transmitter high Z jack.

The FV-101 permits split frequency operation and control from either the FT-101 or FTDX401.

The FT-101 exciter covers 160, 80, 40, 20, 15, (CB), and 10 meters and comes complete with microphone cable and plug, fused DC power cable and plug. AC cable with plugs and all necessary plugs are furnished. AC and DC supplies are internal.

The FL-2100 linear amplifier needs only 3 wire cable and coax cable. Connectors are furnished.

FTdx401 features high power, super sensitivity and sharp selectivity. The FT-dx401 includes: AC power supply, noise blanker, 100 KC and 25 KC calibrators. VOX break-in, phone patch terminal, cooling fan. Covers 3.5 through 10 MHz plus WWV. 560 watts PEP. All that is required to get on the air is a microphone and speaker.

The FV-401 permits split frequency operation for the DX chaser or net operator. Covers 80 through 10 meters.

FL 2000 B 1200 watts PEP, 1000 watts CW, 600 watts AM. Drive power required 100 watts. Has two cooling fans and uses two 572 B tubes.

FRdx400 includes 2 mechanical filters plus "T" notch rejection tuning, and clarifier for easy zero set for SSB. Crystal control 1st mixer and tunable 1st I.F. provides stable operation and high spurious rejection. 100 KC and 25 KC calibrators. VFO can be used in transceive operation in conjunction with F series transmitter.

FLdx400 operates SSB, (USB LSB selectable), AM, CW and FSK. Circuitry can be built in for RTTY operation. 240 watts PEP. VOX, PTT, and break-in CW.

FL-2000B grounded grid linear uses a pair of 572 B tubes. Plate meter VSWR monitor, 2 fans, built-in power supply, 80 through 10 meters, 1200 watts PEP with distortion product in excess of 30 DB down.

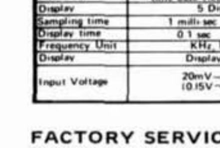
FT2FB $239.00
144-148MHz
10W HIGH 1W LOW

FT2 AUTO $379.00
144-148MHz

FP-2 AC power supply specifications: Output - 13.5 volts, 2 amps. AC input - 100/117/220/234 volts. Speaker - 5" x 3-1/5". Portable or home base operation can be achieved with the addition of the optional FP-2 power pack. This AC power pack provides regulated DC power for the transceiver and charging voltage for optional leak proof rechargeable colloidal type batteries. *In addition, a high fidelity elliptical style speaker is built into the pack.*

The FT-2FB opens the door to noise free broadcast quality 2 meter operation, and thanks to the repeater stations throughout the country, the 2 meter band is no longer restricted to line of sight. General coverage 144 to 148 MHz, 12 channels (3 supplied). Push to talk. Receiver .3 amps, transmit 1.7 amps, power source 13.5 volts + 10%. Dimensions 6-3/8" w. x 2-1/2" h. x 10" d., weight 4 lbs. Comes with dynamic microphone, connector plug, DC cord, fuse and mobile mount.

YC355-D $289.00
DIGITAL COUNTER BUILT IN PRESCALER



Complete with
A.C. Power Cord 6 Ft.
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Signal Test Lead with
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The FT-2 auto is a compact base or mobile VHF/FM transceiver, covering 146 to 148 MHz, featuring electronic scanning up to 8 stations between 146-148 MHz with priority channel sampling while locked on another channel. Adjustable tone burst push-button lock on for repeater actuation. The FT-2 auto is self-contained. Two power cables are supplied with the transceiver, including all mounting hardware, cables, connectors, and accessories required for both mobile and base installation, as well as dynamic push to talk microphone. Operates from various AC voltages or 13.5 DC. Dimensions 8-3/4" w. x 4-1/4" h. x 11-5/8" d. Weight 9 lbs.

SPECIFICATIONS

	YC 355D
Frequency range	5Hz to 35MHz (50Hz to 200MHz)
Accuracy	± time base stability + 1 count
Display	5 Digit
Sampling time	1 milli-sec or 1 sec
Display time	0.1 sec 2 sec
Frequency Unit	KHz, MHz
Display	Display tube
Input Voltage	20mV – 20Vp-p (0.15V ~ 5Vp-p)

		YC 355D
MAX Input Voltage		60Vp-p, less than 10 sec (5V p-p)
Input Impedance		HIGH: 1 M ohm, Low: 56 ohms
Input Capacity		Less than 20pF
Time base Frequency		1 MHz Crystal controlled
Stability		0.0005% at 25° C 0.0025% at 0° – 40° C
Power Require-ments	A C	100/110/117/200/220/234V 50/60Hz 18V A
	D C	12 ~ 14.5V 1A

		YC 355D
		220(W) X 80(H) X 270(D) (8 3/4 W X 3 3/4 H 10 1/2 inches)
Weight		
Tube	Display tube	5
Semi-conductors	Silicon diode	12
	Silicon transistor	9
	FET	1
	IC	26

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RF Freq. (MHz) †	50-54	144-148	220-224	432-436	1296-1300
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N.F. (typical)	2.5dB	2.8dB	3.4dB	3.8dB	9.0dB
Nom. Gain	30dB	30 dB	26dB	28dB	9.0dB
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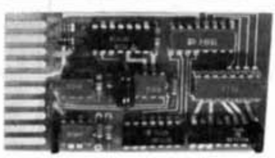

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or 144-146 MHz

25 WATTS

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  PA-1501H — 15-20 WATTS OUT
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- ONLY 2″ x 2½″ x 6″
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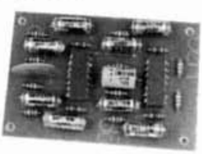

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Build the 2"x3" CWF-2 PC card into your receiver or get the self contained and ready to use CWF-2BX and plug in!

SPECIFICATIONS

BANDWIDTH: 80 Hz, 110 Hz, 180 Hz (Switch selectable)
SKIRT REJECTION: At least 60 db down 1 octave from center frequency for 80 Hz bandwidth
CENTER FREQUENCY: 750 Hz
INSERTION LOSS: None. Typical gain 1.2 at 180 Hz BW, 1.5 at 110 Hz BW, 2.4 at 80 Hz BW
INDIVIDUAL STAGE Q: 4 (minimizes ringing)
IMPEDANCE LEVELS: No impedance matching required
POWER REQUIRED: CWF-2 ... 6 volts (2 ma.) to 30 volts (8 ma.). CWF-2BX ... standard 9 volt transistor radio battery
DIMENSIONS: CWF-2 ... 2"x3" PC board. CWF-2BX ... 4"x3 1/4"x2 3/16" (black winkle steel top, white aluminum bottom, rubber feet)

TRY this fantastic CW filter. If you don't think it is the best you have ever used, ask for your money back. We will cheerfully refund it. These filters carry a full one year warranty.

Write for FREE brochures and magazine test reports. Other IC active filters available: CW mini filter (1½"x2"), low pass, high pass, and wide bandpass filters. Audio amplifiers: ½, 1, 2 watts. Crystal calibrator.

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TUCSON HAMFEST, 28 October, 144 West Lester, Tucson, Arizona. Inside flea market, prizes. Tucson Repeater Association, Old Pueblo Radio Club, Box 6497, Hucson, AZ. 85733.

WANTED: Precision Selectivity Model Motorola Recvrs, P8116C or P8116D. Must be in good working condition, converted or non-converted. WB8NLM, 146 Schonhardt, Tiffin, O. 44883.

QRP TRANSMATCH, Vari Q filter, FM crystal logic oscillator kits, Write Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

SELL — Meteor SB175 xmtr $40.; HM10A tunnel dipper $20; 19″ Emerson portable TV (needs work) new picture tube, $8; BC-221T Zenith frequency meter, $40; RT-35 A/M radio sonde xcvr (675-725 MHz) $3; HA-26 VFO (unused) less manual, $20; Britannica Junior Encyclopedia (1965) $40; Collins F250A85 6409 filter (new) make offer; R-394/U 152-172 MHz receiver with power supply and cabinet for mounting receiver, transmitter & transmitter power supply, $15; 600 watt AC power supply for SB-175 (two power supplies in one) $25; 4 - 6UF 400 VAC Sprague vitamin Q capacitors #GA-51425, $2 each; 4 - antenna relay #423/AR, $.75 each; Western Electric power supply, less tubes #J86219A-3 L-2A 300 V 0.45A D.C. less tubes, $20. Richard Mayer, 3131 Decatur Avenue, Bronx, N. Y. 10467.

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WANTED: Old QSL cards. Vendy Johnson W6CWK, 4960 5th Street, Fallbrook, Calif. 92028.

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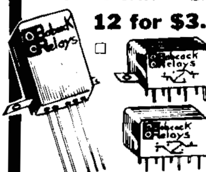

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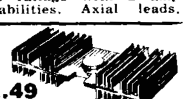

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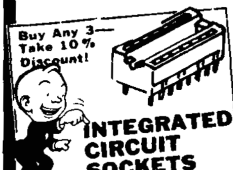

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Receiver-wise, "Cloverleaf" has a front end with unity conversion gain that converts 450MHz band signals to I-F frequencies corresponding to 144MHz channels. Limiter, discriminator, output audio and loud speaker in the 2 meter transceiver continue to function in the usual manner.

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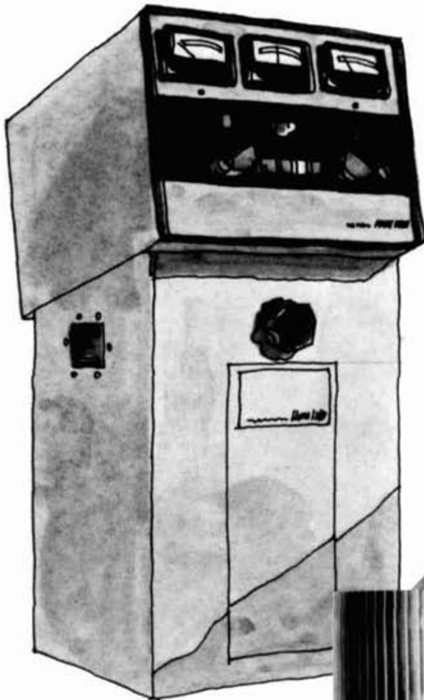

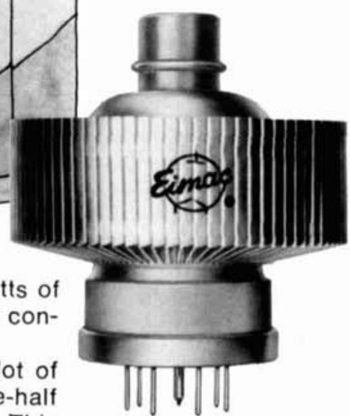

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