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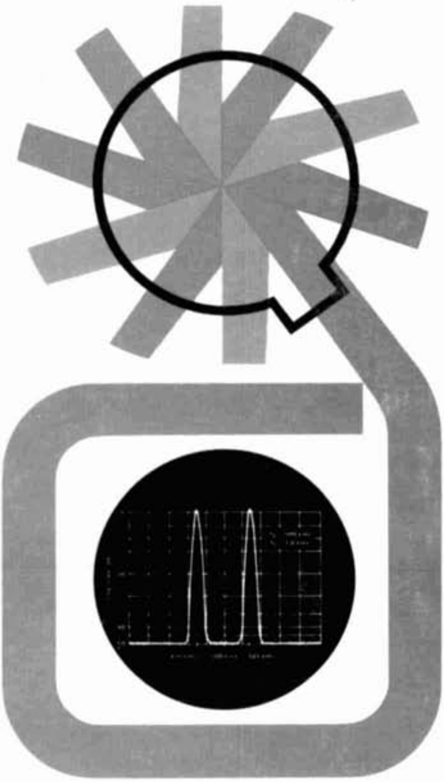

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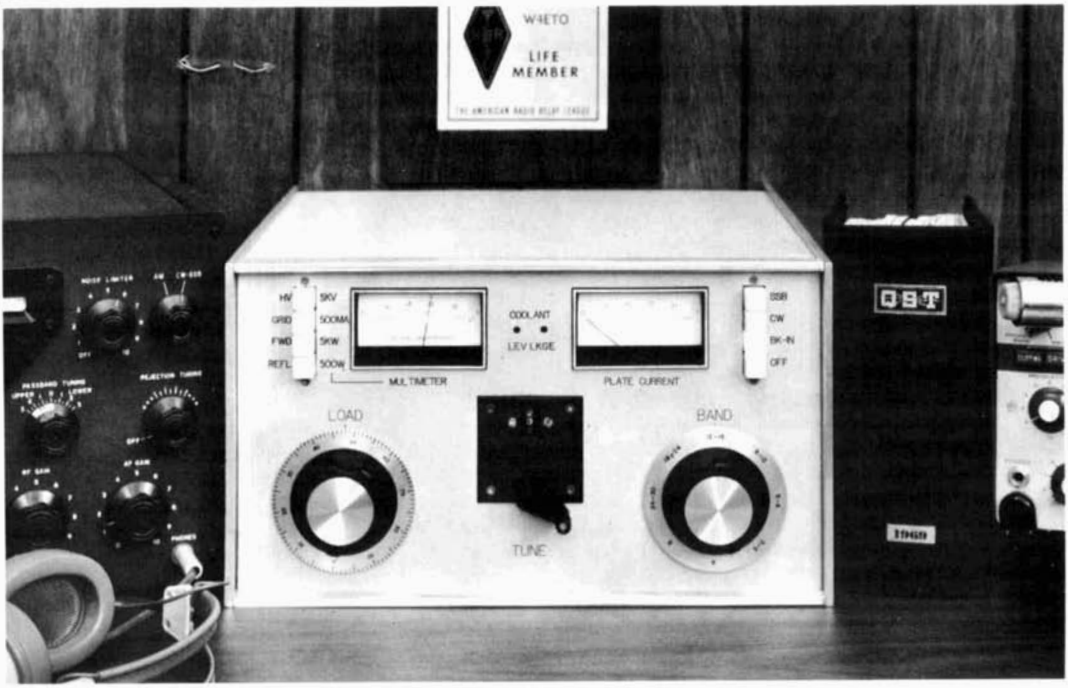

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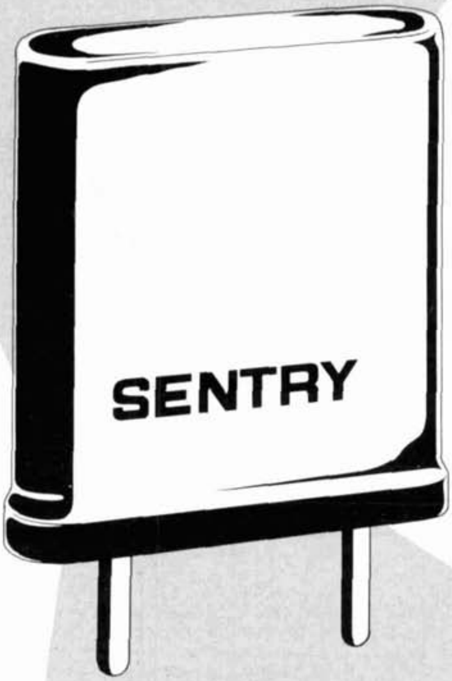

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contents

a second look

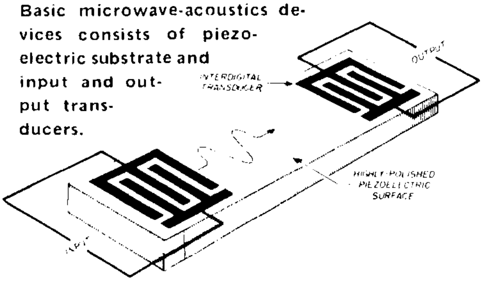
by jim fisk

Microwave acoustics is a new technology that promises to revolutionize communications equipment design in the future. Recent laboratory experiments with microwave acoustics have resulted in amplifiers, oscillators, resonators, signal couplers and delay lines. Nor is this a scientific curiosity confined to the inner-sanctum of the lab—Zenith is working on a 40-MHz acoustic-wave bandpass filter to replace the tuned circuits in color television sets.

The word "microwave" has traditionally been used to describe work in that part of the spectrum where wavelengths are defined in terms of centimeters. However, in "microwave acoustics" the term "micro" is associated with the micron wavelength of an acoustic wave on the surface of a crystal—at 30 MHz, for example, a 100-micron wavelength is possible because acoustic waves travel 100,000 times slower than electro-magnetic radiation.

At the heart of all acoustic-wave devices is the delay line shown below. It consists of a piezoelectric substrate, such as quartz, and input and output transducers. The input transducer is basically a transmitting antenna that converts the incoming electrical signal into an acoustic wave on the surface of the substrate. When the acoustic wave reaches the opposite end of the substrate it is converted back into an electrical signal by the output transducer. The wave propagates rather slowly across the surface of the substrate, and by changing the spacing between transducers signal delay can be controlled.

Basic microwave-acoustics devices consists of piezoelectric substrate and input and output transducers.



Transducer design is extremely important because mass and shape affect efficiency, and size affects bandwidth. The interdigital structure shown in the drawing consists of two separate arrays of metal electrodes which resemble interlaced fingers. By changing the number of fingers the bandwidth can be tailored to circuit requirements. Thus, the basic acoustic-wave delay line becomes a resonator that may be used in place of tuned LC circuits.

Engineers have recently come up with an acoustic-wave amplifier that uses a semiconductor structure spaced a short distance away from the substrate. Since the electric field associated with the propagation of the surface wave extends out of the surface of the substrate it interacts with the electrons in the semiconductor. If the dc supply voltage is low, the surface wave is attenuated because energy flows from it to the slower moving electrons in the semiconductor. However, as the supply voltage is increased, the electrons speed up, and when their speed exceeds that of the surface wave, gain results. Laboratory-built amplifiers using this system have netted gains on the order of 30 dB.

Although it will be some time before these new devices will be available for amateur use, they lend themselves to batch processing. And batch-processed components mean big savings after the design becomes standard—look at the proliferation of low-cost batch-processed ICs currently on the market.

Jim Fisk, W1DTY
editor

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February 7, 1969

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I have drawn a chart of the locations of the various instruments in the rack and am also enclosing it.

The man in one of the pictures is Stan Allison of our transmitter staff.

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Sincerely yours,
Gerald King Allison
Gerald King Allison
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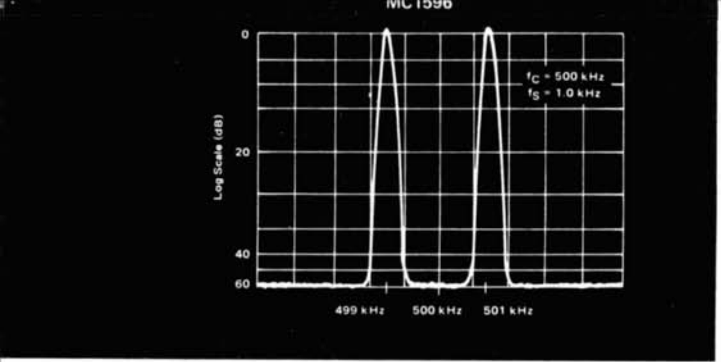


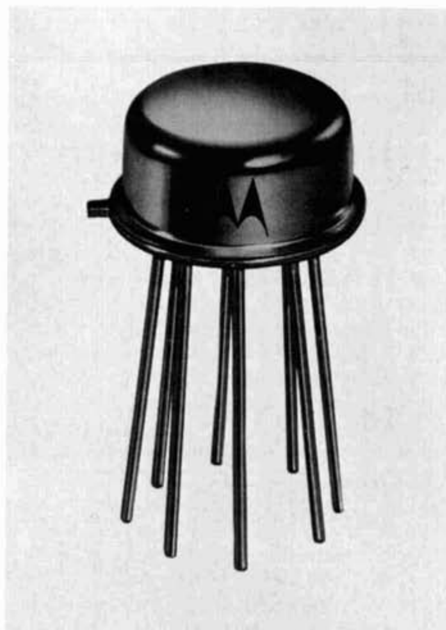
an integrated-circuit balanced modulator

Although designed for balanced-modulator service, the Motorola MC1596G is readily adaptable to many other circuits for amateur use

Roy Hejhall, K7QWR, P. O. Box 3265, Scottsdale, Arizona 85257

Integrated circuits are being designed that can perform more and more of the circuit functions in amateur communications equipment. This article describes one of these new circuits, the Motorola MC1596G balanced modulator. Included are circuits showing the MC1596G as a balanced modulator and several other applications including an a-m modulator, a-m detector, a product detector, a mixer, and a frequency doubler.

The Motorola MC1596 integrated-circuit balanced modulator contains 8 transistors, 3 resistors and 1 diode.



The MC1596G is available now. However, a less-expensive version of the device with a slightly lower carrier-suppression specification and a limited operating-temperature range will be available soon under the type number MC1496G. The MC1496G will still provide very adequate performance for many amateur applications, and the circuits and information in this article apply to the MC1496G as well.

MC1596 performs this multiplication is beyond the scope of this article, and interested readers are referred to the references. However, to put the MC1596G to work it's helpful to have a basic knowledge of the output signal characteristics. Therefore, a brief discussion of the multiplication process with two ac signals follows.

Let's assume we have two sine-wave input signals called A and B at frequen-

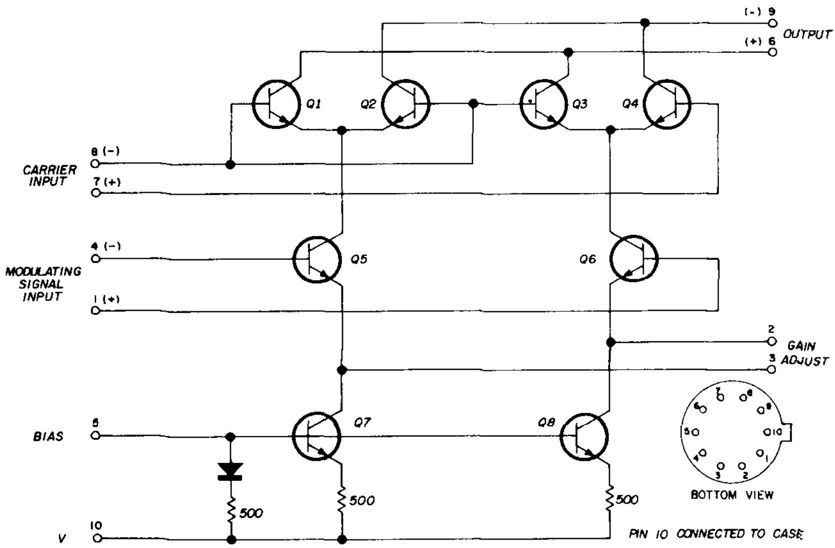


fig. 1. Internal circuit arrangement of the Motorola MC1596G IC.

description of the MC1596G

Fig. 1 is a schematic of the MC1596G. It consists of a dual differential amplifier driven by a standard differential amplifier. Transistors Q1 through Q4 make up the dual (upper) *differential amplifier*, while transistors Q5 and Q6 form the standard (lower) differential amplifier. Transistors Q7 and Q8 are constant-current sources for the lower differential amplifier.

The MC1596G has terminals for two input signals and one output signal. In operation, the circuit produces an output signal that is the product of the two input signals. A detailed discussion of how the

cies f_A and f_B respectively. And suppose we have a device that multiplies signal A times signal B and produces a third signal, C, which is the product of A and B. A device that performs this task is the MC1596G, and signal C will then have the following characteristics:

1. The amplitude of signal C will be the product of the amplitudes of signals A and B.

2. Signal C will contain two (and only two) frequency components, $(f_A + f_B)$ and $(f_A - f_B)$.

Note there is no output at either of the input signal frequencies, f_A and f_B.

An example may be helpful at this point. Suppose we apply two input signals, one at a frequency of 1 MHz and the second at 4 MHz. The output signal will then contain frequency components at 3 and 5 MHz. In other words the output will be two separate, single-frequency sine-wave signals, one at 3 and one at 5 MHz. There will be no output at 1 and 4 MHz.

The signal amplitudes need a little further clarification at this point. The MC1596G has no built-in output load resistors; they must be added externally to develop an output-signal voltage from the output current. The magnitude of this output-voltage depends on the value of these outboard load resistors. Therefore, varying the load resistors will change the gain of the MC1596G. This means that the output signal will be the product of the input signal amplitudes times some constant, which is a function of the external load resistors and other circuit adjustments that affect gain. These gain-related items are described next.

gain considerations

In addition to the load resistances, two other parameters that affect gain, or amplification, from the modulating-signal input to output are the magnitude of the resistance between pins 2 and 3 and the dc-bias currents. For all applications shown here, a bias current of 1 mA in each transistor of the lower differential amplifier, Q5 and Q6, has been used. This is generally recommended for most applications.

The resistance between pins 2 and 3 can be readily tailored to the needs of a particular circuit. Increasing this resistance decreases gain but increases the signal-handling capability of the IC. This means that output-signal amplitude will be decreased, but a higher-amplitude input signal can be handled without distortion. Decreasing this resistance has the opposite effect. The resistance between pins 2 and 3 can be anything between zero and several thousand ohms, depending on the optimum desired combination of gain and signal-handling capability.

Now let's see how the MC1596G can be put to work.

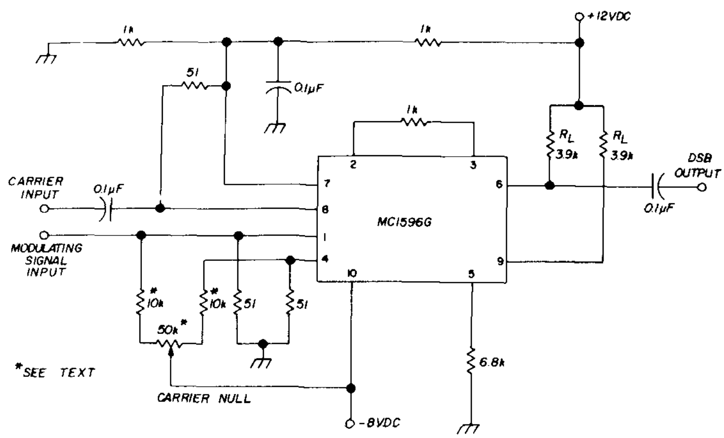


fig. 2. Balanced modulator circuit. For use as an a-m modulator, merely change the two 10k resistors to 750 ohms. The pot can be adjusted for carrier null (ssb) or carrier insertion (a-m).

balanced modulator

Fig. 2 shows the MC1596G as a balanced modulator. This circuit has the following advantages over a conventional diode ring balanced modulator often found in amateur equipment:

1. Circuit simplicity. No transformers are required; only resistors and capacitors. Only a single carrier-null adjustment is used, while diode ring modulators often have two null adjustments. Further, the MC1596G carrier-null adjustment is in the dc portion of the circuit. This means the carrier-null potentiometer need not be located physically near the remainder of the balanced-modulator circuit—it could be located on the rear panel of the equipment or at any remote location.

2. Greater carrier suppression. This balanced modulator will provide typically 65 dB and 50 dB carrier suppression at 500 kHz and 9 MHz respectively.

3. Broadband operation. The basic circuit requires no modifications for carrier frequencies from audio to 100 MHz.

The balanced modulator shown in **fig. 2** also has an extremely clean double-sideband output signal and a low-carrier-oscillator power requirement. Spurious sidebands are down 55 dB or more, and the carrier oscillator must deliver only 0.072 milliwatt to the modulator.

Construction and operation are simple. Reasonable care is needed to isolate output and carrier input. The modulator has such excellent carrier suppression that if care is not paid to circuit input-output isolation, there may be more carrier output signal passed through circuit stray capacitance than through the IC itself.

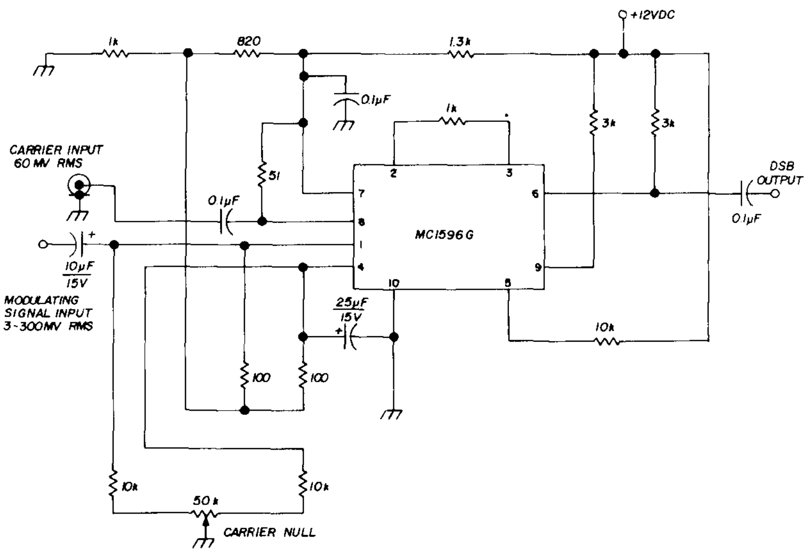


fig. 3. A balanced modulator using a single 12-Vdc supply.

To place the modulator in operation, apply dc power and carrier signal. The recommended carrier input level is 60 mV (0.06 V) rms, but this isn't critical and variations in this figure of as much as 50% won't seriously degrade circuit performance. The carrier-null adjustment is set for minimum carrier output.

The audio modulating signal is then applied. This signal should have a value of 300 mV rms on peaks; but again, this level is not critical.

If you have no equipment available to adjust the carrier injection level to 60 mV, apply both carrier and a single audio-tone modulating signal and observe the double-sideband output level on a voltmeter or scope. As the carrier injection is increased from a very low level (1–10 mV rms) the output will increase as carrier level is increased. Finally a point will be reached where further increase in carrier level causes no change in the output. The carrier level should be set at the point where the output signal just begins to level off.

If only a single 12-Vdc power supply is available, the circuit shown in **fig. 3** may be used. Signal levels and operating instructions are the same as for the circuit in **fig. 2**.

amplitude modulator

The circuit shown in **fig. 2** can be used for an a-m modulator if the 10k resistors are changed to 750 ohms. This modification gives the carrier-null adjustment the greater range necessary for carrier insertion.

could be used in an ssb/a-m transmitter or in an rf signal generator.

doubly balanced mixer

A doubly balanced mixer delivers only sum and difference frequency outputs and suppresses both the local oscillator and rf-input frequencies. The MC1596G may be used in this application also.

Fig. 4 shows a doubly balanced mixer with broadband inputs and a tuned output at 9 MHz. This means that the rf and local oscillator inputs may be any two frequencies with a sum or difference of 9 MHz. The circuit will operate with input frequencies from 160 meters up to 300 MHz. With a 10-meter input signal, the mixer has a conversion gain of 13 dB and a sensitivity of 7.5 μV for a 10-dB signal-plus-noise-to-noise ratio at the i-f output. At 220 MHz, it has 9-dB conversion gain and a 14-μV sensitivity.

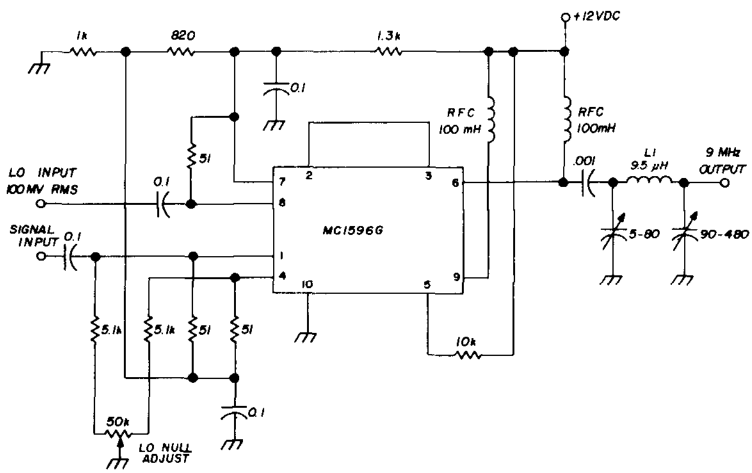


fig. 4. Doubly balanced mixer with broadband input and tuned 9-MHz output. Circuit operates up to 300 MHz. L1 is 44 turns no. 28 enameled on a Micrometals 44-6 toroidal core.

The a-m modulator is operated with the same signal levels as the suppressed-carrier modulators described above. It can be used in a system where both suppressed carrier *and a-m are required. Simply* adjust the 50k pot for either a carrier null or carrier insertion, depending on which operating mode is desired. The modulator

Several variations of this mixer may be used. There are three signal ports, two inputs, and one output. These three ports may be used with any combination of *either tuned tanks or broadband coupling* circuits. Thus, with broadband circuits on all three ports, the mixer will deliver sum and difference frequency outputs from audio through vhf. With tuned input and output circuits, much higher conversion gains of 20 to 30 dB may be achieved due to more efficient impedance matching.

the MC1596G, causes the product detector to have only the desired demodulated audio output.

Fig. 5 shows a product detector circuit. With a 9-MHz ssb i-f input, the

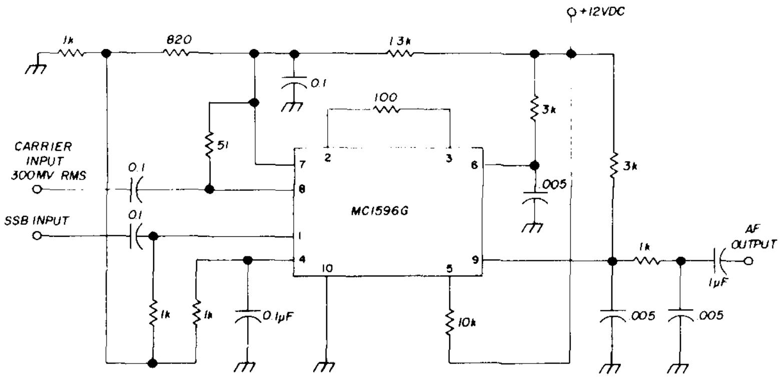


fig. 5. Product detector with 90-dB dynamic range. The high sensitivity of this circuit lends itself to direct conversion techniques.

product detector

An extremely sensitive product detector can be built with the MC1596G.

A product detector is really a mixer with its output in the audio frequency range. The ssb signal forms the rf input,

detector has a sensitivity of 3 μV for 10-dB s + n/n at the audio output. For 20 dB s + n/n, the sensitivity is 9 μV. This means that for an ssb receiver with a 50-ohm antenna input impedance, a 0.5-μV rf input signal would require only 12 dB over all signal power gain from

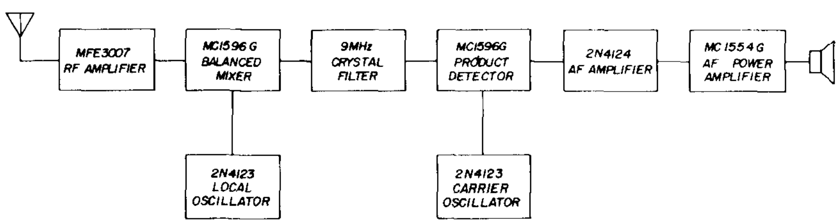


fig. 6. Block diagram of a receiver with no i-f gain. Sensitivity on 15 meters is less than 0.1 μV for 10-dB s + n/n ratio.

and the carrier injection signal is the local-oscillator input. The audio output signal is at the difference (audio) frequency between the ssb and carrier frequencies. A low-pass filter that cuts off above 3 kHz is used at the output. This, together with the inherent suppression of both input-signal frequencies provided by

antenna to detector input to produce a demodulated audio signal with 20-dB s + n/n at the detector output.

Of course there would be many other practical limitations on such a receiver such as agc range, audio-amplifier sensitivity, etc. But the point is that detector sensitivity would certainly not be a problem in any receiver employing the MC1596G as a product detector.

This high sensitivity product detector permits some interesting receiver techniques. For example, the circuit lends itself well to direct conversion.[3,4] Furthermore, it's possible to build a sensitive superheterodyne receiver with no i-f gain. The latter principle has been realized in an hf ssb receiver (block diagram shown in **fig. 6**). The sensitivity of this receiver on 15 meters is less than 0.1 μV for a 10-dB s + n/n ratio.

The product detector has a dynamic range of 90 dB. This means that it will detect i-f input signals from 3 to 100 mV without significant distortion.

The input-signal-handling capability may be increased at the cost of some decrease in detector sensitivity and gain by increasing the resistor between pins 2 and 3 to 500 or 1000 ohms.

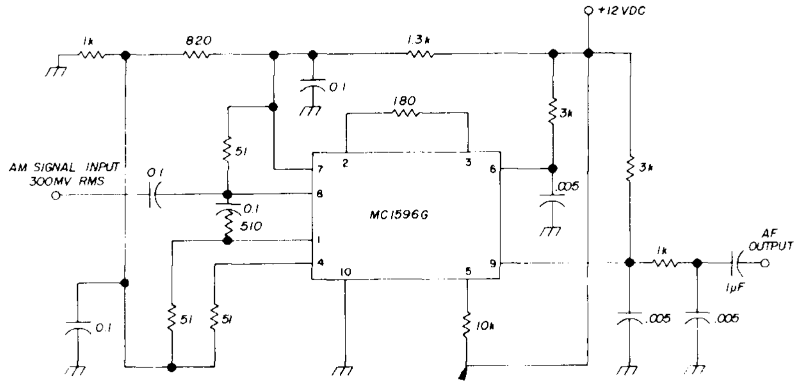


fig. 7. Optimized a-m detector based on the circuit of fig. 5.

a-m detector

The product detector shown in **fig. 5** may also be used as an a-m detector. Thus it can be used as the only detector in an a-m ssb cw receiver.

For a-m operation, simply inject carrier and modulated signal inputs, as for

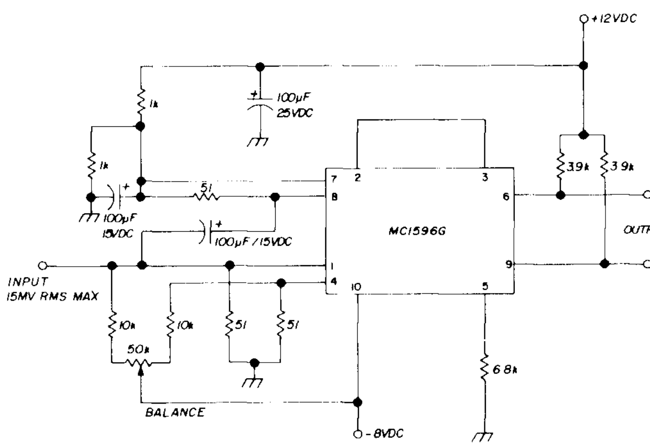


fig. 8. A low-frequency doubler. Circuit will deliver doubled output from input signals up to 1 MHz; all other frequencies are 30 dB or more below the desired output.

ssb. The carrier injection is most conveniently obtained from the i-f signal, thus avoiding any drift problems that may be encountered if the carrier is generated locally, as for ssb reception.

range through vhf.

Fig. 8 shows a low-frequency doubler. The two input terminals are simply ac-coupled, and untuned RC coupling is used at input and output. At input frequencies

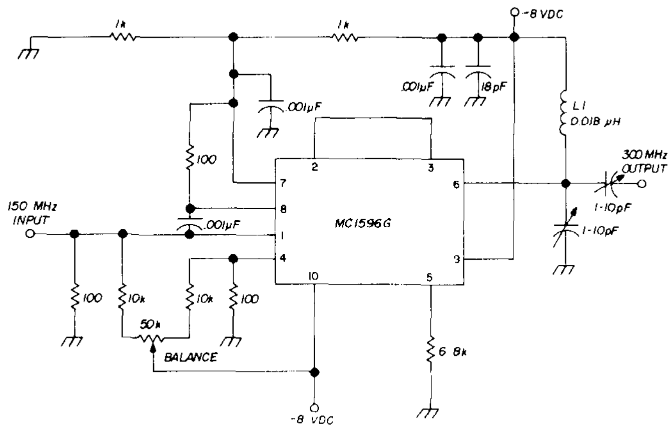


fig. 9. A frequency doubler for the vhf range. Inductance L1 is 1 turn of no. 18 wire, 7/32-inch I.D.

Normally, a constant-amplitude local-oscillator (carrier) signal is injected at the local-oscillator input. To achieve this with a carrier signal obtained from the receiver i-f signal, a limiter would have to be used to remove the modulation. However, if a carrier injection level of 300 mV rms is used, the fully modulated signal may be injected directly.

While the product detector shown in **fig. 5** may be used directly as an a-m detector by injecting a 300-mV signal into the carrier input and up to 30-mV signal into the signal input, a few modifications can be added to optimize the detector for a-m. The resulting circuit is shown in **fig. 7.**

frequency doubler

Injection of the same signal frequency at both inputs produces an interesting result. The sum frequency output is twice the input frequency, and the difference frequency is zero. Therefore, the output consists of a single-frequency signal at double the input frequency, and we have a frequency doubler. The MC1596G operates as a frequency doubler without any tuned circuits from the audio frequency

up to 1 MHz, this circuit will deliver a clean doubled output with all other frequencies 30 dB or more below the desired output.

When modified with suitable coupling and bypass capacitors, the basic MC1596G doubler has been used at input frequencies as high as 200 MHz.

Suppression of input frequency and other spurious signals is not as good in the hf and vhf range. Therefore, it may be desirable to use a tank circuit at the output to obtain a cleaner output signal. **Fig. 9** shows a vhf doubler with such a tank. This circuit doubles from 150 to 300 MHz, with all spurious outputs 20 dB or more below the desired output signal.

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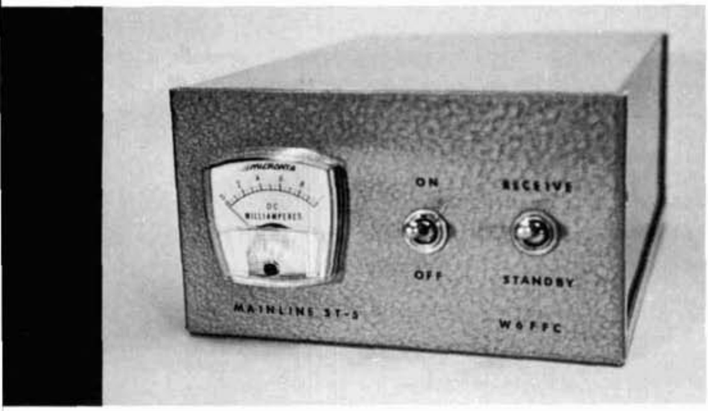
ham radio



the

Mainline ST—5

rtty demodulator

This basic building block demodulator featuring linear IC's can be used for future expansion of your rtty station

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Many newcomers to rtty have complained that a current yet simple demodulator hasn't been published for them to build. The W2PAT unit in the ARRL handbook is nearly 15 years old. In 1964 an attempt was made to replace the W2PAT design with a modestly priced updated unit, the TT/L.[1] This design, together with the subsequent TT/L-2,[2] is now the standard of the serious rtty enthusiast. However, the original goal was missed by a country mile, since the TT/L-2 costs over $160 just for parts and has 14 tubes.

The ST-3[3] was a successful solid-state design that introduced integrated linear operational amplifiers to rtty. It was still moderately complex, however, and fell short of the goal to supply the beginner with something that could be built in a few hours.

the ST-5 demodulator

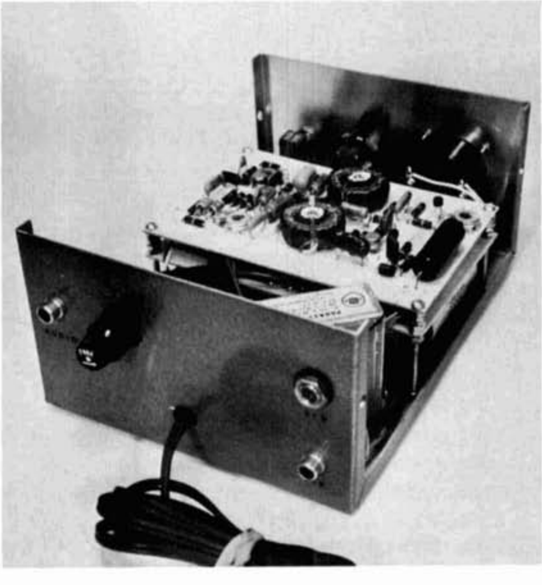
While developing a unit based primarily on ICs to replace the TT/L-2, a very simple modulator with great potential was developed: the ST-5. As with any simple circuit, the cost of the power supply is out of proportion with the rest of the unit. At current prices, the ST-5 costs only $14.50 less loop supply ($8) and a plus-minus 12-volt supply ($11).

The total cost of $33 is not overly impressive until you realize this unit can, if desired, be used as a building block for the more exotic ST-6, which will be published later in the year. Almost every component used here can be used in that unit. The ST-5 is a basis from which the beginner can expand—it's not just a collection of parts that will find no further use when he is ready to broaden his horizons to more sophisticated equipment.

features

The ST-5 uses two operational amplifiers **(fig. 1)**. One is an audio limiter, and the other is a trigger stage to drive the keyer. It has a 175-volt loop supply of the same type used in the TT/L, which provides plus-minus voltages for keying a transmitter and also features narrow-shift cw identification. Finally, the ST-5 has a symmetrical plus-minus 12-volt power supply.

Rear panel of ST-5 shows jacks for audio, fsk and loop power. Enclosure is 6 x 8 x 3½ inches.



limiter

The 709C op amp has over 90-dB gain and is good to over 10 MHz. It makes an ideal limiter. The zener diodes on the input don't assist in the limiting; they merely protect the 709C against damage in the event of excessive audio input (hardly likely but worth the protection). The limiter puts out square waves and is so powerful it starts working on input signals as low as 200 μV. The 25k pot merely balances the small offset input voltage for maximum gain. This voltage varies slightly from one unit to another, so a control pot was added rather than a fixed resistor, which many units use.

discriminator/detector

It's difficult to use the same value inductor with different capacitors and expect to obtain two similar filters of equal characteristics. To get similar bandwidth, voltage output, noise response, etc., some loading is necessary. Most simple demodulators merely balance the voltage or ignore all the problems completely. Without belaboring the point, it's not a simple job to get all these factors to balance suitably; but it *is* possible, and the Mainline units all have filters that have been designed with care.

The ST-5 offers a choice of the 2125-2975 mark and space tones (considered standard), or the 1275-2125 low tones necessary in some modern receivers. (Actually nearly all these receivers respond beautifully to 2975 tones and higher, but a new bfo crystal is needed.) The best results come from the 2125-2975 tones, since the two frequencies are only about 28% apart while the 1275-2125 tones are 40% apart; thus it's a more difficult job to separate the harmonics and achieve proper filter design.

The detector features full-wave rectification for most efficient filtering of the dc ripple remaining after the audio has been rectified. A simple RC low-pass filter removes the remaining audio component.

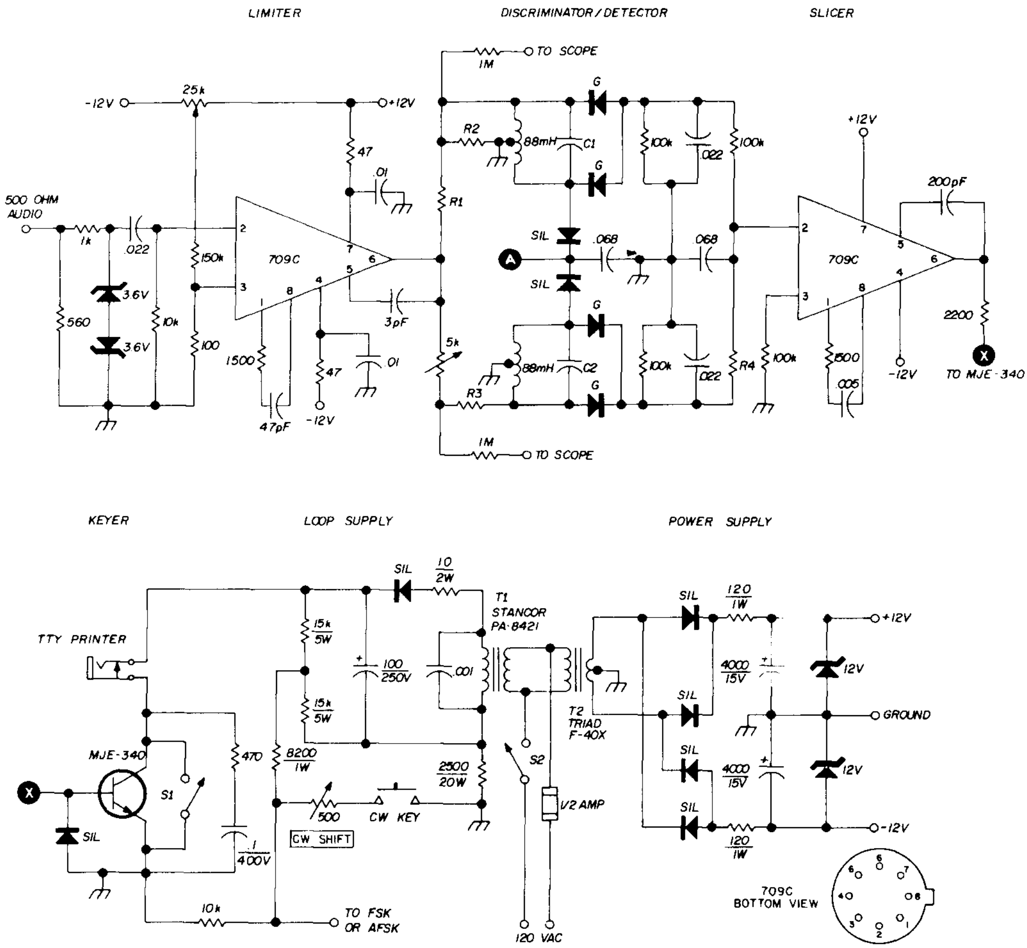


fig. 1. Schematic diagram of the Mainline ST-5 RTTY demodulator.

slicer

The slicer takes the small voltages from the filters and changes them to roughly +10 volts for mark and -10 volts for space, regardless of the original amplitude. This in reality is a dc limiter, as a signal as small as a 100 μV or so will cause the unit to saturate completely, either plus or minus, depending upon the polarity of the applied signal voltage. The unit has so much gain that at the crossover point, a change at the audio input as small as one or two Hz will cause this trigger stage to flip from +10 to -10 volts. This is another way of saying shifts as low as 3–4 Hz could be copied on the ST-5 if tuned in properly.

keyer stage

A 300-volt Motorola 25W transistor selling for $1.06 is used. The normal loop-supply current for tty machines is 60 mA. This transistor has a large amplification factor and acts like an on-off switch. When on, the power consumed in the transistor is only 0.012 W; so in the ST-5 there's no way you could ever damage that transistor.

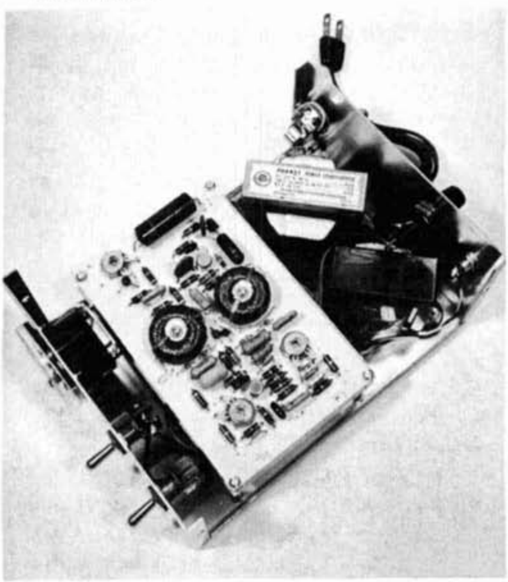
An RC network in the MJE-340 collector takes care of the back emf developed by the inductance of the selector magnets in the printer during the transition from space to mark. The transistor is biased off during space. A diode in the base circuit keeps this negative voltage below the point at which the base-emitter junction would be reverse-biased.

loop supply

Here comes $8 of the $33 total right now. This unit uses the well-known "floating loop" I developed for the TT/L. As you go from mark to space, the voltage at the fsk output switches from negative to positive. This offers excellent keying characteristics for the transmitter, and provides a simple method of keying ssb transmitters needing conduct on mark instead of conduct on space, such as the Collins S-line. The Hallicrafters (and others using a 9-MHz heterodyne scheme with a vfo running from 5.0 to 5.5 MHz) needs both systems. If you are "upside down," merely reverse the diode in the fsk system. Few (if any) other systems offer this potential. The narrow-shift cw identification system can be set appropriately with the 500-ohm pot. If you are using a transmitter that conducts on mark and can't get suitable cw shift, try putting the connection to the 500-ohm pot on the other end of the 8.2 kilohm resistor. One of these two places has always been adequate in the past.

The 2500-ohm resistor sets the loop current, which is in no way critical. Unless more than 10 mA in error from 60 mA (you may have used a different transformer and need a different resistor), don't bother changing anything.

Printed-circuit boards hold all the components except the two transformers and the control switches.



loop transformer

The Stancor PA-8421 is far from cheap. However, I've never found a more suitable transformer at any lower price. Don't be alarmed at the 50-mA rating, which requires some explanation. The primary is capable of handling about 20 VA in the secondary. Since there's also a 6.3-volt winding rated at 2 A, this is almost 13 of that 20 VA. That only leaves about 7 VA for the high-voltage winding, or roughly 50 mA. However, if the filament winding is not used (and I don't use it), then the entire 20 VA is available to the high-voltage winding. This represents around 160 mA. So don't be alarmed at the 50-mA rating. You could pull twice that in this circuit and it wouldn't tax the transformer. Don't worry if the transformer gets warm; all transformers get warm. It's when you burn your hand on them that you have to watch out. I've had a loop supply similar to this running in the TT/L twenty-four hours a day for six years, and others throughout the country are doing the same.

power supply

The 709C op amps will take up to ±18 volts. If you wind up with more than the indicated ±12 volts, but less than ±18 volts, think nothing about it. You can lower the voltage by increasing the value of the resistors if desired. The plus voltage goes up or down at the same rate as the minus voltage, since both voltages are supplied by the same transformer. The op amps use symmetrical voltages. The 4000-μF capacitors are Sprague type 39D at $2.43 each. Other large-value brands may be used, and I suggest you use at least 3,000 μF for this purpose if substituting.

standby switch

When S1 is closed, the unit is placed in mark. When S1 is opened, the printer can follow whatever is fed into the limiter from the receiver.

As explained previously, the unit has so much gain that a signal as small as 3–4 Hz can be copied if tuned correctly; this is called straddle tuning. However, for 170-Hz shift you may wish to add a switch that changes the space filter to the new frequency. **Fig. 2** shows the way this would be accomplished if using the normal 2125-2975 tones, and **fig. 3** shows the circuit for the low tones of 1275-2125. This is merely an expedient and doesn't result in proper filter balance, but it provides good 170-shift reception with the switch closed, or normal 850-shift reception with it open.

tuning indicator

Provisions are provided for connections to the vertical and horizontal amplifiers of a scope (**fig. 1**). It is customary to connect the mark signal to the horizontal amplifier and space signal to the vertical

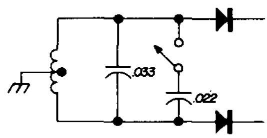


fig. 2. Switching circuit for adding 170 shift to space filter for 2125-2975 tones.

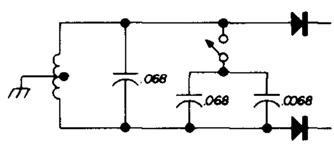


fig. 3. Switching circuit for adding 170 shift to space filter for 1275-2125 tones.

amplifier, although many reverse this method.

Most people prefer a scope indication, but an excellent tuning indicator is provided at point A (**fig. 1**). A voltmeter connected to this point will give equal voltage indication for mark or space. With rtty signals the meter should stand still. If it doesn't, retune the receiver until it does. If straddle tuning a signal, the meter may read less than normal, although it won't move. This is normal and merely indicates the shift being copied is not the correct shift for the filters you're using.

Fig. 4 also shows how a 0–1 mA meter may be added. An inexpensive npn transistor is used, such as the MPS-3394, although any npn transistor would be satisfactory here. The capacitor merely dampens the meter so it doesn't flip around too violently. If your meter is too damped, remove the capacitor or try a smaller value. This was suitable for the inexpensive imported meter used in my unit.

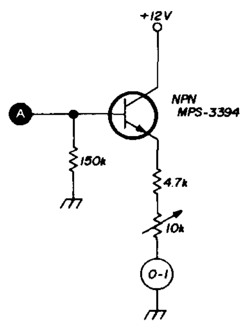


fig. 4. Simple tuning indicator uses an inexpensive 0-1 mA meter.

the transmitter keyer

Fig. 5 shows a typical fsk keyer for installation in the transmitter. The components can be mounted on a small terminal strip and placed near the vfo tube under a convenient mounting screw which also serves as a ground return. The trimmer is connected to the cathode of the vfo tube and the tube replaced in its socket; thus, no changes of any type are made to the transmitter and its resale value is not affected. There should be room for several keyers if you wish to have the convenience of both 170 and 850 shift.

Although a 3-12 pF trimmer is shown in **fig. 5**, some transmitters only require a 1.5-7 pF trimmer. It is suggested that you do not substitute for the 1N270 diode as it is superior to most other types in this application.

If your signal is reported as "upside down," reverse the 1N270 diode. If you do not obtain sufficient cw shift with this connection (conduct on mark), the

500-ohm cw-shift pot should be connected to the opposite side of the 8.2k resistor at the junction of the two 15k resistors (**fig. 1**).

components

The 709C op amps are supplied by various manufacturers including Signetics, Fairchild, and Motorola. The prices in the order named are $2.62, $2.65 and $2.80 as of this writing. Prices are constantly being reduced as devices become available from more companies. When I first started working on a super deluxe demodulator in the fall of 1967, I paid over $10 each. Now they're too cheap not to use.

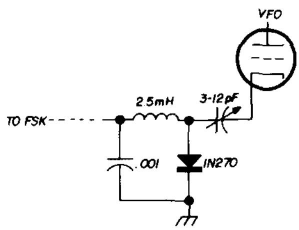


fig. 5. To add fsk to literally any transmitter, connect a 3-12 pF trimmer to the cathode pin of the vfo tube.

The Motorola unit can be purchased through most distributors, including Allied and Newark. The Fairchild unit can be mailordered from the firms below.* Include about $1 extra for handling and postage; the surplus will be refunded. Specify the TO-5 can, as this is easier to work with than the dual in-line 14-pin type (same cost).

The diodes marked G in **fig. 1** are 1N270 germanium at 32¢ each. Those marked SIL are most any silicon type, such as the 1N2069. The one in the loop supply should, however, be a minimum of 400 volts PIV. Fifty-volt PIV is suitable everywhere else.

*Hamilton Electro Sales, 340 East Middlefield Road, Mountain View, Calif. 94040 and G. S. Marshall Co., 732 North Pastoria Avenue, Sunnyvale, Calif. 94086 (also carries Signetics). If buying Motorola version, ask for the MC-1709CG. Texas Instruments 709 op amps are $1.50 each (or 7 for $10) from HAL Devices, Box 365H, Urbana, Illinois 61801; ask for SN72709L.

If you don't wish to spend the money for the zeners on the limiter input, you can substitute regular silicon types as shown in **fig. 6**. These start clipping at 0.6 volt, however, and offer an inferior form of limiting, although they more than adequately protect the input to the limiter from excess voltage. The zeners are a much better choice.

The 88-mH toroids are available from various sources for about 40¢ each.† They're wired in series for 88 mH, and the junction of the two windings is grounded.

If you have an accurate means of determining the frequency, you can tune the filters by removing turns of wire from each of the two sections concurrently to keep the turns ratio in the two windings the same. One turn from each of the two windings will increase the frequency about 6 Hz at the 2125 frequency, for example.

Use Mylar capacitors, such as the Sprague Orange Drop. Twenty-five-volt capacitors are adequate, but you'll probably wind up getting 200V types. They are only 15–21¢ each.

The pots can be the inexpensive 39¢ Mallory PC board MTC types. Other

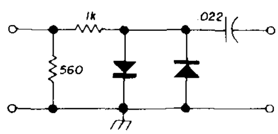


fig. 6. Ordinary silicon diodes may be substituted for zener diodes at the input, although limiting is somewhat inferior.

power transformers may be used, but the Triad F-40X is an excellent buy.

printed-circuit board

The printed-circuit boards shown in the ST-5 in the photographs hold all of the components except the two transformers and the control switches. This greatly enhances construction, and at the same time makes it possible for nearly anybody

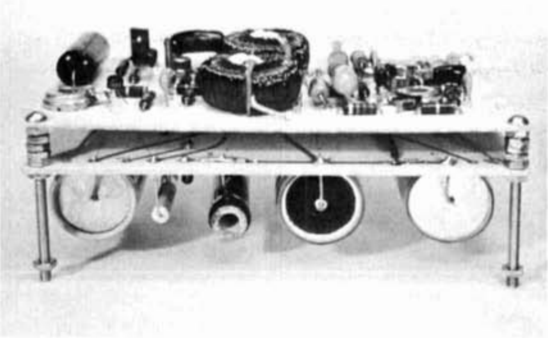
† An excellent source for 88-mH toroids is M. Weinschenker, Post Office Box 353, Irwin, Pennsylvania 15642. He will send you five 88-mH toroids for $1.50, postage paid.

to build an extremely nice-looking unit. The printed-circuit board includes one section for the power supply and another for everything else. The board may be split down the middle and the two sections mounted back-to-back as I did in my unit, or the board may be left intact and used with a more shallow chassis.*

adjustment

With no input signal, or with the input grounded, adjust the pot on the limiter for zero volts dc at pin 6. If this isn't possible, you'd better write me and explain thoroughly, as you probably ruined the op amp somehow. By the way, unless

In this model the printed-circuit board is cut in two sections and mounted back-to-back.



you get too much voltage on pins 2 or 3, like the full power-supply voltage, or get the plus-minus hooked up backwards, it's very difficult to ruin these things. By following even the most elementary construction practices, you'll have no problems with the 709C.

*A printed-circuit board for the ST-5 RTTY terminal unit is available from Stafford Electronics, Inc., 427 South Benbow Road, Greensboro, North Carolina 27401. The undrilled board is $3.00; with critical holes drilled, $3.75; complete, ready to mount components, $6.50. A complete kit of components (less circuit board), including ICs, transistors, diodes, resistors, capacitors and toroids is available for $37.50 from HAL Devices, Box 365H, Urbana, Illinois 61801.

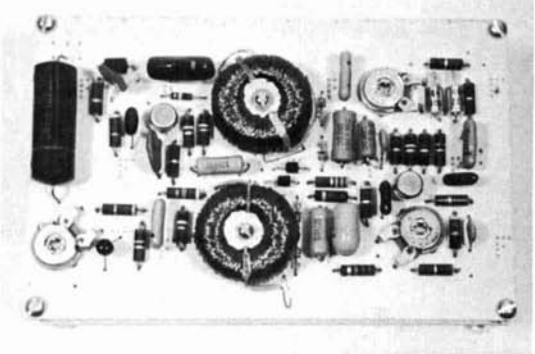
After balancing the limiter for zero volts output, connect the receiver and tune to maximum mark and note the indication on your tuning indicator **(fig. 4)** or on a voltmeter connected to point A. Tune to space on the receiver and again note the reading. If the indications are not the same, adjust the 5k pot on the limiter output until they are. You have now finished all the adjustments and they should require no further attention at any time unless you switch to 170 shift, for instance. In this event you may or may not want to reset the filter balance pot. I suggest you leave it for the 850 setting and take what you get on the 170-switch position, as this is a somewhat artificial method of getting good 170-shift reception.

When transmitting be certain to first close the standby switch or you can get feedback, which will produce errors similar to those you would get when using a microphone if you didn't turn off the speaker.

other op amps

The 709C is to other op amps what the Ford V-8 was to other automobiles. It not only led the way; it's still in use. The 709C was (and is) one of the cheapest ICs of its type available. One would gain very little and stand to lose a lot by trying to substitute other units. The 741

Component layout on the main section of the printed-circuit board.



and 748, for example, have a bit more gain, higher input voltages, and require no frequency compensation. They cost $4.85 each, but their biggest disadvantage here is that they're not at all suited as audio amplifiers. At 2 kHz they have only 30-40 dB gain and make a poor audio limiter compared with the 709C. So unless you know what you're doing, stick to the 709C.

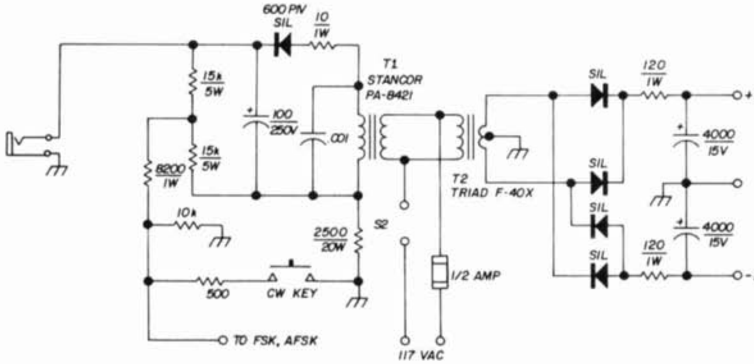
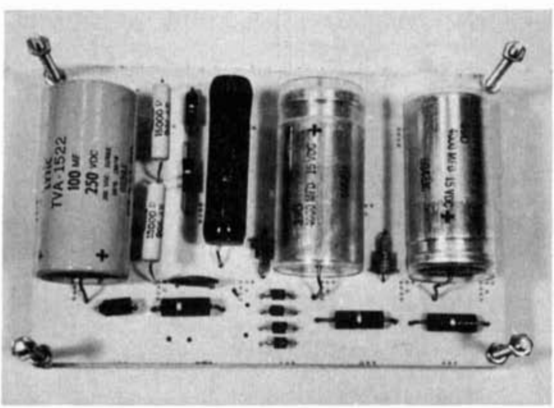


fig. 7. Although this unregulated power supply may be used with the ST-5 RTTY demodulator, the circuit shown in fig. 1 is recommended.

the Mainline ST-6

This demod will be published as soon as possible. The ST-6 uses the same limiter and slicer as the ST-5 and the same keyer and loop supply. It uses the same power supply to which has been added some regulation. It has a total of 7 op amps and 9 transistors. Also featured are an antispace circuit, an active 3-pole Butterworth low-pass filter, autostart with delayed motor control, and optional features such as bandpass input filters for layed motor control, and optional features such as bandpass input filters for either 170 or 850 shift, fast-slow autostart, etc. This is only mentioned at this time to illustrate that if one builds the ST-5, the same parts may be used later for the more exotic ST-6 if you wish to expand your station.

Component layout on the power-supply section of the printed-circuit board.



conclusion

The ST-5 was designed as a simple but highly effective rtty demodulator using the best of currently available concepts. It should be a very popular unit for some years to come, as it's impossible to imagine at this time how any additional performance could be made available—it's already ridiculous to talk in terms of 90+ dB amplification. Only a completely different concept of rtty processing could outdate the ST-5, and that seems quite unlikely to occur until we all get computer terminals in the shack.

references

1. *RTTY,* November, 1964; also *QST,* August, 1965.
2. *RTTY Journal,* September, 1967; also *QST,* May, June, 1969.
3. *RTTY Journal,* September, 1968; also *QST,* April, 1970.

ham radio

an fm receiver

for two meters

Conservative design, solid-state construction, readily available parts— all add up to a really solid vhf fm receiver

Ron Vaceluke, W9SEK and Joe Price, WA9CGZ

Within the last several years, hams have been giving increased attention to vhf fm operation. Although this mode has been used by hams for over 15 years, it is just beginning to be widely accepted. One of the reasons is because of the release from commercial service of the older, wide-deviation, tube-type gear. Commercial users had to go to a narrower deviation to open up more frequencies. One of the authors was on wideband fm 16 years ago using what was at that time fairly current equipment. After all these years, the same type of gear is still widely used. With today's techniques and technology, much of the older tube-type equipment is obsolete. The tragic part is that since the demand for vhf fm equipment has increased, the law of supply and demand has raised the prices of these antiques out of proportion to their usefulness.

Lately several distributors have been selling imported fm gear designed for ham use. Although we haven't tried this equipment, it probably works quite well. However, we felt that it should not be necessary to get a second mortgage on our homes to be able to purchase hobby equipment. A problem that can exist with imported electronic gear is the availability of adequate repair parts when needed. With all of these objectionable possibilities in mind, it was felt that the best approach to vhf fm operation was of the do-it-yourself type. The solid-state, conversion receiver described here is the result.

In establishing our design criteria we felt that it would be wise to spend a few extra dollars, build more than just a basic receiver, and do the job right the first time to obtain good rather than mediocre operation. In looking over readily available semiconductors, best use was made of both discrete and integrated devices.

The first version of the receiver was built on punched board, using eyelets at every point of component entry. Although this technique is adequate for initial design, it leaves something to be desired as far as a finished product is concerned. The second version was made using printed circuits.* All parts are readily available and standard. The i-f transformers are imported, but are a standard type used in portable radios. These are also available from J. W. Miller. Don't overlook those defunct transistor radios at hamfests, etc., as a source of transformers. The coil forms in the rest of the circuit are made by Cambion (avail-

*A set of G-10 epoxy PC boards, tinned, drilled, and with swaged terminals is available for $9.50 P.P. A PC board for the power supply, G-10 epoxy, drilled and tinned is available for $2.50 additional. Order from RMV Electronics, P. O. Box 283, Wood Dale, Illinois 60191.

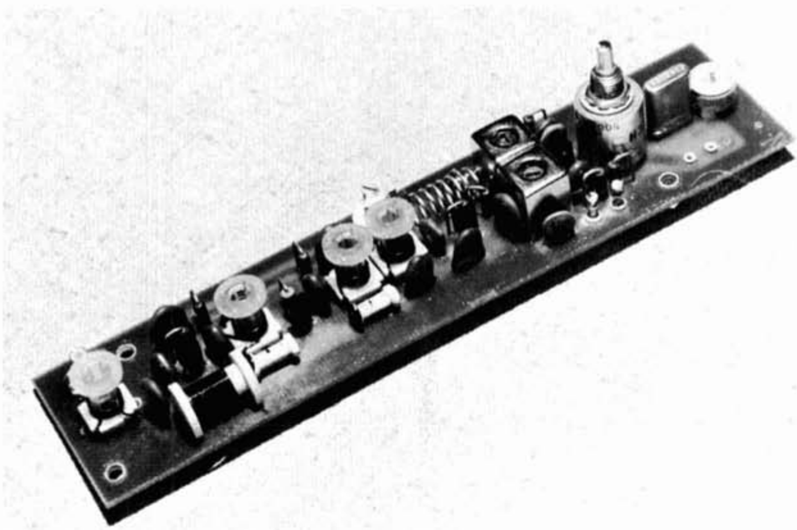
able at Newark Electronics, etc.) and can be shielded by the same size coil shields used on the i-f transformers.

the front end

For good cross-modulation and overload characteristics, fet transistors are used in the front end **(fig. 1)**. The first rf amplifier, Q201, is a conventional, grounded-source, neutralized MPF-102. Other types may be somewhat better; the MPF-105 or MPF-107, for example.

variety is recommended since the larger ones have too much inductance. In the printed circuit version we used an ALCO MRA-3-3S switch. The pins went right through the board and to the foil.

The first mixer is a MPF-102 fet. The signal and local oscillator frequencies are fed to the gate. The source resistor was chosen empirically for the proper amount of local-oscillator injection. The mixer output is 10.7 MHz and is taken off by a link on T201.



View of the PC front end. At the left is the rf input and Q201, followed by Q202. At right center is Q203 mixer, and at the right end of board is the oscillator. The board will hold up to three crystals although only one is shown here.

A grounded gate stage, Q202, is used as the second rf amplifier to simplify construction and for ease of adjustment (no neutralization).

The first local oscillator consists of a crystal oscillator, Q204, operating at 45+ MHz, which is tripled to approximately 135 MHz by a 1N914 diode. The crystal frequency can be determined as follows:

$$\text{xtal freq} = \frac{\text{operating freq} - 10.7\ \text{MHz.}}{3}$$

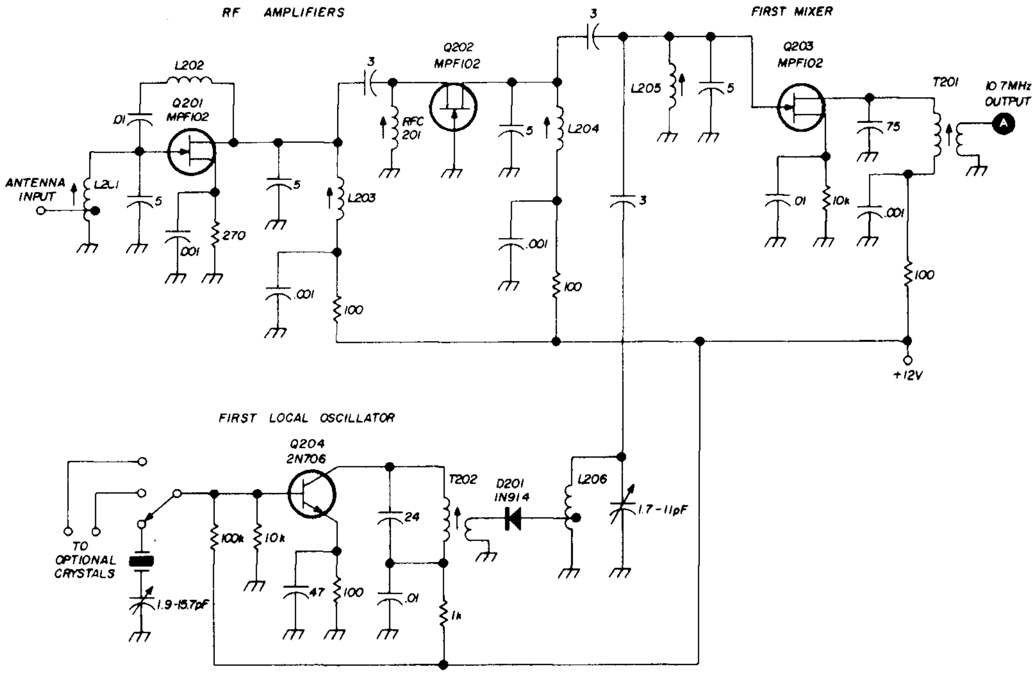
A trimmer capacitor in series with the crystal allows frequency zero adjustment. Be sure to use short leads from the base of Q204 to the crystal, or it won't oscillate. If a switch is used, the miniature

main i-f board

The output of the first mixer is fed to the gate of Q101, the first i-f amplifier **(fig. 2)**. Link coupling from the first mixer provides an impedance mismatch, which prevents Q101 from oscillating without neutralization.

Due to the low Q of the transformers in the mixer and first i-f amplifier outputs, there may be a problem with an image 910 kHz from the desired signal. If you're in an area with lots of activity, or discover later that this problem exists, an external filter can be added between the converter and first i-f.

The output of Q101 is transformer-



L201	4½ turns no. 18 on 0.210" diameter slug-tuned coil form (CTC 3624-4), tapped at 1¼ turns
L202	18 turns no. 26 on 0.214" diameter slug-tuned form (CTC 3104-4)
L203, L204	4 turns no. 18 on 0.210" diameter slug-tuned form (CTC 3624-4)
L205	3 turns no. 18 on 0.210" slug-tuned form (CTC 3624-4)
L206	7 turns no. 20, 3/16" diameter, air wound, tapped at ½ turns
RFC201	8.2 µH (Nytronics DD-8.20 or equivalent)
T201	CTC 3624-2 coil form wound full with no. 32, 8-turn link, shielded (winding area of CTC 3624 is 0.183" long, 0.210" diameter)
T202	CTC 3624-3 coil form wound full with no. 22, 3-turn link, shielded (see coil form info under T201)

fig. 1. Front end of the vhf fm receiver. First rf stage neutralization adjustment (L202) will depend on individual transistor characteristics. Neutralization isn't required in grounded-gate second stage.

coupled to the base of the second mixer, Q102, where it is mixed with 10.245 MHz from the second local oscillator, Q103. The difference frequency, 455 kHz, is fed to a three-stage filter consisting of T102, T103 and T104. The selectivity depends on the coupling capacitors and also the tuning of the transformers.

Other types of filters are available that have better bandpass characteristics, but their cost and limited availability prohibit their use. By using i-f transformers as a filter, they can be adjusted for either wide- or narrow-band use. The i-f transformers are the type used in transistor radios.

The second i-f amplifier, U101, is an RCA CA3012 integrated circuit, which is low priced and has approximately 65 dB gain with good limiting characteristics. Because of the very high gain, the lead dress is important, and the bypass capacitors should have short leads to prevent self-oscillation. A problem developed on the original printed circuit layout with this stage due to ground loops, which caused U101 to oscillate.

Transformer T105 couples the i-f amplifier output to the fm detector U102, which is a Sprague ULN 2111A. This integrated circuit contains an i-f amplifier, limiter, fm detector, and an audio stage. Other IC fm detectors are on the market, but this unit doesn't require an expensive transformer and provides excellent a-m rejection.

The audio output of the fm detector feeds the audio gain and squelch pots. Transistor Q105, a noise amplifier, drives Q106, the noise detector. Transistor Q107 is a dc amplifier, which is driven by the noise detector and is used to turn the audio squelch gate Q104 on and off by biasing the emitter-to-base junction on and off.

Audio from the squelch gate drives the audio output stage, which is a relatively new integrated circuit from Motorola. The MFC4000 is inexpensive, small, and provides 250 mW output. While this is not quite enough for mobile use, it provides more audio than you can stand with a 4- or 5-inch speaker in normal room conditions.

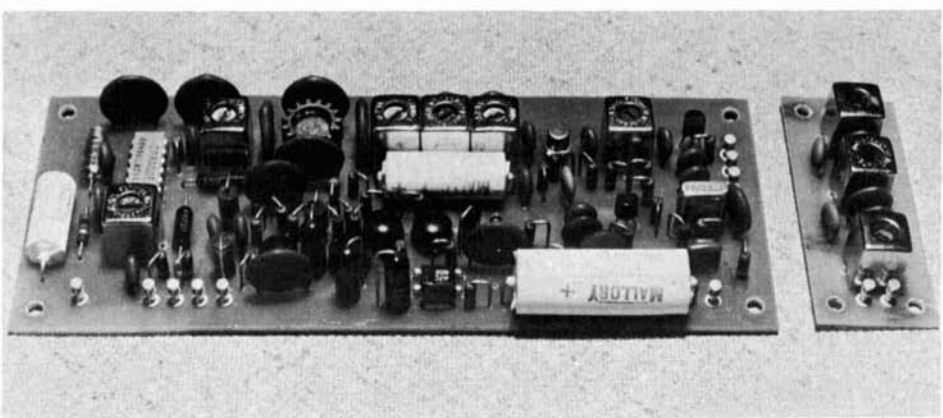
power supply

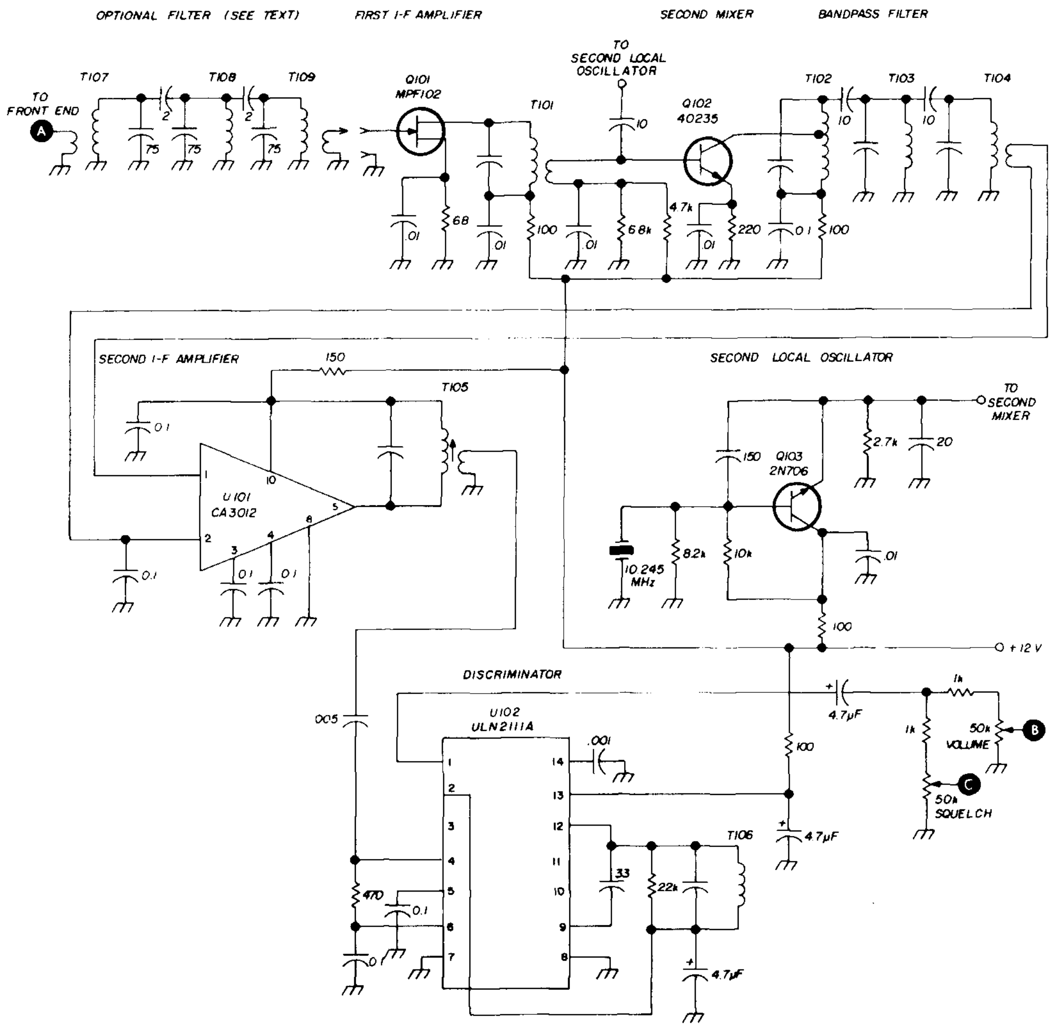
The entire receiver operates from a 12-volt source. Audio output stage U103 requires 9 volts, which is obtained from a zener. This method was used rather than a dropping resistor, because the current to U103 varies widely and so would the voltage drop. If the receiver is used for mobile work, some additional filtering of the battery supply may be desirable. The amount can vary from car to car. To use the receiver in the home, a small ac power supply can be made. One of the authors has poor line regulation, so an electronically regulated supply using a Motorola MC1460R integrated-circuit regulator was built to power the receiver. The receiver draws approximately 60 mA when squelched (no audio) to around 150 mA on speech peaks with the audio gain wide open. As far as mobile is concerned, since the power requirements are so low, many hours of listening can be had without straining the car battery.

alignment

Although an expensive fm signal generator of the type used for commercial radio work would be nice, it's not at all necessary. The i-f board can be initially tuned using any 10.7-MHz signal generator with a-m modulation. Some may question this; however, if the generator output is kept low as well as the percentage of modulation, the receiver will respond if the level is low enough to prevent limiting.

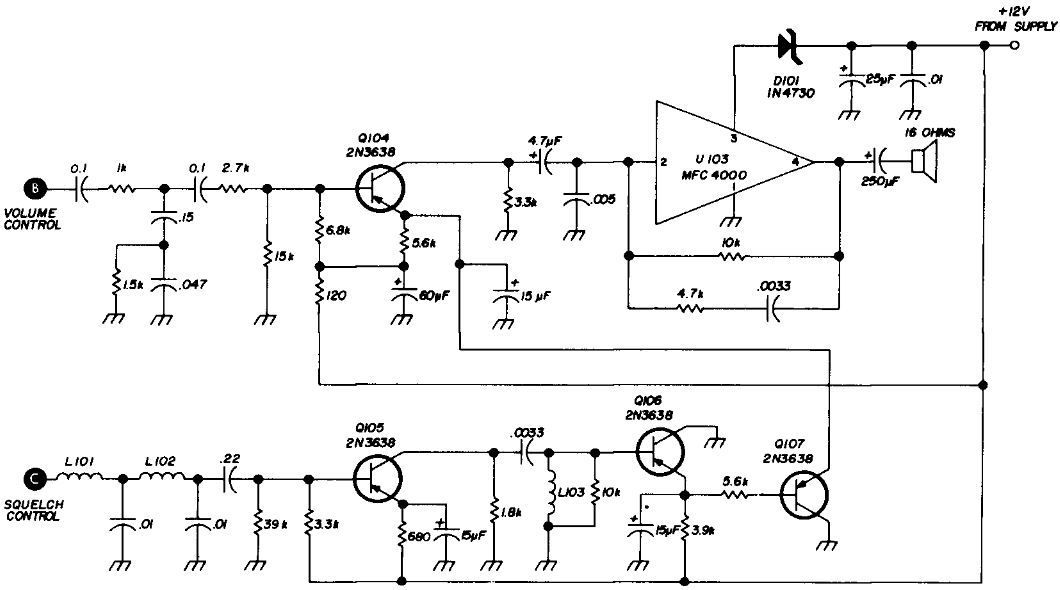
The large board (approximately 6 x 2-7/16 inches) holds the i-f amplifiers, discriminator, audio and squelch circuitry. The smaller board at right holds T107, 108 and 109. The layout is compact but not miniaturized.





To start, inject 10.7-MHz to the gate input of transistor Q101. Since we need a dc return, (normally provided by the link on T201) a 2.5-mH rfc should also be connected from the input to ground. The generator coupling should be through a capacitor of approximately 100 pF. Peak all transformers for maximum audio output, but be sure to keep the generator output low to prevent limiting. Next, connect the front end to the i-f board. (Be sure to remove the rfc.) Place an rf probe connected to a vtvm at the diode (D201) cathode, and tune T202 for maximum reading. Next, couple a gdo or wavemeter to L206 and tune C218 for maximum. Then couple a signal at the operating frequency to the input of the front end. Peak L201, L203, L204 and L205 for maximum audio output. If Q201 oscillates, adjust neutralization coil L202. Turns may have to be removed or added, depending on the individual characteristics of Q201 and because of the limited tuning range of the coil.

The above has been a preliminary adjustment. A cooperating station on frequency can be used for final on-the-nose alignment. If an fm generator and sweep generator are available, by all



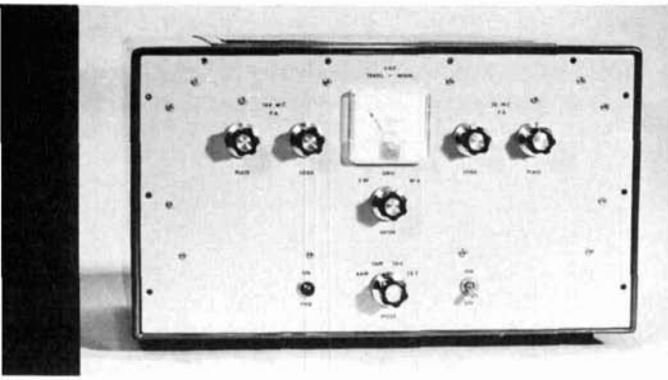
D101	3.9 volt zener diode, 1 watt (1N4730)
L101,L102	5.6 mH (Nytronics SWD-5.6 or equivalent)
L103	82 mH (Nytronics SWD 82000 or equivalent)
T101	CTC 3624-2 coil form wound full with no. 32, 3-turn link, shielded (winding area of CTC 3624 is 0.183" long, 0.210" winding diameter)
T102,T103 T104,T105 T106	Miniature 445-kHz i-f transformer, 25k—600 ohms (J. W. Miller 9-C1 or equivalent)
T107,T108 T109	CTC 3624-2 coil form wound full with no. 32, 8-turn link (see coil form info under T101)

fig. 2. I-f strip, discriminator, and af circuits. Bandpass filter following mixer consists of ordinary i-f transformers, allowing adjustable selectivity.

means use them; however, an on-the-air signal is adequate. Peak all coils in the converter as well as in the selectivity strip (if used) and transformer T201 in the i-f strip for maximum audio output. If an on-the-air signal is used, attenuation may be necessary at the antenna input to keep the receiver from limiting. A weak signal is best. Transformers T102, 103, and 104 are stagger-tuned for desired bandwidth. Using a properly modulated on-the-air signal, tune these transformers for minimum distortion and clipping. T105 is adjusted for maximum audio output. T106 quadrature coil is adjusted for best audio quality. If a scope is connected to the audio output of U102, T106 should be adjusted for a symmetrical waveform. If more than one frequency is used, tune the rf coils for approximately the center frequency spread. All that's necessary now is to adjust the volume and squelch controls for desired levels.

One of the authors has been using this receiver for mobile operation daily for several months and has found that it performs quite well. The mobile transmitter is a hybrid affair, but work is under way on a 100-percent solid-state transmitter. We will present this to *ham radio* readers at a later date if sufficient interest is shown. There is no reason why anyone who wants to get on vhf fm today can't build a receiver such as we have presented and produce a unit that will perform to expectations at reasonable cost.

ham radio



a multimode transmitter for six and two meters

PC-board mixers featured in an earlier article are combined with linear amplifiers and control circuits for improved operation

D. W. Bramer, K2ISP, 45 Thayer Road, Fairport, New York 14450

A previous article in ham radio featured transmitting mixers for the two- and six-meter bands.[1] Regular low-frequency equipment provided excitation for the six-meter unit, while a 50-MHz source provided drive for the two-meter mixer.

Although these converters produce low output power, they are sufficiently complete to form basic subassemblies for a medium-power dual-band vhf transmitter. This article shows how to combine these printed circuit assemblies with a bias supply, control circuit, and linear amplifiers. Construction techniques will depend on individual requirements, so I've only highlighted physical details; these are shown in the photos and sketches. Circuit details are shown in the schematic **(fig. 1)**.

Operating modes can be selected with a single control:

1. 6-meter a-m, using a low-power 50-MHz exciter such as the Heath "Sixer."

2. 2-meter a-m, using the same exciter as above.

3. 6-meter ssb, using a regular low-frequency ssb transmitter or transceiver.

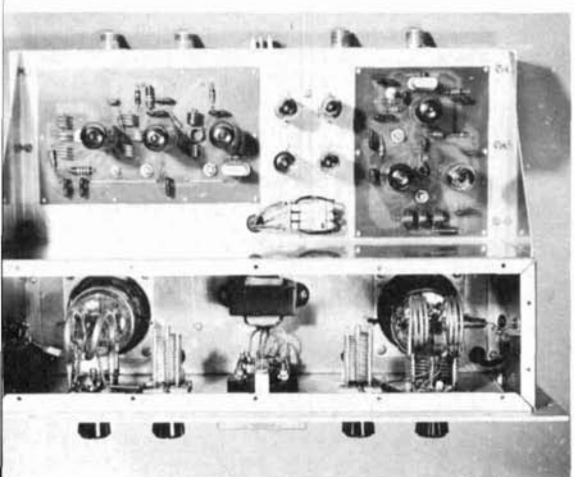
4. 2-meter ssb, using the same low-frequency ssb source as above.

Carrier insertion allows cw operation in both mode switch positions. Break-in cw and vox are included to enhance enjoyment.

The Amperex 5894, which is interchangeable with the 829B, may be more efficient, particularly on two meters. However, if the 5894 is used some reduction will be required in neutralizing capacitance. Also, the grid-circuit inductance will have to be increased.

metering system

With the selector switch in the center position, grid current of either amplifier can be monitored. In the counter clockwise position, 2-meter output power will be indicated; 6-meter output is indicated when the switch is in the clockwise position. No grid current should flow in Class AB_1. However, an indicator is required to show when grid current begins to flow to obtain maximum output with minimum distortion.

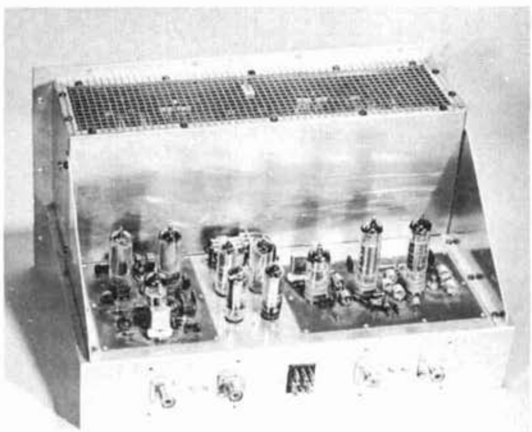


Chassis layout of the 6- and 2-meter *multimode transmitter*. The 2-meter transmitting converter is rear left; the 6-meter transmitting converter is right rear. The 829B final-amplifier stages are in the shield compartment to the front.

power supplies

Regulated +210 volts are required for the 6U8 oscillator-buffer, 5763 screens, and 12AT7 plates. Regulated Screen voltage for the 829B's is also required. Two independent series strings of OB2 regulators are used. This provides isolation between the oscillator and 829B screen-current fluctuations. A simple bias supply is used. A small 6.3-volt transformer is connected backward in the filament line. Its unloaded output (−145 volts) is applied through two resistors to 18- and 22-volt zeners. When the control relay is unenergized, both diode supplies float, allowing the bias lines to assume full

Rear view of the multimode transmitter shows the neat layout and construction used by the author.

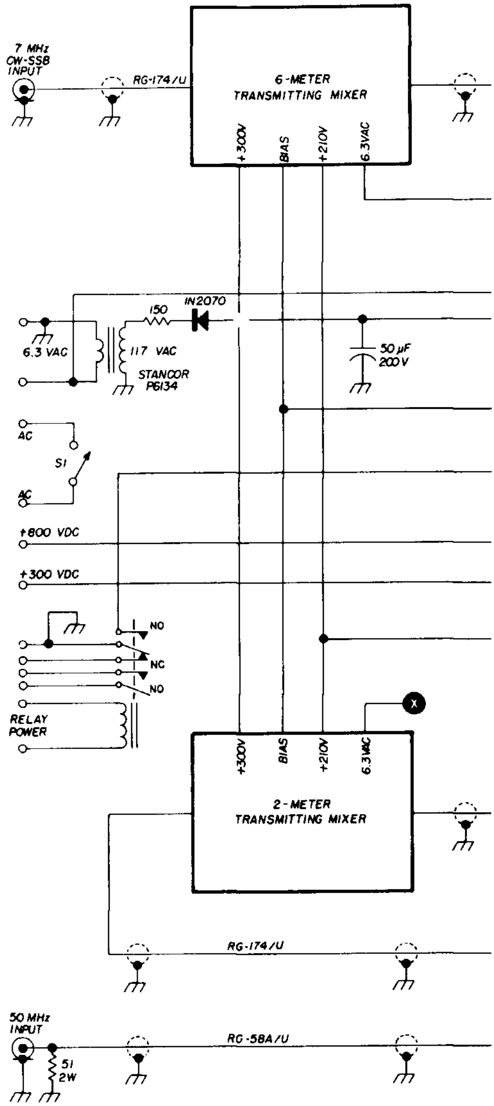


muting potential of -145 volts. In the energized position, the zener circuit is grounded, providing -18 and -22 volts bias for the mixers and 829B grids respectively.

Supply voltages are brought in via a 12-prong male Cinch-Jones plug mounted on the chassis rear apron. Supply requirements include +800 Vdc, 150 mA; +300 Vdc, 275 mA; and 6.3 Vac, 6.5 A.

construction notes

My construction techniques are apparent in the photos. Three major structures are used; an L-shaped main chassis plate, a U-shaped rf shield for the final amplifier compartment, and a 1/8-inch-thick front panel. Rough dimensions are shown in **fig. 2.** Angular side supports can be added, as shown, to strengthen the assembly. Except for the front panel, all pieces are made from 0.05-inch-thick 5052-H34



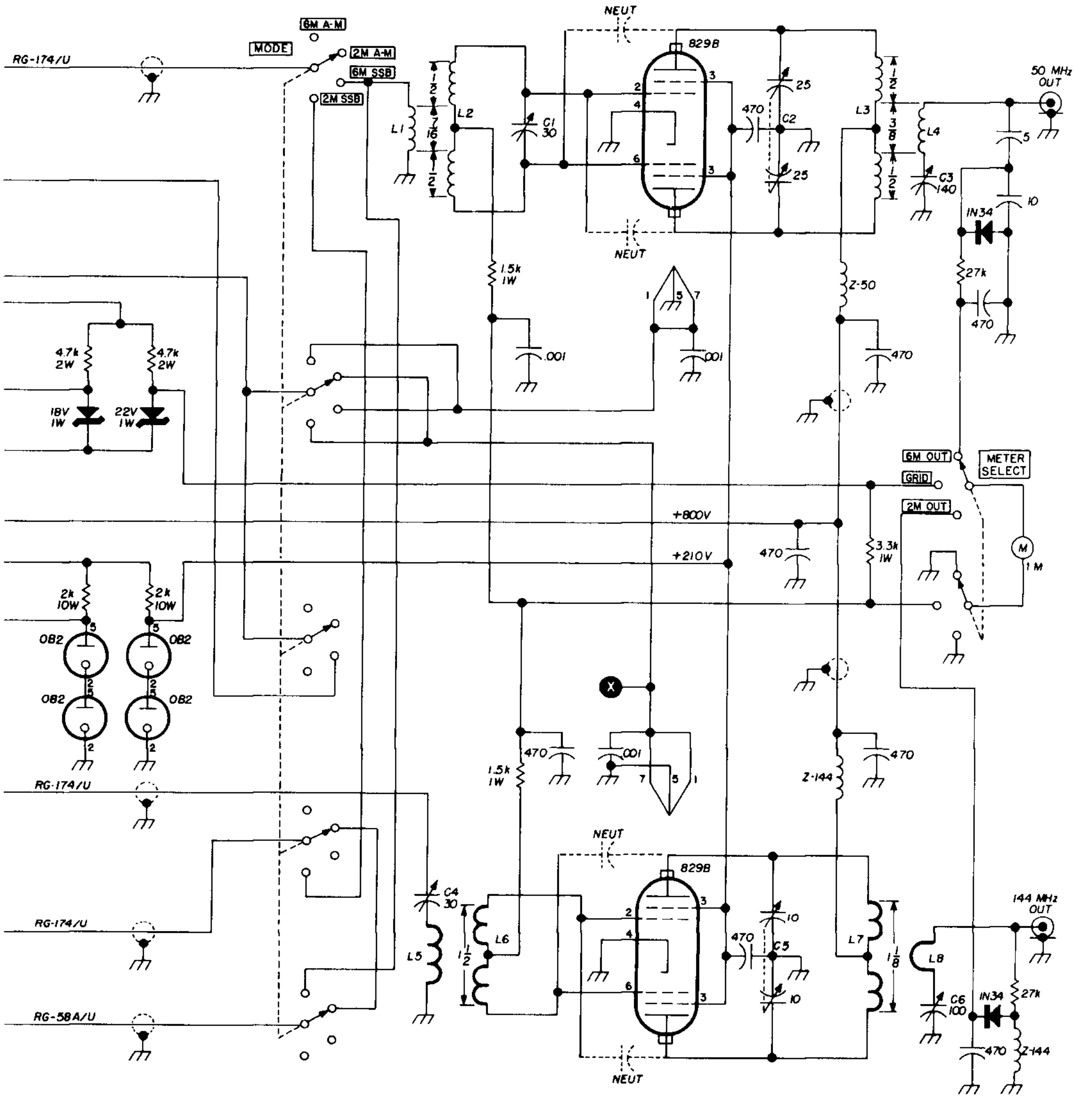
C1, C4	**30-pF trimmer (Arco 461)**
C2	**25-pF butterfly variable (E. F. Johnson 167-22)**
C3	**140-pF air variable (Hammerlund HF-140)**
C5	**10-pf butterfly variable (E. F. Johnson 167-21)**
C6	**100-pF air variable (Hammerlund HF-100)**
L1	**5 turns no. 16, 7/16" diameter**
L2	**6 turns no. 16 each side of center, 5/8" diameter**
L3	**3 turns no. 10 each side of center, 1¼" diameter**
L4	**4 turns no. 14, 7/8" diameter**
L5	**2 turns no. 16, ½" diameter**
L6	**3 turns no. 12, 5/8" diameter, 1½" long, center tapped**
L7	**2 turns 3/16" silver-plated tubing, 1¼" diameter, 1-1/8" long, center tapped**
L8	**1 turn loop around center of L7. Refer to photo for approximate size and position.**

fig. 1. Schematic diagram of the 6- and 2-meter multimode transmitter.

aluminum. An aluminum angle, ½ x ½ inch, was drilled and tapped to form the fourth top surface edge of the rf shield compartment. I tailored the front panel to accomodate the assembly in a Heath "Seneca" cabinet, which I obtained from a surplus outlet. The knobs are from Heath.

Final amplifier components, panel meter, meter switch, and the bias supply transformer are mounted above deck inside the shield compartment. The control relay and voltage regulator tube sockets are mounted near the chassis rear center, outside the shield compartment.

The 829B sockets are E. F. Johnson type 122-105-100. The circular portions extend well below the chassis. Grid coils are suspended from the socket terminals with number 14 insulated solid wire leads. These are criss-crossed and extend up through ceramic wafer holes to form

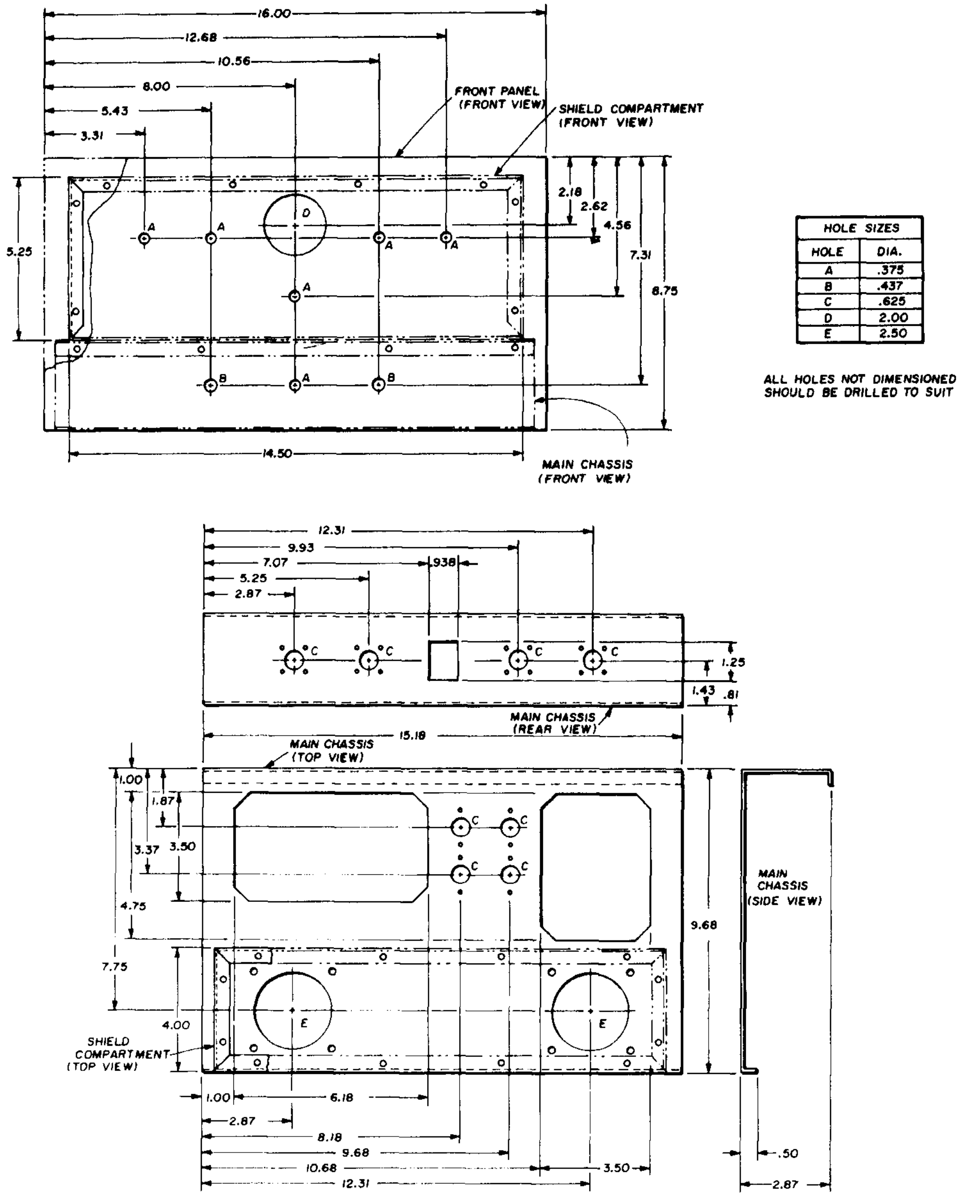


neutralizing capacitors. The leads should run about one-half inch above the chassis surface. Capacitance is adjusted by bending the leads.

Link input coils, 1.5 kilohm grid resistors, and the bypass capacitors are suspended from a small terminal strip soldered directly to the base of each socket between pins 1 and 7. High voltage for the 829B tank is brought above deck with RG-58/U coaxial cable. Fahenstock spring clips, soldered directly to butterfly variable capacitors, provide 829B plate pin connection.

adjustment and operation

Reference 1 should be reviewed before attempting to set up this more complex system. If the printed-circuit subassemblies have been preadjusted, the remaining task is to stabilize the 829B stages and optimize their input coupling. Reference



HOLE SIZES	
HOLE	DIA.
A	.375
B	.437
C	.625
D	2.00
E	2.50

fig. 2. Mechanical construction details of the 6- and 2-meter transmitter chassis.

to the ARRL Handbook will be helpful in obtaining correct adjustment.

First, apply heater power only. The appropriate heaters should light with the mode switch in each position. Next, check for -145 volts on the 829B grid and mixer-board lines. With the relay armature depressed, -22 and -18 volts respectively should appear on 829B grid and mixer lines.

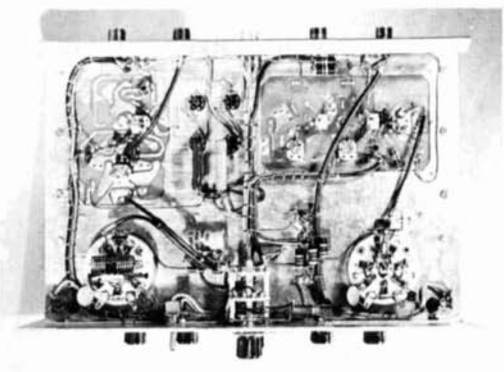
two-meter a-m adjustment

Connect a 50-ohm dummy load to the 50-MHz output jack. Position the mode switch to "6 AM." With heater power and drive applied, the 829B stage can now be neutralized. (A standard procedure is given in the ARRL Handbook.)

Apply 300- and 800-volt dc power. (It might be helpful to reduce the 800-volt power to 500 volts for this initial adjustment.) Adjust the controls for maximum power into the dummy load. This can be observed on the panel meter with the meter switch in the "6M OUTPUT" position. Peak trimmer capacitor C1 for maximum output indication on the meter.

two-meter a-m adjustment

Connect the 50-ohm dummy load to the 144-MHz output jack. Leave the 50-MHz exciter connected. Rotate the mode selector switch to the "2 AM" position and the meter switch to "2M OUTPUT." Neutralize the 829B stage. Apply dc power and tune for maximum rf power into the dummy load. Alternately adjust capacitor C4 and the two-meter mixer board output capacitor (butterfly variable) for maximum 144-MHz output.



Under-chassis shows good vhf construction.

six- and two-meter adjustment

Connect suitable excitation to the low-frequency input jack. Rotate the mode switch to the "6 SSB" position. With the meter switched to read "6M OUTPUT," gradually increase drive for 50-MHz output indication. Peak the 6-meter mixer-board output tank butterfly variable capacitor for maximum indication. Position the meter switch to read grid current, and increase drive until an indication of grid current is just apparent. This represents the maximum drive level at which maximum output will occur consistent with minimum distortion.

The unit is now adjusted for the "2M SSB" operating mode. Now rotate the mode switch to the extreme clockwise position, and adjust the controls for maximum output while observing the drive level.

Rear panel of the multimode transmitter.

final notes

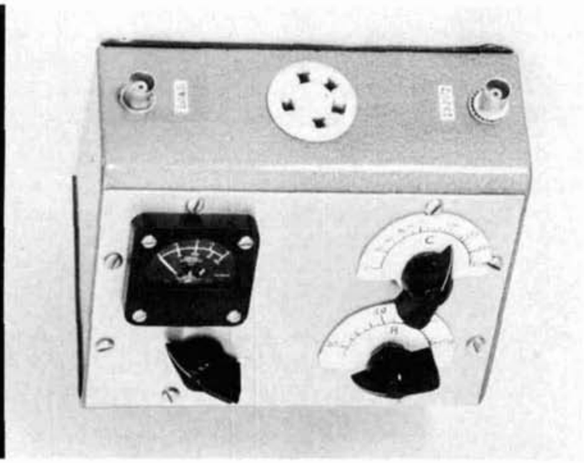
It may be necessary to alter the value of the resistors in the output sampling circuit. Try different values near 27 kilohms until a good meter deflection is obtained at maximum output.

For a-m operation, tune for maximum output in the ssb mode first, then switch to the a-m mode position. The 50-MHz a-m input source must be attenuated to produce about one-half the ssb-mode output as indicated on the meter. Modulation up to 100 percent will then be amplified linearly. Peak envelope ssb output is about 50 watts, while output on a-m is about 12 watts on either band. Third-order distortion products on 144 MHz measure about 30 dB down.

reference

1. D. W. Bramer, K2ISP, "Heterodyne Transmitting Mixers for Six and Two Meters," *ham radio*, April, 1969, p. 8.

ham radio



a simple bridge for antenna measurements

This instrument allows measurement of both resistive and reactive components of your antenna with the aid of the Smith chart

Henry S. Keen, W2CTK, 64 Schuyler Drive, Commack, New York 11725

Articles published in the amateur literature on the subject of the Smith chart[1] have been of more academic interest than practical value. How many hams, for example, have slotted-line and swr—indicator facilities appropriate to the frequencies involved? The little device described in this article can be built by anyone and will provide useful data for antenna measurements based on the Smith chart, even for the "dc bands."

basic bridge circuit

The need for more accurate evaluation of a recently erected 15-meter quad was indicated when a borrowed vswr meter provided a value that was obviously too good to be true. The bridge described here was the outcome and has provided much useful information.

The basic bridge circuit **(fig. 1)** will measure a purely resistive load with acceptable accuracy; but when a reactive component is present, the null will not only be broad, it will be displaced and therefore inaccurate.

Several methods are available for balancing and measuring the reactive component, but the simplest appears to be placing a parallel-resonant tuned circuit across the load, as shown in **fig. 2**. By detuning this circuit from resonance, it's possible to introduce a controlled opposite reactance across the load and adjust the bridge to a perfect null. The reactive component of the load is measured by the direction and degree of this detuning. The equivalent circuit of the antenna therefore is represented by a resistance shunted by a reactance. Because these elements are in parallel, it's probably more accurate to regard this bridge as an admittance rather than an impedance bridge, and to perform any computations on that basis.

Two arms of the bridge are made up by the 100-ohm linear potentiometer. A third arm is a 51-ohm, half-watt carbon resistor, while the fourth arm is the load, shunted by the parallel-tuned circuit.

The variable capacitor should be at least 250 pF for most applications, while the switched coil sections (see photo) are merely sufficient in number that resonance on each band can be obtained on at least two settings of the selector switch. This provides for adjustment in either direction from resonance when a complex load is present.

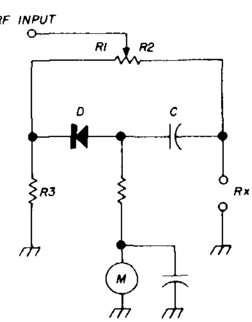


fig. 1. The basic rf bridge. At balance, R1/R2 = R3/RX.

A germanium rather than a silicon diode was selected because of its lower barrier potential. Although a 100-μA meter was used, its choice was dictated more by small size than current rating. For greater sensitivity, the 1.5k isolating resistor can be replaced by an rf choke.

calibration

No measuring device is better than its calibration. Many hams may feel they have too little precision equipment with which to calibrate a device of this type. However, a nice thing about this bridge is that the *resistive and reactive controls can be calibrated separately*. Before wiring the bridge, the resistive control can be calibrated against a purely resistive load. The resistive calibration can even be made with a couple of flashlight cells using the circuit of **fig. 3**. A handful of 100-ohm, half-watt resistors used in various combi-

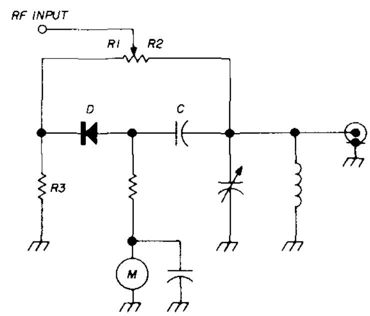


fig. 2. Modification to measure complex impedances.

nations should provide sufficient useful calibration points. If you use as many resistors as possible (within reason of course) in these combinations, individual differences will tend to cancel.

Calibration of the capacitor is greatly simplified if you have access to a Q meter. Circuit capacitance, and the minimum capacitance of the variable capacitor itself can be ignored, as we are interested only in the incremental accuracy; i.e., the difference in capacitance between any two settings of the dial. A handful of silver-mica capacitors of known value can be used with a grid dipper for the calibration, ignoring any points that don't lie on a smooth curve. A preliminary calibration by this substitution method was later repeated, using a precision decade-resistor box and a Q meter. The difference between the before-and-after measurements were well within 10%, which is a good working value around a ham station.

measurement procedure

A necessary device for this application is a reasonably well-matched 50-ohm load. A half-watt, 51-ohm resistor, mounted in a BNC connector performed quite well.

At a given frequency, the rf input to the bridge is first adjusted to produce full-scale deflection of the meter with the controls set at either extreme. The bridge is then balanced at this power level against the reference load, and the capacitor reading is recorded. The antenna is then substituted for the reference load, and the bridge is rebalanced. These resistive and capacitive readings are recorded, and the process repeated on other frequencies of interest.

To compute the equivalent antenna impedance, (or admittance) a table is set up as in **table 1**. For each frequency the setting of the capacitor that was obtained with the reference load in place is listed, then the resistive and capacitive readings with the antenna in the circuit. The two capacitive readings for each frequency are compared and their differences listed. The reactance, which must be computed for each frequency (represented by this differential capacitance) is then listed in the next column.

Normalized conductance and susceptance are next computed for a 50-ohm reference level by dividing 50 by the measured resistance and the computed reactance of the capacitive differential. If you're using 70-ohm line, you should substitute 70 for 50; otherwise all procedures are the same. If the setting of the capacitor was reduced by substitution of the antenna for the reference load, the susceptance is positive (capacitive). If it was increased, the susceptance is negative (inductive).

Table 1. Data for computing equivalent antenna impedances.

Freq. (MHz)	R (ohms)	C (pF)	Ref C (pF)	ΔC (pF)	X	G'	B'
21.0	60	20	75	55	137.6	.833	.363
21.1	61	26	72	46	163.8	.820	.325
21.2	64	30	73	43	175.0	.780	.285
21.3	68	29	72	43	174.0	.737	.287
21.4	76	24	71	47	158.0	.659	.315
21.5	81	14	70	56	132.0	.618	.379

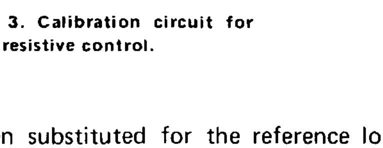


fig. 3. Calibration circuit for the resistive control.

These normalized admittances can now be plotted on a Smith chart, as in **fig. 4**, and successive frequency points connected by a smooth curve. This curve showed my quad to be resonant at about 21.15 MHz. As I normally operate ssb at about 21.35 MHz, the resonant point represented about 4 or 5 inches of wire in the length of the driven element.

smith chart

To convert your admittance measurements to more familiar impedance values, all that's necessary is to replot each point on the Smith chart at the same distance from the center of the chart, but located diametrically opposite the admittance points. When making this transformation, bear in mind that when we deal in admittances the equivalent circuit is a resistive and reactive element in parallel; *while the equivalent circuit for impedance* is a resistive and reactive element in series. To return to absolute values from the normalized quantities on the Smith chart, merely multiply the impedance components by 50 or 70, as the case may be.

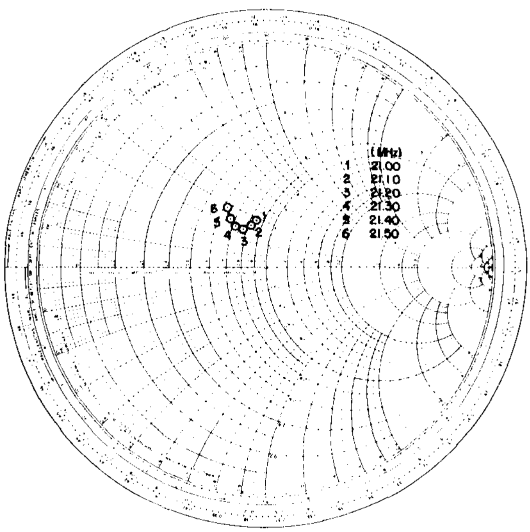


fig. 4. Admittance plot of 21-MHz quad as measured through the 50-ohm coaxial line.

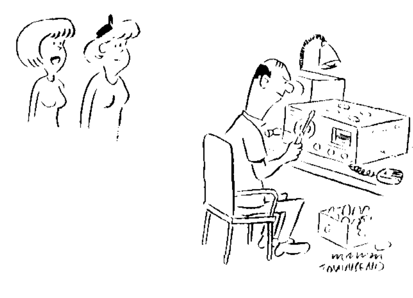
The distance of the plotted point from the center of the chart **(fig. 4)** indicated that the vswr at resonance, (the points closest to the center, in this case) is about 1.5.* Although this is a reasonable figure, the match can be improved in several ways. First, and most practical, is to replace the 50-ohm cable with 70-ohm cable. Second, the spacing between antenna reflector and driven element can be reduced until the antenna input impedance more closely approximates 50 ohms; and third, a matching network of some kind can be installed at the antenna end of the coaxial line.

Occasionally one hears of someone who trimmed the length of his coaxial line "to make it load." What probably happened was that the antenna was so badly mismatched that, coupled with an odd length of coaxial line, a shunt reactance appeared across the output network that was so large the network just couldn't compensate for it. Trimming the coaxial line didn't reduce the vswr, but it did reduce the reactive component to where the overworked matching network was finally able to handle the situation. If you're forced to this quick and dirty solution, at least the Smith chart will tell you exactly how much transmission line to remove. In **fig. 4** it can be seen that *moving the impedance plot* about .092 wavelength nearer the load should remove most of the reactive component.

transmission-line length

A fair representation of what the antenna looks like to the feed line may be obtained by transferring each admittance (or impedance) point toward the load (counter clockwise) a distance equal to the electrical length of the feed line in wavelengths at each frequency. Each half-wavelength represents a complete revolution around the center of the Smith chart. This operation assumes the line to be lossless, frequently an optimistic assumption!

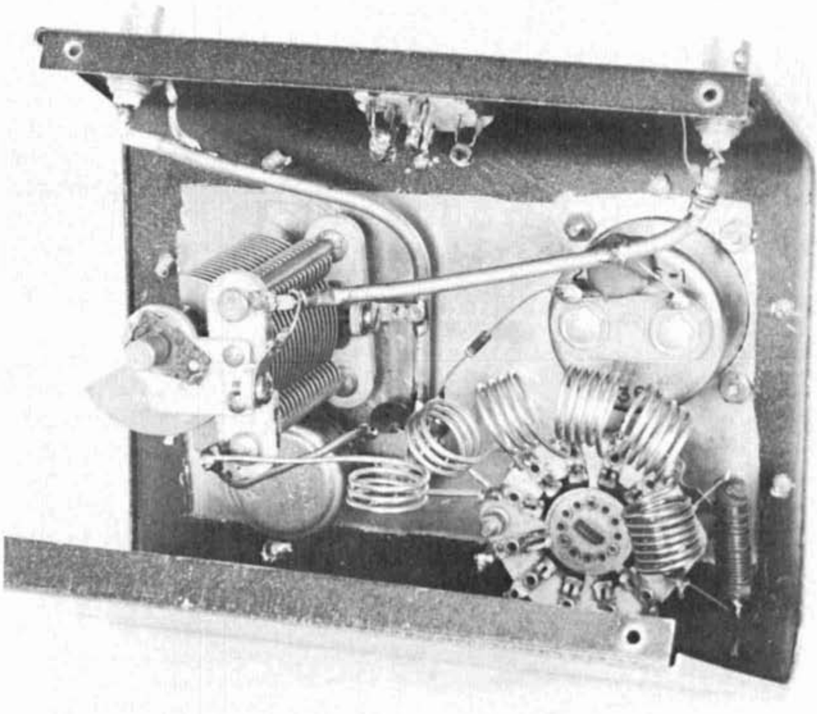
However for this move, the exact electrical length of the transmission line in wavelengths must be known. One method is to disconnect the center con-



"George is one of the silent majority . . . He can't get his transmitter to work."

*Because the circle through the resonant point intersects the R/Z_O axis on the right-hand side of the chart at $R/Z_O = 1.5$. **Editor.**

ductor at the antenna and grid-dip the line from the bottom end, coupling to a small one-turn coil. Above all don't trust the frequency calibration of the grid dipper, but check with a good wave-meter or calibrated receiver. If the transferred measurement points don't center on the scaled diameter of the Smith chart, your electrical length was inaccurate. Grid dipping to the odd harmonic nearest the operating frequency is probably best, but be sure you know which harmonic it is. The electrical length of the feed line is proportional to the frequency.



Inside the admittance bridge. Switched coil sections resonate on each band of interest.

conclusion

Transferring the admittance or impedance data up to the business end of the feedline in this way will give you a piece of information otherwise hard to come by. Suppose, for example, an antenna shows a vswr of 2:1. You can't always be sure the antenna looked like 100 ohms, for it could as well appear as 25 ohms to the 50-ohm cable. You'll therefore be able to tell in which direction the loading adjustment should be made.

Although this type of antenna measurement may seem to involve considerable pencil pushing, it results in much better than a ball-park estimate. Those microvolts you're able to put into the other follow's receiver are expensive and hard to come by. It's nice to feel that you're getting everything possible from your installation because you did the job thoroughly, rather than settle for the too-frequent lick and a promise of which so many are often guilty.

reference

1. F. E. Terman, "Electronic and Radio Engineering," 4th ed., McGraw-Hill, New York, p. 100.

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neutralizing small-signal amplifiers

How to obtain maximum performance from preamps and converters

Earnest A. Franke, WA4WDK, 108 Matawan Terrace, Matawan, New Jersey 07747

In most of the converters and preamplifiers built today the tubes or transistors are operated in the grounded-cathode or grounded-source configuration. Neutralization of the amplifier is necessary to compensate for the interelectrode capacitance in the tube or transistor. If not neutralized, this capacitance allows energy developed in the output circuit to be fed back to the input. The amplifier then acts as a tptg oscillator.

Amateurs unfamiliar with the theory of this phenomenon frequently encounter instability problems and find amplifiers hard to neutralize. This article provides basic data on neutralization and gives some hints on how to solve this problem. A properly neutralized amplifier will reward you with maximum gain and stability consistent with minimum noise figure, which is especially important at the higher frequencies.

gain measurement

Amplifier gain can be measured in the usual manner, from input to output. Gain in the reverse direction can be measured also, and should be zero or nearly so. **Fig. 1** illustrates this concept.

The interelectrode capacitance in the amplifier allows a signal to be fed back from output to input. As the feedback capacitance is increased, the reverse gain also increases. Thus it is more important to neutralize amplifiers at the higher frequencies because of the decreasing capacitive reactance.

neutralizing methods

When neutralization is required over a small bandwidth (as in an i-f amplifier), a resonant system can be used. An inductor, L_n, is made to resonate with the interelectrode capacitance and any shunt capacitance existing in the circuit **(fig. 2)**. The inductance value can be approximated by

$$L = \frac{1}{4\pi^2 f^2 C}$$

where f is the mid operating frequency, and C is the grid-plate or gate-drain capacitance of the tube or transistor. A large blocking capacitor, C, is inserted in series with the inductor to prevent shorting the supply voltage to ground through the input circuit. This blocking capacitance must be large enough to offer a negligible reactance at the operating frequency. Button capacitors are excellent because of their short lead length. Air-wound inductors are preferred for the neutralization coil, because they have less stray capacitance than the slug-tuned type. The neutralization point may be found by using a tuning wand with a brass or iron slug at each end. Proximity of the iron slug increases the neutralizing coil inductance, requiring the coil to be squeezed. Brass decreases the inductance, requiring the coil to be expanded to reach resonance.

At resonance the impedance of a parallel circuit is a resistance that is Q times the reactance of either the inductance or capacitance.[1] Very high impedances can be developed by parallel resonance. Therefore, a higher-Q neutralization circuit will result in a larger isolation, as seen in **fig. 3**. As gain or bandwidth increases, the Q or impedance of the inductance must be improved to isolate the input and output circuits. Shunt capacity to ground must be avoided, which reduces the gain-bandwidth product.[2]

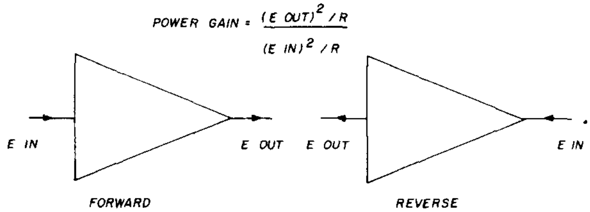

fig. 1. Amplifier gain can be measured in forward or reverse direction. In the equation for power gain, R is the tuned-circuit impedance.

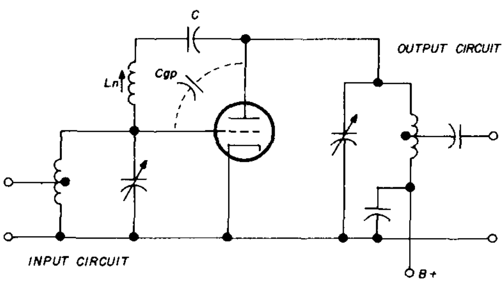

fig. 2. Neutralization scheme for an amplifier operating over a very narrow bandwidth or at a fixed frequency. This is known as the coil neutralizing method.

Neutralization is necessary to reduce tuning interaction between input and output. Many amateurs simply tune the neutralization coil in the amplifier until the output circuit doesn't affect the input. When the "blurps and squeals" disappear, neutralization is considered to be complete.[3]

tube neutralization

When tubes are used, neutralization is simplified because the plate voltage is removed. The neutralizing coil is adjusted for a strong signal applied to the input. At the minimum output point, the coil resonates with the grid-plate capacitance, and circuit impedance is maximum. A simple neutralization procedure can be used, requiring a strong local signal to be fed to the input and a detector coupled to the output.

fet neutralization

Field effect transistors are being used more today because of their low noise figure, ability to accept large signals with small cross modulation, and low cost. Inside the fet is a capacitance, C_{rss}: the common-source, short-circuited, reverse-transfer capacitance. This capacitance between the gate and drain is a parallel combination of a voltage-sensitive depletion capacitance and lead capacitance. The behavior of the depletion capacitance is similar to the variable-capacitance effect obtained by reverse biasing semi-

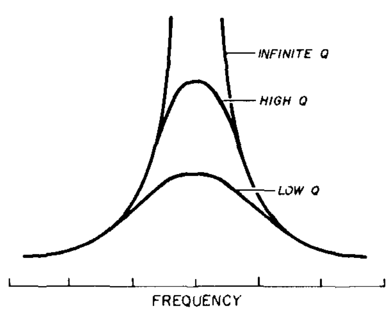

fig. 3. Impedance as a function of frequency with circuit Q as a parameter (parallel-resonant).

conductors.

Differing from the tube method, attempts to neutralize the fet by minimizing a signal applied to the input without the drain supply voltage applied will not work. The junction capacitance changes significantly when the supply voltage is applied.

mosfet and grounded-gate circuits

Neutralization usually may be avoided by operating the amplifier in the common-gate mode. The grounded gate acts as a radio frequency shield between input and output. Feedback capacitance is no longer the gate-drain capacitance, but is the source-drain capacitance.

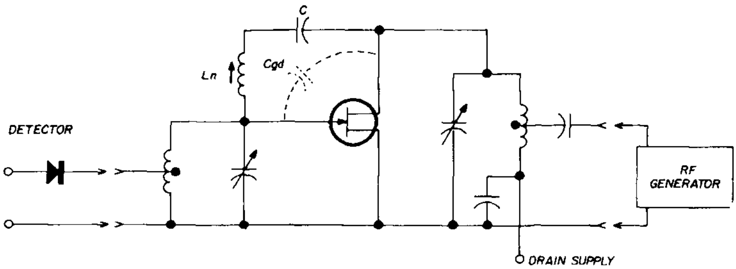

fig. 4. Method of measuring reverse gain in a typical amplifier.

neutralizing with applied voltage

The following method can be used to neutralize a tube or fet amplifier with the supply voltage applied. An rf signal generator or sweep oscillator is coupled to the output circuit, as shown in **fig. 4**. Generator level is increased until an output is noted by a sensitive detector or receiver connected to the input circuit. The neutralizing coil, L_n, is tuned for a minimum indication at the detector. The amplifier is "analyzed backwards" in this method, and reverse gain is measured. If a sweep generator is used, the oscilloscope presentation would resemble **fig. 5**. A "suck out" will occur at the neutralized frequency,. caused by the isolating high-impedance resonant circuit. The bandwidth of the "suck out" will be determined by the Q of the resonant neutralizing circuit and must be wide enough to cover the amplifying bandwidth of the amplifier in the forward direction.

Complete neutralization may still not be achieved if stray inductive or capacitive coupling exists within the amplifier components or extends to another stage. The answer to this problem is shielding or rearrangement of components. Proper bypassing, with short lead lengths, must be maintained to avoid oscillation.

With the mosfet semiconductor, any necessary neutralization is largely due to socket-lead capacitance. The feedback in the common-source configuration is less than that in a vacuum tube in the grounded-grid mode. You might get by

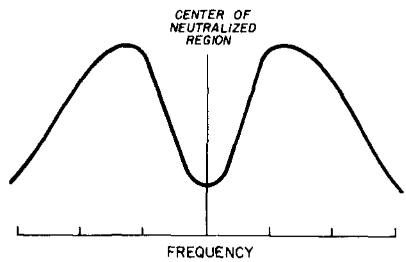

fig. 5. Oscilloscope display of "suck-out" at neutralization frequency.

without neutralization and still have a stable amplifier, but the noise figure and gain will be improved with it. Once tried, it is a simple procedure.

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1. Frederick Terman, *Electronics and Radio Engineering,* McGraw-Hill, p. 52.
2. Ibid., p. 415.
3. D. D. DeMaw, "FET Converters For 6 and 2 Meters," *QST,* May, 1967, p. 11.

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electronic counter dials

A direct frequency readout for your receiver vfo using inexpensive ICs

E. H. Conklin, K6KA, Box 1, La Canada, Claifornia 91011

One evening at the home of Stan Dixon, VK3TE, several of the VK3 gang were chatting about things of general interest when Harold Hepburn, VK3AFQ, mentioned that dials for accurate frequency indication were difficult to obtain in Australia. He was building a two-meter receiver using a 10–10.5-MHz vfo and wanted something reasonably accurate for frequency readout. He asked about the possibility of using a built-in electronic counter for the vfo that would indicate frequency directly.

This set me thinking about the problem. After considering the many angles, I concluded that this would be a feasible project, using inexpensive digital ICs. This was before any details of the *signal/one* receiver were available, which uses a digital dial readout that gives calibration accuracy of 100 Hz on each band.

conversion-frequency correction

All equipment won't necessarily have the convenient vfo range that VK3AFQ had in mind, but may begin at some odd frequency. Furthermore, heterodyne conversion oscillators will have some error (unless they're adjusted to exact frequency), which will be part of the readout.

This error may be different for different bands. Computer techniques could be used to remember the error and add or subtract the correction for the conversion frequency for each band, but a much simpler method is at hand.

If the bandswitch is provided with suitable contacts, the flip-flops in the frequency counter can be preset so that the counter will show the correct happening at the right time.

Some vfo's may tune up on some bands and down on others. This sounds like the last word in complexity for an electronic counter dial, but it's not impossible. By using gates that switch the circuit to count up or down, a single external contact can make the counter go either way. Several such circuits appear in references 2 and 3.

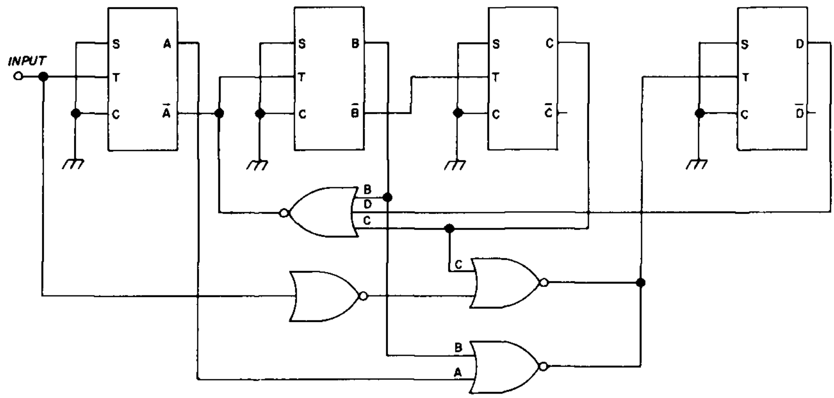

fig. 1. A decade divider that counts down instead of up. Gating is such that, on leaving zero state, the circuit returns to the 9 state directly by inhibiting the B flip-flops and causing the D flip-flop to toggle during this condition only.

frequency at the end of the dial. It would be necessary only to note the readout at the end of the tuning range, then wire the *preset* terminals of the counter ICs so that the additional count at the end-of-band vfo frequency would set the indicator to zero (000.0 kHz).

backward vfo's

Many vfo's tune backward: the highest vfo frequency is the lowest receiver frequency. Most counters count up, not down. It would be useful to avoid a computer subtraction process. Motorola's Application Note AN-251[1] states, " . . . a ripple counter will count down when the complemented output from one stage drives the clock input of the following stage. . ." Motorola's circuit showing how this is done is reproduced in **fig. 1**. Four NOR gates are added to keep events

time-base generator

A time-base generator will be required, as in other counters. If you would like to have an accurate frequency standard[4,5] it would be worthwhile to combine functions and use this accurate base for the electronic counter dial as well. However, if you wish to settle for a dial that displays only four digits (three for integral kHz and one for tenths of a kHz), some inaccuracy can be accepted. Two decades of the counter won't have any readout. You could use the 60-Hz power-line frequency to generate the time base. The time-base generator may require only four dual JK flip-flops for a one-second count. Hewlett-Packard, in the instruction book for their little four-digit counter, states that an accuracy of 0.02 percent or better is expected from the use of the power frequency for gating.

I have completed experimental work using the power-line frequency for gating. A report on this will be in a forthcoming article. It confirms that the inaccuracies, which are within 0.02 percent, will not affect a counter dial with two digits not displayed. It appears that you may expect an error of one or two counts (tenths of a kHz) occasionally in a four-digit dial indicator readout that displays kHz and tenths of a kHz in four digits. This should be good enough for tuning indicators.

display methods

It's possible to count vfo frequency, store the count, and display only the last completed count until it must be changed. (See the data on the Fairchild CL 9959 buffer-storage element and the CL 9960 decimal-decoder driver.) The count can remain unchanged until some later count requires the readout indicator to change to a new frequency. However, although storage, decoding, and lamp-driver ICs are available, we can get along with something even simpler.

Let's say we can use some of the available percent error to reduce the count period to 0.1 second rather than the more common 1-second interval. Without storing, or blanking the indicator during the count, we would then have 0.1 second during which the indicator runs through a count and gives no clear indication of frequency. This can be followed by nearly 0.9 second during which an unchanged frequency is displayed, even when the dial is being turned.

This action can be accomplished by using some of the ICs to produce a 0.1-second gating input, the total 1-second display period, and a preset signal before the start of the next count. As indicated earlier, this can preset the combined heterodyning frequencies for one end of the tuning dial, so that when the vfo frequency is counted at this point, the indicator will show 000.0 kHz.

The indicator can be any of the several types of digital displays. The less-expensive displays, such as the gas-filled National Electronics NL-950 shown in the Newark and Allied catalogs, cost about $6.25 per digit, or $25.00 for the entire four-digit display. For those who are satisfied with the binary-digital combination[5] and are willing to read each digit in an 8-4-2-1 combination using four lamps, a somewhat lower cost is possible.

conclusion

Now let's see what we have. Inside our set will be a PC board about 4 by 6 inches, plus some space for the decoding gates or lamp-driver transistors. The vfo bandswitch will have contacts that preset the counter ICs to the correct frequency for the end of the vfo tuning range.

Outside, the unit will have a suitable knob with a convenient tuning ratio, but no associated indicator. Above it will be either four indicating gas-filled tubes or four rows of four lamps each for binary readout.

When the unit is turned on, the error or drift in the crystal-controlled heterodyne conversion oscillator won't be corrected in the tuning readout, but vfo error will be. If the knob is turned rapidly, the indicator will show an approximate frequency at 1-second intervals, thus lagging a bit behind a rapidly turned knob. When the knob rests for about a second, however, the readout will show the frequency to an accuracy close to 0.1 kHz.

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2. "DCL Series 800 Handbook," May, 1968, Signetics Corporation, 811 E. Arques Ave., Sunnyvale, California 94086.
3. "Signetics Utilogic II Handbook," Signetics Corporation, 811 E. Arques Ave., Sunnyvale, California 94086.
4. E. H. Conklin, K6KA, "Amateur Frequency Measurements," *ham radio,* October, 1968, p. 53.
5. E. H. Conklin, K6KA, "Frequency Counters and Calibrators," *ham radio* November, 1968, p. 41.

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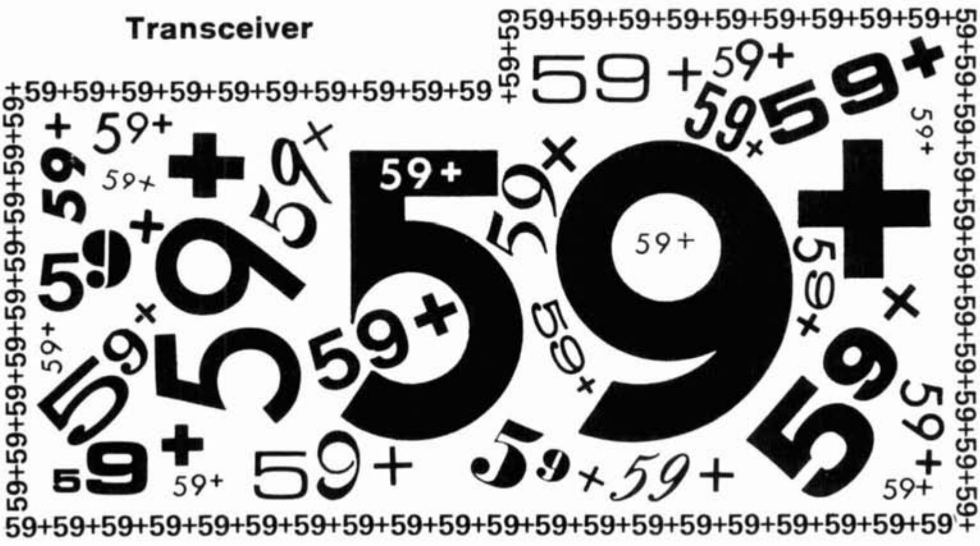

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solid-state audio oscillator-monitor

The basic for this efficient, low-power circuit is an isolated integrator network

N. J. Nicosia, WA1JSM, 85 North Street, North Reading, Massachusetts 01864

Prompted by a recent article in *ham radio,*[1] I fulfilled a long-time desire to design an audio oscillator that works with digital integrated circuit supply voltages, produces a clean sine wave at 1 kHz, drives a speaker, and is inexpensive and easy to build. The result is described in the following paragraphs.

circuit description

A minimum number of components is used in an efficient sine-wave oscillator circuit **(fig. 2)**. Transistors Q1 and Q2 form a high input and low output impedance amplifier, a feature of operational amplifiers. A "basic isolated integrator network"[2] is inserted between input and output. The circuit can be made to

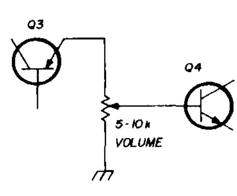

fig. 1. Optional circuit for volume control.

oscillate at the frequency determined by the network constants by omitting a resistor between the base of Q1 and ground. Slight adjustment in frequency

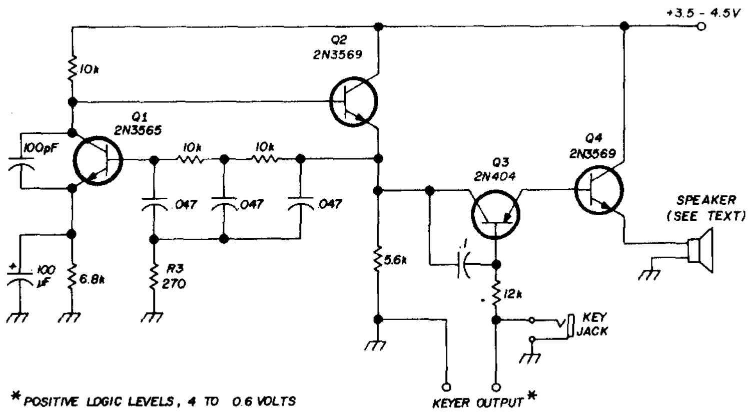

fig. 2. Schematic of the 1-kHz audio oscillator. Output frequency is determined by integrator circuit between Q1 and Q2.

can be made by changing the value of R3; however, don't make too big a change since stability may be affected. Q3 operates as a series switch and is turned on by grounding its base resistor. The transistors specified for Q3 are germanium types because of their low saturation voltage, which promotes efficient switching. Q4 is an audio power amplifier and drives a speaker directly.

I use a 3.2-ohm speaker in series with a 22-ohm resistor as a load. Any speaker will work, even the 40-ohm transistor-radio types. Use a scope to ensure a good waveform across the whole load. Q4 can be prevented from heating up by adding more resistance in series with the speaker. I don't use a volume control, since the level of my unit is perfect for my requirements. However, if one is desired, see **fig. 1**.

power supply

With a 4-volt supply and key down, the total power drain is 35 mA. For a power supply you can use three D cells in series or obtain 3.6 to 5 volts from your electronic keyer. A simple power supply is shown in **fig. 3**.

Logic levels of 4 to 0.6 volts are commonly encountered with keyers. I would be interested in hearing how this circuit can be tied into the many keyers described in the amateur literature.

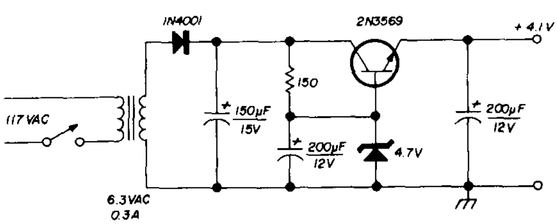

fig. 3. Suggested regulated power supply.

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1. G. D. Young, VE7BFK, "Electronic Keyer," *ham radio,* November, 1969, p. 32.
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3. Paul V. Wanek, "Nomograph Charts—A Fast Way To Build A Notch Filter," *Electronics,* September, 1969.

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nomograph for reactance problems

An easy-to-use computational tool that lets you solve complex inductive and capacitive reactance problems without mathematics

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Nomographs are aids for quickly solving many electronic circuit problems. A straight-edge placed across appropriate scales allows you to solve equations without using a slide rule or pencil and paper.

The nomograph in **fig. 1*** can be used to solve reactance problems when one quantity is unknown and two are known. Chart A is used to determine magnitude and decimal-point location. The significant figures are determined from chart B.

practical example

Suppose you're interested in a circuit such as that shown in **fig. 2**. It's a Q-multiplier that can be added to your strip for increased selectivity. Let's say you have an inductance of fairly high Q whose value is 5 mH. You'd like the circuit to resonate at 1 kHz; what value capacitor should you use in the op amp feedback circuit?

You could determine the capacitor's value by well-known mathematical formulas, but the nomograph of **fig. 1** will provide the answer much quicker. Here's how it's done.

In **fig. 1A**, a line is drawn between the two known values: 5 mH and 1 kHz. This is labeled 1 in **fig. 1A**. The intercept of line 1 on the X_L scale of **fig. 1A** shows the inductive reactance of this combination to be somewhere between 10 and 100 ohms.

Moving to **fig. 1B**, a line is drawn between 5 on the L scale and 1 at the top of the F scale. This location of line 2 was

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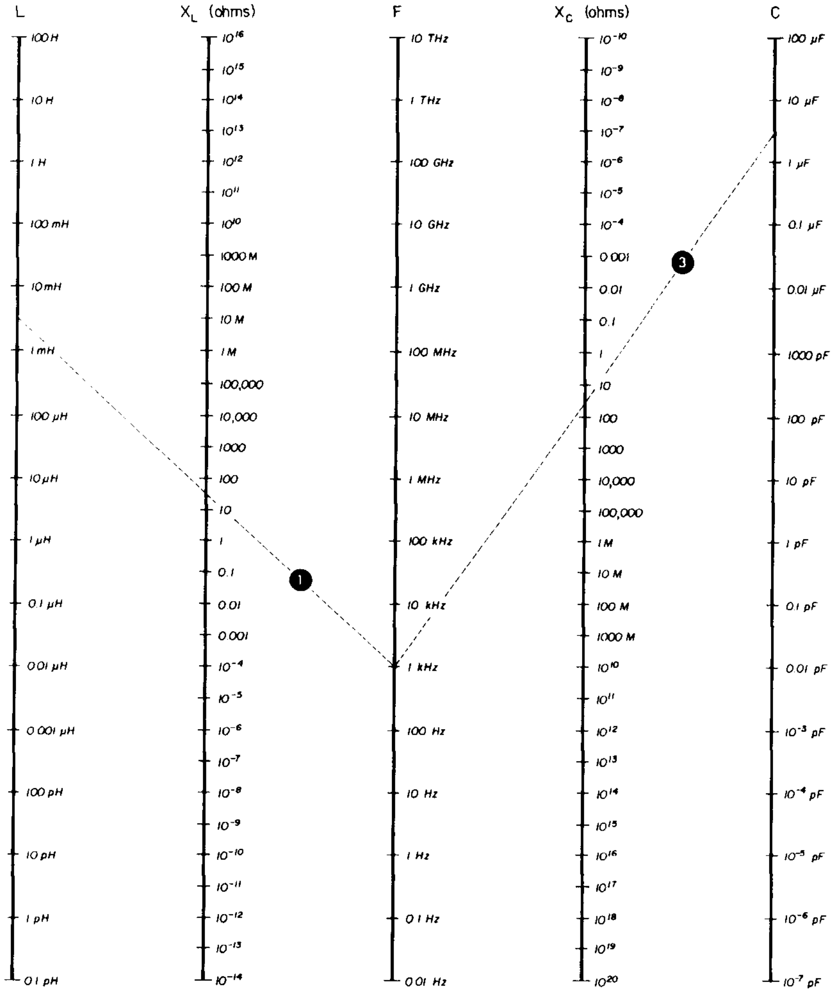


fig. 1A. Reactance nomograph. This chart is used to determine magnitude and decimal location; significant figures are found from fig. 1B.

purely arbitrary; the line, in this case, could just as easily have been drawn between 5 on the L scale and the bottom of 1 on the F scale.

The intercept of line 2 on the X_L scale of **fig. 1B** is what's important. It intercepts the X_L scale at approximately 3.2. This is the significant figure, or decimal multiplier, for the value determined from **fig. 1A**. The inductive reactance is therefore 10 (from **fig. 1A**) multiplied by 3.2, or 32 ohms.

determining capacitance

Returning to **fig. 1A**, a line is shown between 1 kHz on the F scale and 32 ohms on the X_C scale. This is labeled 3 in **fig. 1A**. Line 3 intercepts the C scale at 5 μF, which is the desired capacitance for C1 of **fig. 2**.

An inductive reactance of 32 ohms is used to find C1's capacitance, because basic theory says that resonance in a tuned circuit requires that X_C equal X_L. This is the principle behind the calculations shown here. The nomograph can be used to solve other reactance problems as well.

useful hints

When working with nomographs, the accuracy of the final result will depend

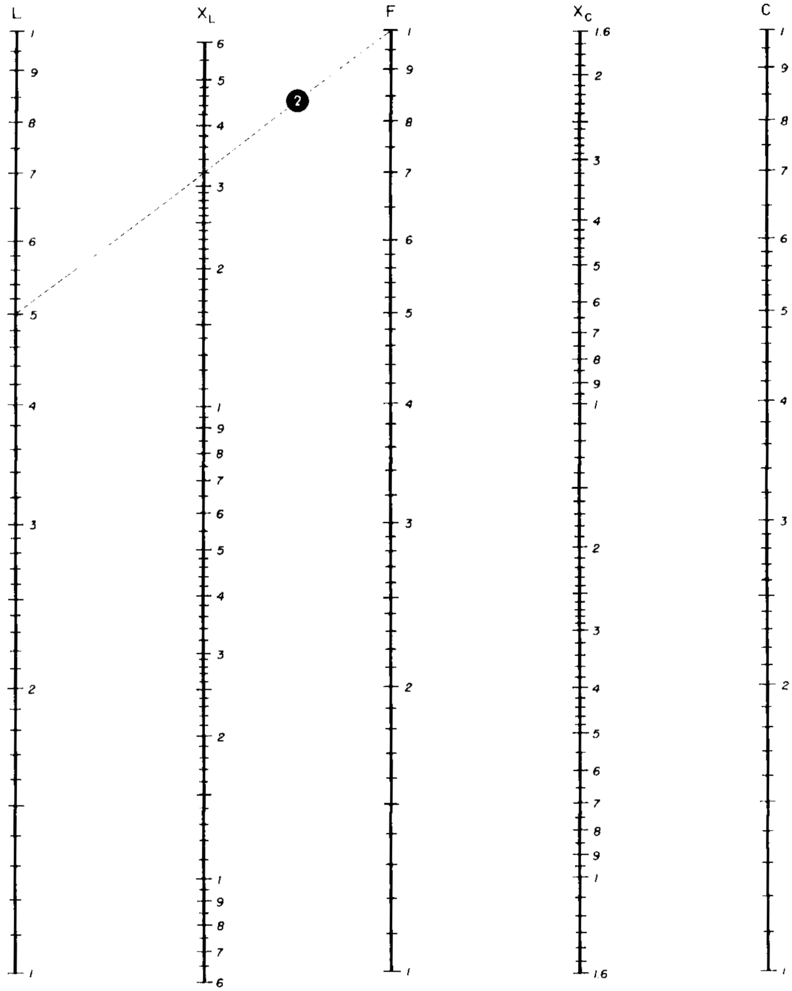


fig. 1B. Reactance nomograph. This chart is used to find significant figures after magnitude and decimal location have been determined from fig. 1A.

on how accurately you draw the connecting lines between the unknowns. Many nomographs give you a "ballpark" answer. If you wish to refine the result, you'll have to use mathematics. If used with care, a nomograph is a great time saver and can provide answers with accuracy sufficient for most practical problems.

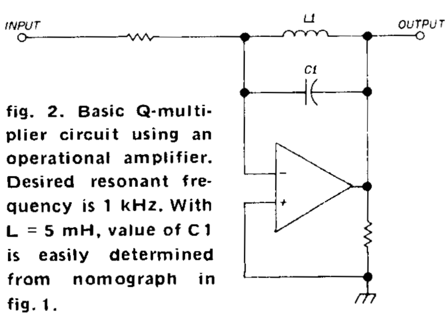


fig. 2. Basic Q-multiplier circuit using an operational amplifier. Desired resonant frequency is 1 kHz. With L = 5 mH, value of C1 is easily determined from nomograph in fig. 1.

A sharp pencil or draftsman's dividers should be used to locate the end points of the two unknown variables. Lay a straight edge against one leg of the dividers, and rotate the straight edge until you pick up the other point. Read the value at the intercept and mark it down on a piece of scratch paper. This will avoid cluttering the nomograph, which can then be used indefinitely as a computational tool.

ham radio

parasitic oscillations in high-power transistor rf amplifiers

Robert C. Wilson, WØKGI, 6577 S. Newland Circle, Littleton, Colorado 80120

More than one design engineer has gained a few grey hairs trying to clean up the output of his 50- or 100-watt transistor transmitter. The reason is parasitic output frequencies, which were not mentioned by the textbooks and which may have been included in the output-power rating by the transistor manufacturer.

The subject of parasitic output has been avoided whenever possible by device salesmen, but in reality it's the one big reason why transistor transmitters aren't found in great profusion. Transistor parasitics are unlike tube oscillations, and the best solution to the problem is still to be found.

transistor parasitic oscillations

There are three types of transistor parasitics. The first is the free-running type. This may be due to tuned or semituned circuits that self-oscillate at frequencies unrelated to the amplifier driving frequencies. The solution is the same as in tube design; that is, the problem circuit is reduced in Q or detuned. The second and third types of parasitics are much more insidious and difficult to eliminate.

The second type is due to the parametrically pumped characteristics of the transistor and is produced as submultiple frequencies of the amplifier driving frequency. These are very difficult to detect, as they are exactly locked to the drive frequency and are normally outside the range usually checked for spurious response.

The third type is low-frequency noise amplification, due again to parametric pumping. The low-frequency noise is not harmonically related to the amplifier drive frequency, but may be of either the so-called 1/f origin or from any low-frequency circuit that can be pumped into oscillation. This type of parasitic may be noticed as a rough-sounding spurious signal in the vicinity of the desired signal, or as white noise centered symmetrically around the desired frequency. If type 3 occurs it is low in frequency, but it will modulate the desired signal in the same way as any modulator, except with highly undesirable results.

The last two types may be detected indirectly by sudden increases in power output, or as changes in collector current while tuning. Because most people tune for maximum output as indicated by a power meter, most likely these spurious frequencies will be maximized.

stabilization methods

Despite the fact that it has often been advocated, detuning the amplifier is no solution to the problem. The real solution is to prevent the occurrence of the parasitic under any condition. To date this has not been possible, but several techniques will make rf amplifiers reasonably stable. One method, recommended widely, is to prevent low-frequency response in the amplifier by using very small values of inductance in the base return and collector dc feed. **Fig. 1** is an example. A problem occurs when the frequency is reduced in that the chokes must be larger, and often the expected result happens; that is, parasitics.

Tube-type neutralization networks are generally ineffective. The reason is that the transistor capacitance varies at an rf rate, while the neutralization components are constant. A method I've found useful is to suppress low-frequency gain by resistive feedback **(fig. 2)**. The cost is a small amount of lost rf gain (perhaps 1 dB) and a somewhat increased dc-circuit current. The thing to remember is not to bias the transistor full-on in the quest for feedback. The result in a noncurrent-limited circuit is the immediate burnout of the transistor—a very costly fuse!

Summarizing, at least three types of rf transistor parasitics occur. These often go unnoticed, because they're not at expected frequencies and are sometimes produced in ways foreign to tube engineers. So far, techniques for parasitic reduction leave much to be desired as they are brute-force methods rather than elegant solutions to the problem. Clean, high-power rf amplifiers are a possible but ticklish proposition and await some yet-undiscovered technique to produce optimum results.

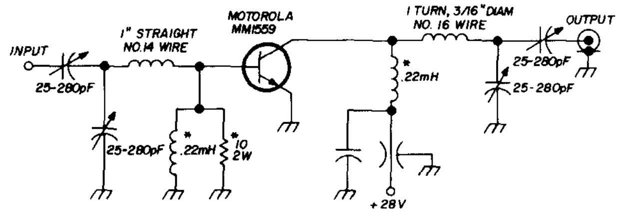
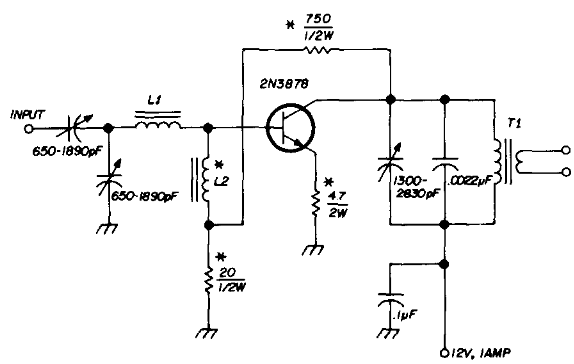


fig. 1. This is the brute-force method of stabilization, where pumped frequencies are swamped by shorting out low-frequency power. Values shown are for 150 MHz. Components marked with an asterisk are used to stabilize the amplifier.



L1 10 turns no. 18 on 1½" toroidal core, Indiana General type Q-1

L2 3 turns no. 18 on 2" toroidal core, Indiana General type Q-1

T1 Primary 7 turns no. 18, secondary 3 turns no. 18 on 1½" toroidal core, Indiana General type Q-1

fig. 2. A 160-meter stabilized amplifier using various methods of reducing types 2 and 3 pumped parasitics. Components marked with an asterisk are used to stabilize the amplifier.

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cw transceiver operation with transmit-receive offset

It's easier to operate cw with a slight frequency offset between direct-conversion transceivers— here's why

Richard S. Taylor, W1DAX, 54 Slade Street, Belmont, Massachusetts 02178

Recently while playing with a design for a direct-conversion cw transceiver, I had some thoughts about the significance of transmitting and receiving on slightly different frequencies, as this design requires. No one seems to have given this idea much thought, although many apparently operate this way all the time. My first reaction was that there might be some situations in which the offset would cause real problems—for example, two identical transceivers trying to talk and getting nowhere because one of them was tuned to the wrong side of zero beat.

My conclusion is just the opposite. The offset adds no new problems to the operation of the transceiver. In fact, if the transceivers are identical (and the offsets the same) it's easier to operate with the offset than without it.

frequency offset

First, the frequency offset occurs in a direct-conversion transceiver (**fig. 1**) because the vfo must be offset from the received signal frequency to produce an audio beat note. Since the transmitted signal is just the vfo output amplified, the transmit and receive signals are different by the frequency of the beat note. **Fig. 2** shows a response curve of a cw direct conversion transceiver (dct) and how two other dcts might communicate with it.

In **fig. 2**, the rf response curve has the form of the audio filter plus its mirror image. The mirror image occurs because signals are audible on either side of zero beat.

Stations anywhere under the curve can be heard. In looking for a contact transceiver one, say, tunes for a signal above zero beat and responds. The home rig will hear him at f_2. Transceiver two tunes for a signal below zero beat, home hears him at f_3. Communication is possible in either case. Note that the beat frequency, although different in the two cases, is the same between pairs of stations. When home and station one talk, they both will have a beat frequency of f_{b1}; when home and two talk, it will be f_{b2}. This requirement for a common beat frequency may be disconcerting to operators whose ears peak at different frequencies, but it doesn't prevent contact from being made.

Operating with another station that's not a transceiver is just as simple and proceeds as above, except that now each station can choose a favorite beat note because of the bfo control at the non-transceiver station.

calling CQ

In answering a CQ with a transceiver the home station does not zero beat, because he can't hear the calling station.

The home station must tune the calling station to one side or the other of

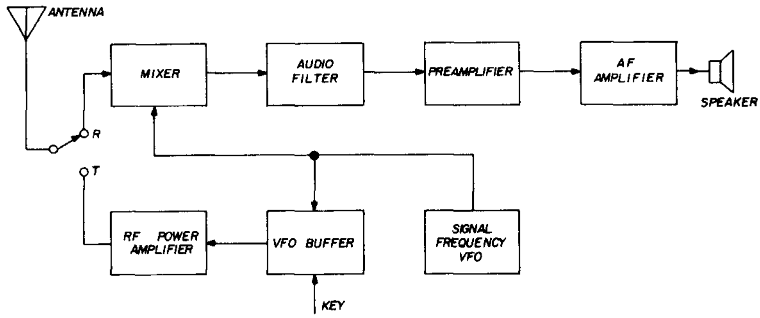

fig. 1. A basic cw direct-conversion transceiver. The vfo is common to receiver and transmitter.

zero beat. The choice will probably be made on the basis of least interference. If the other station is also a transceiver, communication will automatically be established. If the other station is not a transceiver, the offset probably won't be noticed as he tunes for replies.

In calling CQ with a transceiver, however, the replying station, if not a transceiver, will probably zero beat the transceiver carrier. The transceiver would then have to be retuned to hear the reply. On the second go-round the replying station might think the transceiver has drifted (it has, of course), but as long as the replying operator retunes only his receiver and leaves his vfo alone, communication will be established on the second try, and no further retuning would be necessary.

If the transceiver were retuned to reply to a crystal-controlled station, no contact would be made. This is, of course, the same thing that would happen with an ssb transceiver. The solution is to listen around your own frequency and wait for people to come to you, as is done on ssb.

So the offset need not be a problem. At least it should be no more of a problem than the inability to listen on any frequency but one's own, which is the characteristic of any transceiver.

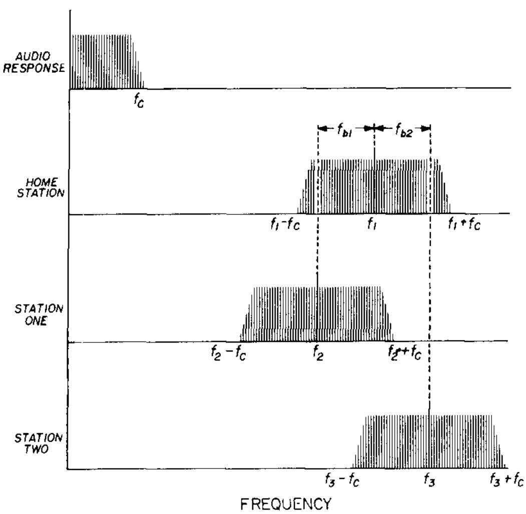

fig. 2. Response curves for three dct's. The audio response for each rig is shown on the first line. The beat frequencies are the difference between carrier frequencies in each case: fb1 = f1 — f2 and fb2 = f3 — f1.

what about audio selectivity?

Surely, the wide-open bandpass shown in **fig. 2** is unsuitable in today's conditions. Audio peaking can improve receive selectivity, but this introduces a problem **(fig. 3)**. Here, both the home station and station one have added audio selectivity, but at different frequencies. The home station is peaked at f_{b2}; station one is peaked at f_{b1}. Let's say station one is calling CQ and the home station hears him, tunes him to his audio peak, and replies. But home station's transmitting frequency is outside station one's audio peak and, therefore, may not be heard. Avoiding this situation requires either opening the bandwidth of both transceivers, or standardizing the offset between them.

I recommend offset standardization by means of identical peaked audio filters at *the two transceivers.* Not only is improved selectivity achieved, but a system advantage is gained as well. Tuning is simpler. Once home tunes the incoming signal to his audio peak, communication is optimized, because the signal transmitted from the home station will be at the audio peak of the other transceiver. One envisions highly selective transceivers with but one frequency control, which is attractive.

bandwidth considerations

The best shape for the peaked bandpass requires careful consideration. A sharp peak is desirable from the standpoint of both offset standardization and selectivity. It would be undesirable to have too accented a peak, however, as it would be hard to hear stations with offsets different from your own. Possibly 6–8 dB would be a good number for the peak above the low-frequency response level (**fig. 4**). An optimum width might be a few-hundred Hz at the 6-dB points. An

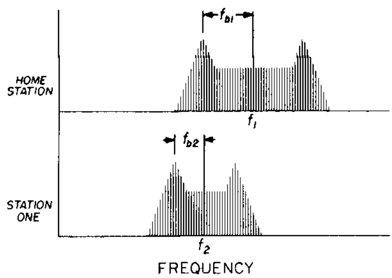

fig. 3. Two dct's with audio selectivity at different frequencies. Note that when the home station centers his audio peak on station one's carrier at f2, station one's audio peak is way off the carrier at f1. In this example station one wouldn't hear the home station at all.

ability to hear stations below the peak (nearer zero beat) is desirable, so that stations which tried to zero beat your carrier would be audible. This would make the direct-conversion transceiver a good performer when working all kinds of rigs—not just when working other transceivers.

It's entirely possible to build a direct-conversion transceiver using phasing techniques that would provide true single-signal reception with a response on only one side of zero beat. It would look much like the direct-conversion ssb receiver I described in reference 1. I don't believe there is any advantage to doing this, however, because the sideband that the single-signal receiver ignores may well be the one on which the responding station replies. If such transceivers became common, it would be necessary to designate which sideband was to be used on each cw band to ensure that the transceivers

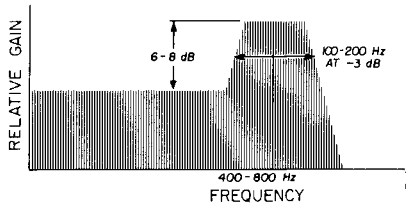

fig. 4. A possible standard passband for cw dct's. The optimum numbers would be determined by in-use testing.

would communicate. The dsb version; i.e., with response on both sides of zero beat, seems the best choice, at least at present.

conclusions

1. Transmit-receive frequency offset is not a hindrance to cw communication, whether between two transceivers or a transceiver and regular stations.

2. Frequency offset standardization is required if good cw selectivity is to be obtained.

3. This standardization is best achieved by means of a standard peaked filter response to be used in all cw transceivers.

Maybe the cw dct would be the thing to get a lot of us back on the cw bands. Certainly the simplicity is attractive.

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1. Richard S. Taylor, W1DAX, "A Direct-Conversion S.S.B. Receiver," *QST*, September, 1969, pp. 11-14.

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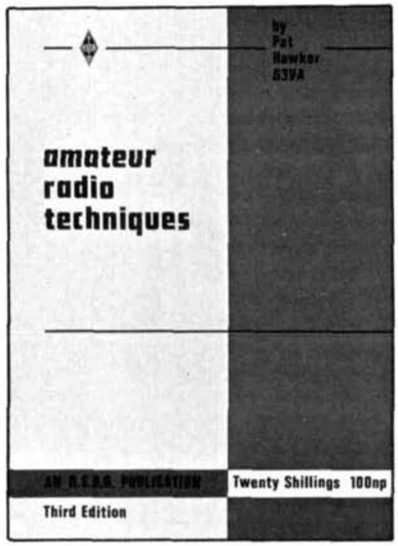

AMATEUR RADIO TECHNIQUES

J. Pat Hawker, G3VA

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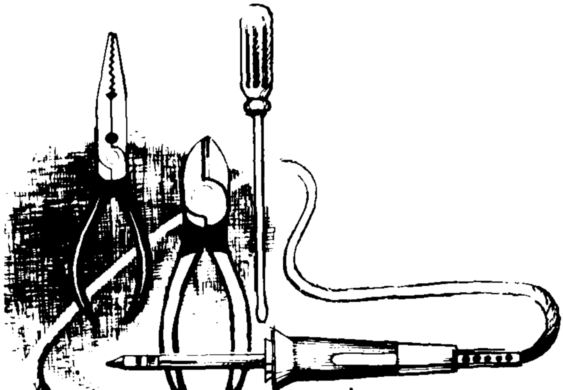

the repair bench

by larry allen

finding faults in rf and i-f amplifiers

What do you do if you can only pick up nearby or strong stations? What if they come in weak and perhaps noisy? First thing you should do is suspect the rf amplifiers in your receiver.

This isn't unusual, particularly with transistor front ends. If the manufacturer (or you, if the receiver is homebrew) designed the rf amp to use the sensitive but delicate field-effect transistor, a gate punctured by static surge isn't at all uncommon.

The problem lies in recognizing a bad front end. If the mixer and i-f stages are naturally quiet, and transistor stages often are, you may not know an i-f failure from an rf one. Or the fault may be in the automatic gain control (agc) system. Only careful testing will tell you for sure.

amplification vs noise

Familiarity with your receiver is the best assurance of knowing when there really is trouble. You should get to know how much natural receiver noise to expect when there's no station.

Then, when suddenly you can't pick up stations you know should be there, make a listening test. Tune the receiver dial to an empty spot. Turn rf and af gain up.

Is receiver noise (the background thermal hiss) up to snuff? If so, the i-f stage must be amplifying. Also, the mixer stage is probably okay; much front-end thermal noise normally originates there.

Yet, some of today's field-effect transistor (fet) front ends are too quiet for this kind of analysis. It's normal to hear almost no receiver hiss. You have no choice then but to rely on other testing methods. You try to determine rf-stage sensitivity. Or, you can just test the stage by regular dc-measurement methods.

what's in a rf stage

Most of what I say about troubleshooting rf stages can be applied to i-f stages, too. There's little difference.

In most a-m receivers, rf stages tune over several different bands. The i-f stages are fixed-tuned. But it's common to tune ssb receivers nowadays by synthesis—the same as the transmitter. In that case, the receiver rf stages are fixed-tuned. You troubleshoot them the same as i-f stages.

A typical old-fashioned tube-type rf stage is drawn in **fig. 1**. Only one deck of the bandswitch is shown; the input band coils are omitted for simplicity.

Most hams can figure out a way to track down trouble if they are sure what a stage is supposed to do. I don't mean in just a general way, but specifically—each part of the stage. Take the stage in **fig. 1** for an example.

First and foremost are the *input* and *output* circuits. T1 and C2 are the input coupling components. R1 is the input load, decoupled by C3. Ordinary signal tracing or injection is the way to make sure these components work.

The output circuit is T2, decoupled by C7. Those decoupling components are an important part of the input and output circuits, so don't forget them when you analyze stage operation.

There are also tuned circuits to contend with, although part of them are not lyze the screen or cathode dc circuit.

The cathode dc circuit, which can be considered dc supply even though it's just a ground return, is through R2 and R3. C4 or C5 can become part of this circuit if either happens to short. Otherwise, they don't affect cathode dc voltage.

Notice that R3 is variable. It's the *rf gain* control for the receiver. (In the receiver from which this example is taken, R3 also is part of the i-f stages.

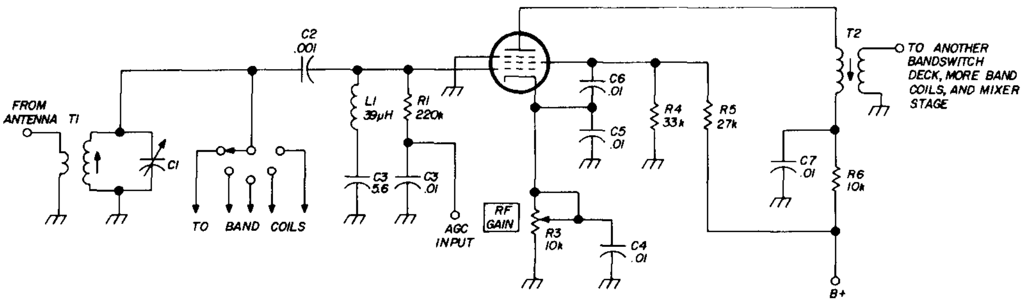

fig. 1. Rf stage contains circuits that must be checked individually if you don't use test procedures that check overall stage performance.

shown in **fig. 1**. T1 and C1 make a tuned circuit. But they are in parallel with band coils you can't see. The band coils, with the inductance of T1, set the band the stage is to tune across and C1 tunes the specific frequency. If there's a fault in one of these coils, that band won't tune properly. If the fault is in T1 or C1, *none* of the bands tune as they should.

Next, consider the *supply* circuits. They carry dc voltage to the tube elements.

The plate supply is through R6 and the primary of T2. Capacitor C7 is important in the plate dc supply circuit only because of the possibility it might short. You therefore must consider it part of the plate supply circuit when you're diagnosing.

Resistors R4 and R5 are the chief components of the screen supply circuit. C6 is a potential part of it—if the capacitor happens to become leaky or shorted. Leakage in C6 would put voltage intended for the screen onto the cathode. Consider that possibility when you analyze the screen or cathode dc circuit.

That connection is omitted here for simplicity.) Changing the value of R3 between R2 and ground varies cathode bias on the tube. The pentode is a sharp-cutoff type; its gain depends sharply on its bias. Thus, by changing bias, R3 controls rf amplification.

Voltage at the grid is controlled from the agc stage. The agc control voltage is fed through R1. Decoupling capacitor C3 is part of the grid-supply circuit only if it shorts or gets leaky. C2 and C3 might become part of that circuit if either went bad.

There's another resonant circuit. It isn't tunable. It may also be called a trap circuit, because that's what it's there for. L1 and C3 form it, and it's resonant to 9 MHz, the i-f of this receiver. It traps out any stray 9-MHz signals, preventing them from being amplified and running through the mixer to upset the i-f stages.

That about sums up the circuits in this rf stage. Input, output, tuning, bypass or decoupling, and trap: those are the signal-carrying circuits. Plate supply, screen supply, cathode return, and grid bias: those are the dc-carrying circuits. You have to consider each when you set about troubleshooting an rf stage like this one.

transistor rf stages

The transistor input stage from one receiver is drawn in **fig. 2**. You can probably identify the circuits. They may look a bit different from those in **fig. 1** because they're in a transistor stage.

The input circuit comprises L1, C1, C2, and both R1 and R2 as load resistors. Decoupling for the two resistors isn't in the diagram, but it's understood. The agc line has a bypass capacitor not shown; so

The transistor is pnp. Therefore, normal forward-bias operation puts the emitter positive, the base less positive (same as more negative than emitter), and the collector far less positive (same as far negative from the emitter).

The emitter gets voltage directly from a positive 12-volt supply line through R4. C3 is the decoupling capacitor, and is a concern to the dc circuit only if it shorts or gets leaky.

Collector goes to ground through L2. The winding has no appreciable dc resistance, so for dc the collector is grounded. That puts it far negative with respect to emitter.

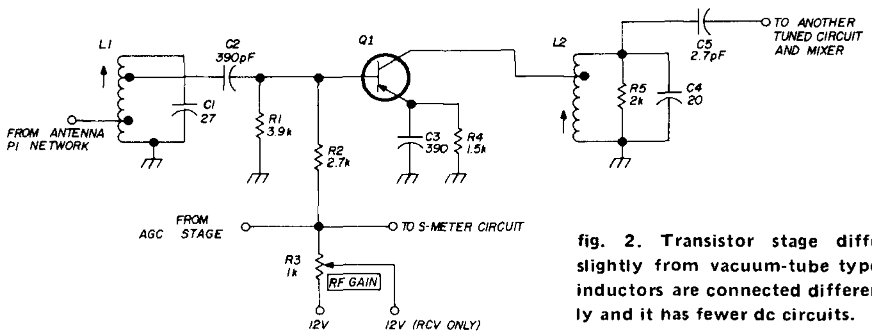

fig. 2. Transistor stage differs slightly from vacuum-tube type—inductors are connected differently and it has fewer dc circuits.

does the 12-volt supply.

L1 and C1 are fixed-tuned, although adjustable with a tuning tool. They form a broadband tuned circuit. (This particular rf stage is part of a one-band transceiver.) The taps on L1 are for impedance matching.

The output circuit comprises L2, C4, R5, and C5. L2-C4 are a tuned tank, with R5 as a band-broadening load across it. L2 is adjustable for band-peaking. C5 couples amplified rf energy to the mixer stage (through another tuned circuit, omitted for simplicity).

The only other signal circuit is emitter bypass capacitor C3. If it opens, substantial degeneration can occur, but it only makes the stage weak—it doesn't make it dead.

There are only three dc supply circuits. That's because a transistor has only three elements to receive voltage.

The base has the only complicated supply network. The main dc comes from the 12-volt line through R1. However, a connection through R2 lets the actual voltage—and therefore bias on the transistor—be varied by the agc line and by the setting of *rf gain* control R3. The transistor operating characteristic is such that bias controls amplification. Thus the *rf gain* control sets optimum gain of the stage, and agc varies it to accommodate signal strength.

Of course, C2 is part of the base dc circuit only if it comes defective. A faulty decoupling capacitor in the agc line could also affect base bias. You need remember these capacitors only if you're troubleshooting and find the base voltage is wrong.

external influences

One other thing you can't forget when

you're troubleshooting rf and i-f stages. A trouble in the stage may be caused somewhere else.

A blocked i-f or rf stage is common. Sometimes, that's traced to an agc stage overdoing its bias thing. Or, the i-f or rf amp may block on strong signals—a sign of overload. That, too, may be traceable to a faulty agc stage—this time not doing enough.

The receiver block diagram is sometimes helpful in spotting stages that affect rf or i-f. Some don't show that much detail and you have to rely on your ability to read the schematic. **Fig. 3** is a partial block diagram of the set from which **fig. 2** is taken. Its detail is enough to be helpful.

You might find the voltage upset in an i-f stage, yet the agc stage works normally. Suspect the s-meter hookup. If any part of that circuit shorts to ground, it could foul up bias on rf or i-f stages. A short inside the s-meter might make the *rf gain* control work wrong.

In other words, examine the schematic or block diagram before you go tearing into any rf or i-f stage. If external stages affect the rf stage, check them out or isolate them from the rf stage some way.

testing rf amplification

One way to see if an i-f or rf stage is working right is to measure its gain. In modern transistor receivers you can expect to find a voltage-gain factor of 20 or more. A tube stage usually gives even higher gain.

You can make this measurement fairly easily if the output of your rf generator is calibrated. First, clamp the agc line with whatever dc voltage produces normal idling (no-signal) bias on the rf or i-f stage you're testing. That bias is usually written on the schematic or on the voltage chart.

Clip your vtvm to the a-m detector output, or through an rf probe to the output of the last i-f amp. Feed the generator signal to the *input* of the rf or i-f stage being measured. Note the dc meter reading. Set the generator output level for some meter reading that's easy to remember. Make a note of the rf output level of the generator.

Move the generator signal to the output of the stage being tested. Turn up the generator level until the meter reads the same as before. Divide the new generator output-level reading by the earlier one. The result is the voltage gain of the stage.

As an example, suppose 0.7 microvolt of signal drives the meter to 1 volt dc when the generator is connected to the stage input. When you connect the generator to the stage output, you have to turn the generator up to 14 microvolts to get

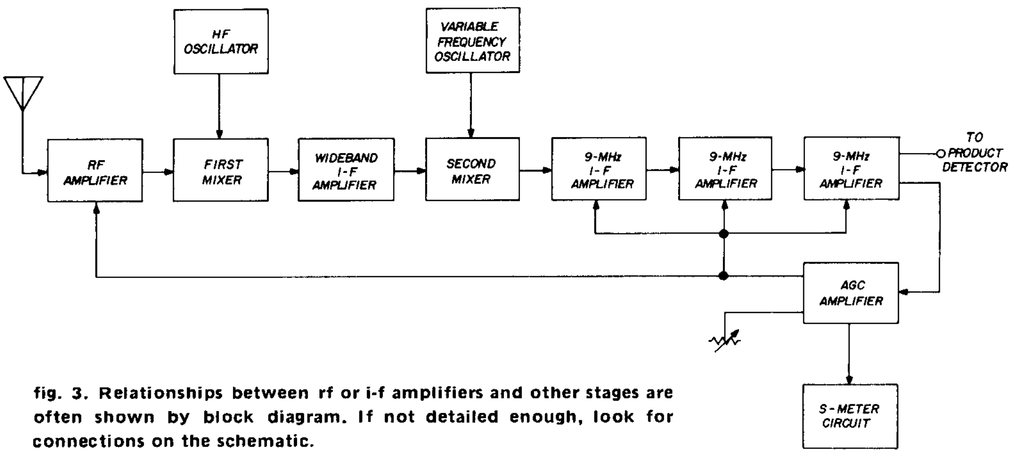

fig. 3. Relationships between rf or i-f amplifiers and other stages are often shown by block diagram. If not detailed enough, look for connections on the schematic.

that 1-volt dc reading on the meter. Dividing 14 by 0.7 gives 20. That's the voltage gain of the stage.

Unfortunately, only the more costly signal generators have calibrated output. You may have to use a less accurate way. It's only relative, and your best bet for using it is to make measurements while your receiver is working normally and record them for reference.

You'll need a doubler-type rf probe for your vtvm. A suitable circuit is sketched in **fig. 4**. It's more sensitive than the ordinary single-diode probe. The vtvm should be a very sensitive one, with lowest full-scale reading 1.5 volts or less.

Again, clamp the agc. Turn the *rf gain* control wide open. Keep the generator unmodulated.

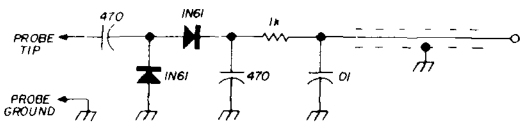

fig. 4. Circuit for a voltage-doubling probe you can use with a vtvm to measure relative rf signal levels in rf and i-f stages. Entire probe should be shielded to prevent hand capacitance from upsetting the reading.

Feed the rf signal to the antenna input jack. Tune the generator to the center of the band if the stage is fixed-tuned; if the stage is tunable to one frequency, set the generator precisely to that frequency.

Connect the vtvm probe first to the base of the rf transistor. Set the meter on its lowest range. Turn up the generator signal until you get a perceptible reading on the meter. Then set the generator output for some very small but definite voltage indication—say 0.01 volt. Don't change the generator setting.

Move the vtvm probe to the base of the mixer. The reading should be much higher now—say nearly 0.2 volt dc. That represents a gain of about 20 if you're using the doubler probe.

Obviously, these figures are approximate. Voltage gain for a tube is roughly the same. For tubes or transistors, however, the surest system is to make a record of normal amplification while your receiver is new. Then use the same measurement method when you test on the repair bench.

dc troubleshooting

You should already know the methods of tracking down the cause of incorrect dc voltages on an rf stage. You might want to make dc tests before you go to the trouble of putting the rf probe on your vtvm.

But usually you'll find that kind of fault is obvious—as when the stage is completely dead. It's for subtle weakness or abnormal overloading you need to clamp the agc and test stage gain. Then you can hunt down the small voltage problem—or bad transistor or tube—that's causing the trouble.

checking the stage another way

You can also use a form of signal injection. Connect the vtvm, without a probe, to the a-m detector of the receiver. If it's handier, keep the probe on the vtvm and connect it to the i-f input of the product detector (or at the output of the last i-f amp). Clamp the agc as before.

Connect the rf signal generator, tuned to the rf frequency as already described, to the input of the mixer stage. Turn up the generator output just enough to cause a reading on the meter. Make a note of the reading.

Move the generator back to the input of the rf stage. Note the increase in the reading. The dc-voltage increase should be about the same as the one already described. Without the doubler probe, a 10-times increase means about 20 gain in the rf stage. With a doubler probe instead of the set's a-m detector, a 20-times increase means roughly 20 gain in the rf stage.

Now go a step further. Be sure you've got the generator frequency set precisely. Try tuning the coils in the tuned circuits. If the meter reading doesn't vary, the coils may be at fault. Before you replace them, though, make sure the capacitor that decouples each coil is not open.

Finally, if tuning is erratic and you can't seem to make head nor tails of how the coils tune, check the bypass capacitors on the supply lines and on the agc line. One of them may be open.

looking to the future

Remember that the ways of checking rf stages outlined here can be used just as well with i-f stages.

There's still another way of troubleshooting rf and i-f stages: with sweep alignment. Unfortunately, no inexpensive sweep generator available today goes down far enough in frequency.

In a future column I'm going to show you how to make your regular sweep generator go down far enough to sweep 60-kHz, 455-kHz, and other i-f amps. You can do it without modifying the instrument you have.

First, though, there's a new troubleshooting system that has come to my attention. It's called *1-2-3-4 Servicing* by its originator, Forest H. Belt. In the next issue of repair bench, I'll tell you what *1-2-3-4 Servicing* is all about. That'll prepare you to understand what goes on when you use the sweep-alignment method of rf and i-f troubleshooting.

ham radio

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the ham notebook

protection for solid-state power supplies

The current crop of Heathkit solid-state power supplies is great. I have three of them, which I use in ham work and in areas far removed from hamming. Some problems developed with these supplies, and the purpose of this note is to recommend a slight design change that you can make to protect your power supply. Even if your supply isn't a Heathkit—home brew, for example, but using similar circuits—you might consider this inexpensive way to add protection against external influences.

Many experimenters have surplus relays, especially the 28-volt type with coils that require about 200 mA for operation. It is convenient to test them with a solid-state power supply. The stored energy in the relay coil can develop quite a wallop when the circuit is opened to de-energize the coil. Standard practice is to connect a diode across the coil to suppress the stored voltage. However, when checking many assorted relays, the tendency is to go the easy way and take a chance that the switch gap, upon opening, will dissipate the stored energy. You might get away with this a few times, but not for long. Sooner or later a large negative voltage spike will back up into the power supply, and you've got problems.

The easiest way to protect the supply in this case is to add a diode across its output terminals. Solder the diode to the lugs on the back of the binding posts, cathode to positive; anode to negative. A 1N4003, 1N538, 1N645 or similar type will work. Any negative transients trying to sneak back into the supply will be shorted to ground at the terminals. So much for protection from one external influence.

Another external influence is a short circuit, often a dead short, across the supply. Heath has a dandy circuit that cuts off the current when a short circuit is sensed. Upon removal of the short, current is restored, and no harm is done. The Heath circuit contains two supplies, each mutually independent. One is the heavy-duty supply, which provides power to the load. The other is a zener-regulated, constant-voltage reference source. Across this source is a potentiometer that provides a variable voltage. The positive terminal of the reference source is tied to the positive terminal of the power supply. The pot output is applied to the base of an error detector/amplifier transistor. During normal operation, the unregulated supply will attempt to match the level of the reference-supply voltage. The error detector senses the difference between these voltages. The result is a slightly positive voltage at the error-detector transistor base. As the pot is varied, the output voltage will tend to follow the reference voltage.

Suppose a short circuit develops. The current-limiting circuit will reduce output current to near zero, reducing the output voltage to near zero. The voltage difference at the error-detector transistor base

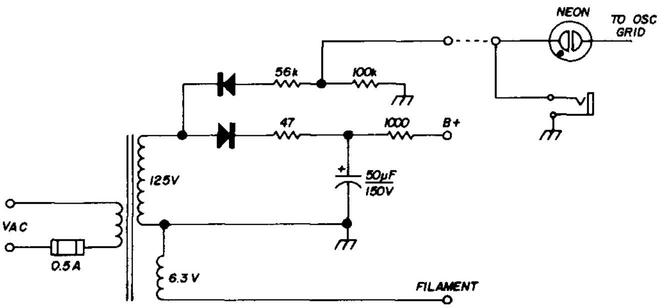


fig. 2. Isolation circuit for use with Heath HG-10B vfo and HD-10 keyer.

will be some voltage between the output voltage (zero) and the reference voltage—anything up to 35 V, for example. Since the reference voltage is negative with respect to common, the full negative reference voltage will appear at the error-detector transistor base, instantly zapping the transistor.

To prevent this disaster, Heath uses a diode in series with the error-detector transistor base. The diode will pass current in the positive direction only. However, if regulation is lost (this happened to me), the output voltage will shoot up to the full unregulated amount, say 50 V. As before, the error detector will sense the difference between output and reference voltages; except this time about +15 V will appear at the error-detector transistor base. Here the diode offers no protection; its job is to protect the supply in the event of a short circuit. The relatively high positive voltage on the error-detector transistor base will pass a high current through the base-emitter junction, and ZAP!

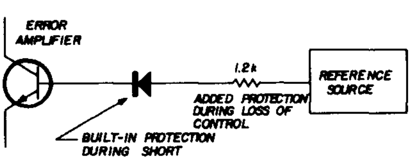


fig. 1. System used by Heath to protect their power supply. Resistor was added by author.

The answer to protection during loss of regulation is extremely simple. A resistor is placed in series with the error-detector transistor base, as in **fig. 1**, to limit base current. The cost in performance will be a slowdown in supply response—about 15 ns.

reference

1. Heathkit Manuals for Model IP-18 and IP-28.

Frank Case, W3NK

independent keying of Heath HG-10B vfo

For owners of the Heathkit HG-10B vfo who might wish to make it independent of other equipment, the following idea will be helpful. My vfo is keyed with a Heathkit HD-10 keyer. The problem was how to use this model keyer with the vfo. The instruction manual for the keyer warns against using a voltage above −105 V.

The circuit of **fig. 2** shows how I solved the problem. All components except the transformer and fuse holder were mounted on a small piece of Vector board. The existing 4-conductor cable was removed and replaced with a length of ac cable. The fuse holder was mounted between the key jack and ac cable. The transformer, an inexpensive Japanese unit, was mounted vertically on the underside of the chassis next to the mode-switch wafer. The Vector board will fit nicely between the transformer and the rear chassis wall next to the terminal strip holding the neon lamp.

The connection from the key jack was removed from the cathode circuit and connected as shown. No switch was used, as I intended to leave the vfo on constantly.

James H. Crouch, K4BRR

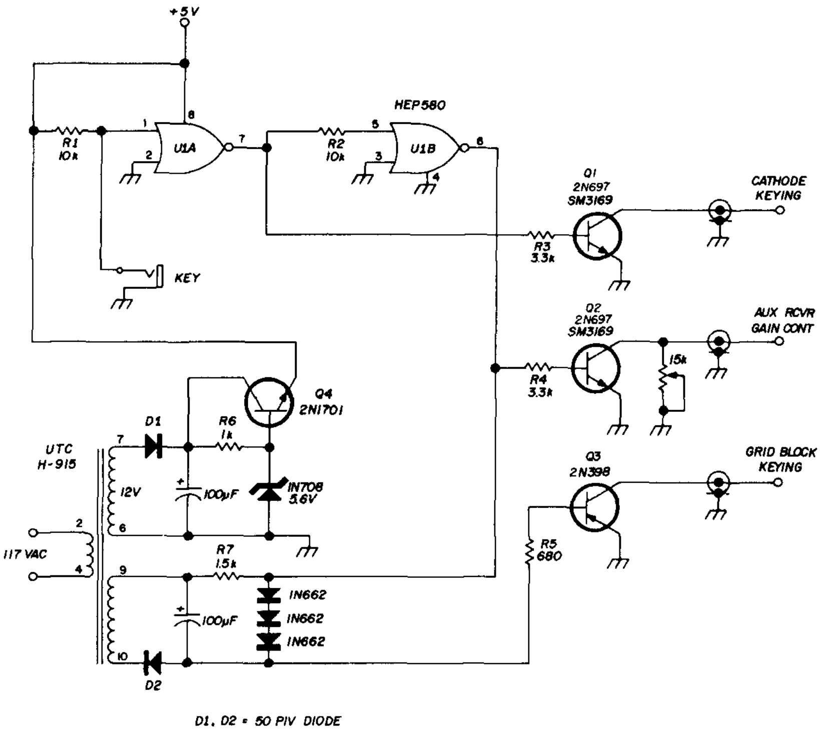


fig. 3. CW break-in control circuit using ICs.

break-in control system

For really fast and effective cw break-in, many amateurs use separate transmitting and receiving antennas. With such an antenna system and fairly low power (under 100 watts output) the control circuit described here is all that's required for full and complete break-in operation. It uses the old idea of inserting additional resistance in series with the receiver gain control when the key is down to automatically decrease receiver gain and prevent overload. This method also allows the transmitted signal to be monitored in the receiver.

Fig. 3 shows the circuit. When the key is up, U1A turns on, and the output at pin 7 goes low (0 volts). The output of U1B goes high because both its inputs are low. This turns Q1 off and Q2 on, so that the transmitter keyed circuit is open, and the auxiliary gain control is shorted to ground. The receiver now has full gain. Closing the key causes U1A to turn off, the output at pin 7 goes high, which forces the output at U1B pin 6 to a low state. As a result, the transmitter is keyed and the auxiliary gain control decreases receiver gain to its preset level.

Since I have two types of transmitters (cathode and block-grid keying), I added a negative source of voltage and Q3 to key the grid-blocked rig. With the key up, the positive voltage on U1B pin 6 overrides the negative supply at Q3's base, holding it off until the key is closed.

This new system replaces a dpdt relay and its keying circuit. In addition it provides for independent cathode and grid-blocking keying of different transmitters. I'm breaking only about 20 Vdc with Q1 and Q2, so the devices shown work fine. The 2N398 would be required for most grid-blocked keying systems.

The whole thing could be powered from a few dry cells, but the surplus transformer was available to make the system operate on 117 Vac.

Cal Sondergoth, W9ZTK

ssb input source for vhf, uhf transverters

Many amateurs require a good vfo for use in the vhf/uhf bands. The 28-MHz output of an ssb transceiver or transmitter can be used as an input source for an up-converter to obtain the desired vhf/uhf frequency. It's highly desirable to maintain the original modulation system of the transceiver or transmitter.

For output on 144 MHz, one could use 166 + 28 MHz; for 432-MHz output, the conversion can be made by mixing 404 and 28 MHz. The best method to obtain output on 1296 is to use 28 + 518 MHz, then mix the 546-MHz resultant signal with 750 MHz. This method will avoid all frequency components except the desired one.

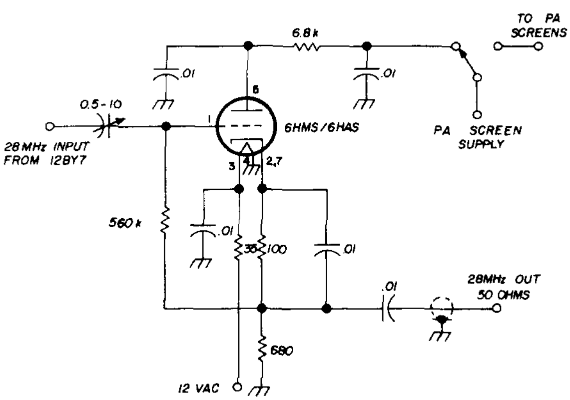


fig. 4. Cathode follower for efficient power transfer from transceiver to transverter.

It's possible to obtain the 28-MHz source form the transmitter or transceiver by (a) running the output into a dummy load and feeding a certain amount of energy through a small capacitor to the transverter, or (b) feeding the output to the transverter through an attenuator. These methods are inefficient and wasteful of power. Therefore, I've taken a new approach to the problem.

I use a type 6HM5/6HA5 tube in a cathode follower (**fig. 1**). This tube is very small, like the 6AK5. The circuit is installed on a little subchassis, which fits nicely in the 6JB6 input compartment of my TR4. The output bnc connector is at the rear, close to the 6JB6 output compartment, and is mounted with the switch on a small panel. When the switch is on, the 6JB6's are disabled, and no power is wasted.

The output impedance is given by $Z_{out} = 1/S = 1/0.02 = 50$ ohms, where S is the tube transconductance. With 135 V on the plate, this tube has a transconductance of 20k μmhos. This results in a normalized value of 50 ohms output impedance.

Because the input impedance of a cathode follower is high, only a very small amount of capacitance is needed to feed the 28-MHz signal from the driver stage, and no misalignment will occur.

Using this method I can obtain 1–10 volts of cw or ssb signal; it also works well on a-m. The stability of the vfo is the same as in the HF bands. The overall stability is determined by the crystal oscillator. I've been using this addition to my TR4 for two years on 144 and 432 MHz with very good reports. The idea can be used in other ssb equipment with equal results.

Jacques Mainardi, F8MK

home-made heat sinks

Anyone who works with solid-state devices knows that when a large amount of power is applied to a transistor, a heat sink is required.

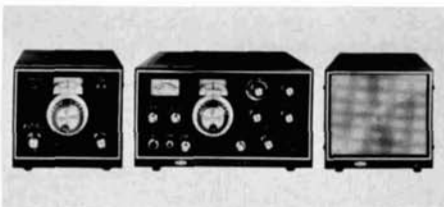
Many tv, fm, and other commercial broadcast stations use tubes such as the 4CX250 and similar types. In most cases these tubes aren't worth rebuilding when they go bad and can be had for the asking.

I found that by cutting the top section from the tube, plus a small amount of filing (and in some cases a little ingenuity), an excellent heat sink can be made for many power transistors. The convenience of having the heat sink around the transistor, rather than spread out along the chassis, can be realized.

Greg Larsen, WAØWOZ

new products

ssb,cw transceiver



Allied Radio has introduced a new five band ssb/cw transceiver that covers the range from 3.5 to 29.7 MHz in seven bandswitched ranges. The new transceiver features a solid-state vfo circuit with a linear tuning system that permits accurate readings to 1 kHz on all bands. The transceiver has built-in sidetone, vox,ptt, fast or slow agc, 25-kHz crystal calibrator, receiver incremental tuning and sharp-cutoff crystal filters, including a 500-Hz filter for cw work.

Power input is 160 watts from 3.5 to 21 MHz, and 120 watts on ten meters. Carrier and sideband suppression are rated at -40 dB. Receiver sensitivity is 0.5 μV for 10-dB signal-to-noise ratio (3.5 to 21 MHz, 1.5 μV on 28 MHz). Selectivity on ssb is 2.4 kHz at -6 dB, 4.8 kHz at -60 dB. On cw the selectivity is 500 Hz at -6 dB and 1.5 kHz at -60 dB.

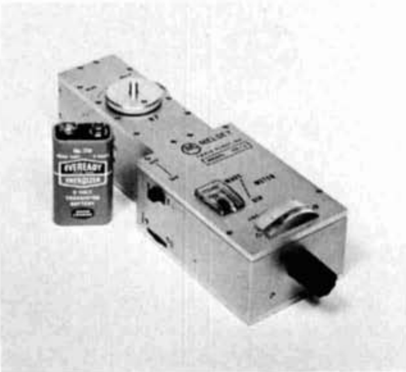
The Allied A-2517 transceiver is priced at $400. The A-2518 matching speaker/ac power supply is $99.95. A matching external solid-state vfo, the model A-2519, provides increased versatility by allowing transmit and receive operation on different frequencies; price is $89.95. Allied products are available exclusively through Allied Radio Shack stores, or by mail. For further information, write to Allied Radio Shack, 100 N. Western Avenue, Chicago, Illinois 60680.

transistor manual

The new edition of the RCA "Transistor, Thyristor and Diode Manual" includes the latest available information on basic technology, operating principles, characteristics and ratings, applications and test of RCA semiconductors. This new manual is 20 percent larger than its predecessor and continues as an authoritative reference on bipolar transistors. In addition, it provides information on mos field-effect transistors, thyristors (scrs, triacs and diacs), silicon rectifiers and other types of solid-state devices. Definitive data are given for more than 900 different semiconductor devices; comprehensive data and design curves for transistors and thyristors are provided. In addition, tabular data are given for silicon rectifiers, other semiconductor diodes and discontinued transistor types.

In the circuits section of this manual schematic diagrams, detailed parts lists and descriptive writeups are provided for 38 practical circuits. Most interesting to the amateur radio experimenter are a mosfet preamplifier for 6, 10 and 15 meters, a two-meter converter, a stable vfo (3.5–4.0, 5.0–5.5 or 8.0–9.0 MHz output), microphone preamplifier, 40-watt 50-MHz transmitter with load mismatch protection, transistor dip meter and an electronic keyer. Other applications include fm tuners, an fm stereo multiplex demodulator, hi-fi amplifiers, power supplies and voltage regulators, battery chargers, an electronic heat control unit, light flashers and dimmers, and several digital circuits. 656 pages. $2.50 from your local RCA distributor; ask for Technical Series SC-14.

uhf wave dip meter



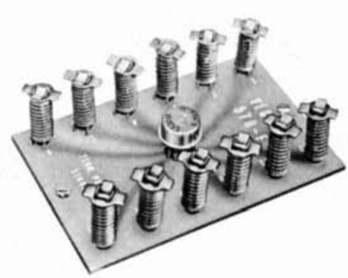
A new solid-state uhf wave/dip meter and marker oscillator has been introduced by the Melsey Corporation. This new instrument provides continuous tuning from 400 to 1150 MHz. Frequency readout—by the use of a 30-inch steel tape—is better than 1%. Design features include a battery-operated transistor oscillator in a cavity configuration that is tuned by a precision glass-invar capacitor. The instrument has an outside coupling loop for general use, and a miniature coaxial receptacle for direct connection with 50-ohm coax cable. The connector assembly can be rotated to obtain variable attenuation (30dB minimum)

In addition, special applications are possible with modifications developed by the manufacturer. With a minor adjustment the output of the unit can be increases to function as a local oscillator for uhf mixers. It can also be tracked to an rf circuit, or used as an fm, a-m or pulse modulated signal generator or target transmitter. The SN-2 wave/dip meter is $185 from Melsey Corporation, 202 Carle Road, Carle Place, L. I. New York 11514

fet applications handbook

If you're looking for practical design data on field-effect transistors, this new expanded 2nd edition by Jerome Eimbinder, managing editor, *EEE Magazine,* contains nearly 25% more material than the previous volume. The in-depth information furnished by editor Eimbinder will be of immediate value to anyone looking for new ideas and unique fet circuit applications, including many basic fet circuit descriptions. Contents include introduction to the fet and basic fet characteristics, biasing fet stages, fets as oscillators, low-noise audio preamplifiers, source followers, phase splitters and switches. Also included are fet measurements, the photo fet, mosfet biasing techniques and fets as voltage-controlled resistors. The appendix includes often-needed data and charts arranged to serve as a convenient quick-reference source. 352 pages. $14.95 from Tab Books, Blue Ridge Summit, Pennsylvania 17214.

ic breadoard socket



Vector Electronic Company has announced a new breadboarding socket that is designed for 12-lead TO-5 integrated circuits. The device consists of an epoxy-glass wafer with a 12-pin socket, the tabs of which have been soldered to two adjacent rows of solderless *Springclip* terminals. The board is furnished with two pins on the bottom that may be press fitted into pre-punched terminal board with 3/32 inch holes such as AA-pattern Vectorboard.

This new breadboard socket simplified integrated-circuit experiments because as many as four solderless connections can be made quickly to any terminal pin, and the user may use as many ICs as he needs by simply using additional breadboard sockets. The price of the 570F IC socket is $3.95 and may be ordered from the manufacturer, Vector Electronic Company, Inc., 12460 Gladstone Avenue, Sylmar, California 91342.

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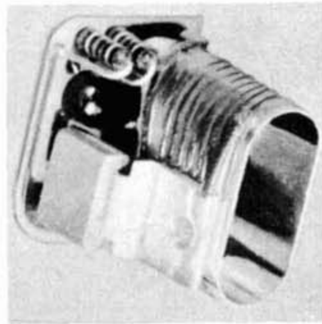
ambient compensator



Isotemp Research, Inc. has recently announced an ambient temperature compensator for quartz crystals that uses a proportional solid-state control circuit to maintain crystal temperature within ±0.05° C at constant ambient temperature and constant voltage supply. The set temperature of the unit is 75 ±2.5° C. Warmup time is approximately 6 minutes from -30° C. Maximum power demand is 4 watts; approximately 1¾ watts are required to maintain crystal temperature at -30° C ambient. Required supply voltage is 12.0 Vdc (6 Vdc to 24 Vdc are standard, 28 Vdc to 48 Vdc are available).

The model 1CL6P-2 ambient compensator is designed for one HC-6/U crystal holder, the model 2CL6P-2 holds two HC-6/U crystals. Models are also available for HC-13/U holders. The small quantity price for the 1CL6P-2 ambient compensator is $10.00. Order from Isotemp Research, Inc., 1216 Harris Street, Charlottesville, Virginia 22901. Isotemp Research specializes in the design and manufacture of temperature-control sub-assemblies for electronic equipment and offers a variety of proportional-control crystal ovens in very small packages.

two-meter amplifier

The new solid-state two meter power amplifier from Dynamic Communications puts out 10 watts with a maximum of 20 mW drive. The 101-500 power amplifier operates with a 12-volt power supply and is an ideal booster for 2-meter fm walkie-talkies or as a mobile final amplifier.

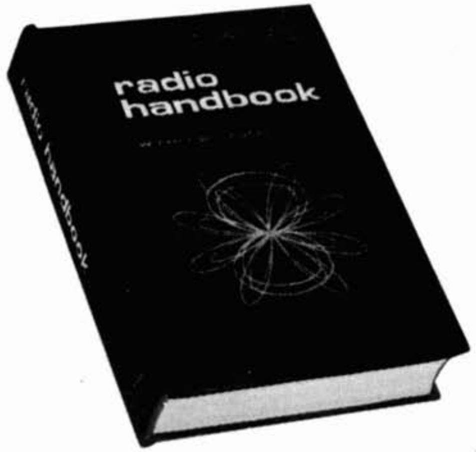
(With a 6-volt supply the amplifier puts out 4 to 5 watts.)

The DyComm power amplifier is completely broadbanded and will operate anywhere between 143 and 149 MHz with no re-tuning. The unit measures 2x2x6", including the heat-sink enclosure. $59.95 from Dynamic Communications, Inc., 301 Broadway, Riviera Beach, Florida 33404.

pulse-generator adapter

The all-new pulse-generator adapter from Blulyne Electronics Corporation is the answer for the experimenter who has a sine/square-wave generator but needs a high-speed laboratory-quality pulse generator. The new pulse generator allows you to calibrate your scope at high frequencies (50 MHz rise time), to test check amplifiers for frequency response, to control chopper circuits, or to test any electronic circuits requiring fast rise times. The generator features variable pulse width from 100 nanoseconds to 500 milliseconds (50% duty cycle maximum); pulse amplitude is variable from 0.6 to 10.0 volts, and rise and fall times are each less than 20 nanoseconds.

The unit may be used with any sine- or square-wave generator with an input from 1 Hz to 10 MHz; input impedance is 5000 ohms. Two models are available: The APG-150 with 50 ohms output impedance, and the APG-100 with 100 ohms output impedance. Price of the APG-150 is $49.95; the APG-100 is $39.95. For more information write to, Blulyne Electronics Corporation, 3 Sand Springs Road, Williamstown, Massachusetts 02167.



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principles of electronic technology

Most textbooks written for electronics fall into one of four categories: (1) Texts and manuals for engineers, usually pertaining to theory with a mathematical approach, generally at the upper college level. (2) Texts and laboratory manuals for training electronic technicians for maintenance of consumer or industrial devices. (3) Texts, handbooks and Q&A books for preparing the reader for an FCC exam, or to build and maintain amateur or commercial radio stations. (4) Hobby books presented for the casual hobbyist and basement experimenter.

These four classes of books sadly neglect a very important category: basic electronic theory, presented at a high school level of math but with attention to detail that would do honor to an engineering text. Carl B. Weick's new book, "Principles of Electronic Technology," fills the gap.

Weick has done an admirable job with his book. It goes into great depth and detail in explaining how and why the basic circuit elements (R, C and L) function. A similar treatment is given to the behavior of fundamental ac and dc circuits. Electronic devices and wiring diagrams are left to other texts. What's unique about "Principles of Electronic Technology" is the thoroughness with which it prepares the student for progression to other levels of electronic technology, whether it be maintenance, engineering, or being a true *amateur* of radio. It is best suited for study in conjunction with an organized class. Like many other books, though, a home reader with real determination can master the text quite well. This, as always, requires reading, rereading, and thinking about each element presented, working out every problem and using every review question.

Don't be too ready to look down on basic electronic theory. There are few, other than practicing engineers and instructors who deal with such topics as daily routines, who truly have a comprehensive understanding of the basic subject. Why? Because engineers touch lightly on basics only as a hasty stepping-stone to the higher and more complex subjects. Once learned, the basics are quickly pushed to the back of their minds, to be recalled, if at all, only with studied effort.

Radio amateurs often scan only the surface of the basics, grasping only those facets we believe will be of use in passing the various grades of licenses examinations. We dig a bit deeper when we plan a construction project but then only in a narrow, specialized field of our immediate interest. If you feel your mastery of electronic theory is not as complete as you'd like, "Principles of Electronic Technology" may be just what you're looking for!

power circuits manual

The latest edition of RCA's "Power Circuits Manual" has been updated and expanded to include the latest information on solid-state power devices. This new manual provides design information on a broad range of circuits that use power transistors, silicon rectifiers, and thyristors. In addition, it includes a brief introduction to semiconductor physics, as well as detailed descriptions of the construction, theory of operation, characteristics and circuit applications for each type of device.

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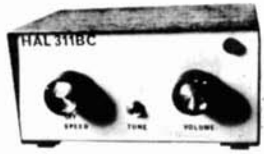

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short circuits

hcd noise blanker

In the hot-carrier noise blanker circuit on page 17 of the October, 1969 issue of *ham radio* the 47k resistor connected to pin 4 of the second Nuvistor *should* be connected to pin 8. Pin 4 is connected directly to the .005 μF bypass capacitor.

digital clock

Several parts values were omitted from the schematic diagrams of the digital clock on page 51 of the April, 1970 issue. R1 through R10 should be 20k for 170 Vdc operation of the Nixie tubes; R11 and R12 are each 270 ohms; C3 in the power supply should be 2000 μF. All gates are Fairchild μL914 or equivalent, and all flip-flops are μL923s.

Stafford Electronics advises the following price changes (first price is complete kit, price in parenthesis is for etched circuit board only): 12- or 24-hour clock, $145 ($13.50); 12- or 24-hour clock less seconds, $125 ($13.50); power supply/ clock generator, $35 ($3.00); alarm circuit, $35 ($3.00); walnut cabinet, $30.

solid-state power supplies

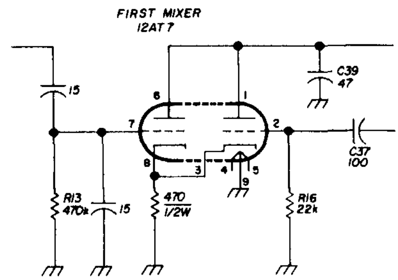
In **fig. 21** of the "Survey of Solid-State Power Supplies" article, page 36, February, 1970, pin number 10 of the MC1460G voltage-regulator IC should be connected to the common bus.

32S1 modification

The wrong crystal was listed with the 32S1 modification that appeared in the *ham notebook* on page 82 of the December, 1969 issue. The correct Collins part number for the 457.550 kHz crystal is 290-8709-00. The parts department at Collins has been alerted to this error, so if you already ordered a crystal you should receive the correct frequency.

75A-4 modifications

There was an error in **fig. 2B** of the 75A-4 modification article that appeared in the April, 1970 issue of *ham radio.* The The correct schematic is shown below.



tilt-over mast

When calculating guy-wire tension (last item in **fig. 7**, page 48, February, 1970 *ham radio*) the wind loading at the top of the mast should be divided by the sin of angle B, *not* the tangent. The sine and tangent of 9.7° (example used in text) are numerically alike and no error of great magnitude exists, but for installations where angle B is much larger the error is large.

2-meter transmitting mixer

The E. F. Johnson Company has discontinued the type 189-253-5 PC butterfly capacitor specified for the 2-meter transmitting mixer featured in the April, 1969 issue of *ham radio* (**fig. 5**, page 13). For C1 and C2, 8.5 butterflys, use Johnson 160-208. At C3, 10 pF butterfly, use Johnson 160-211; C4, 10 pF, use Johnson 160-104; C5, 20 pF, use Johnson 160-110. Capacitors C1 and C2 may be soldered upside down to the foil side of the printed-circuit board to the appropriate connection points.

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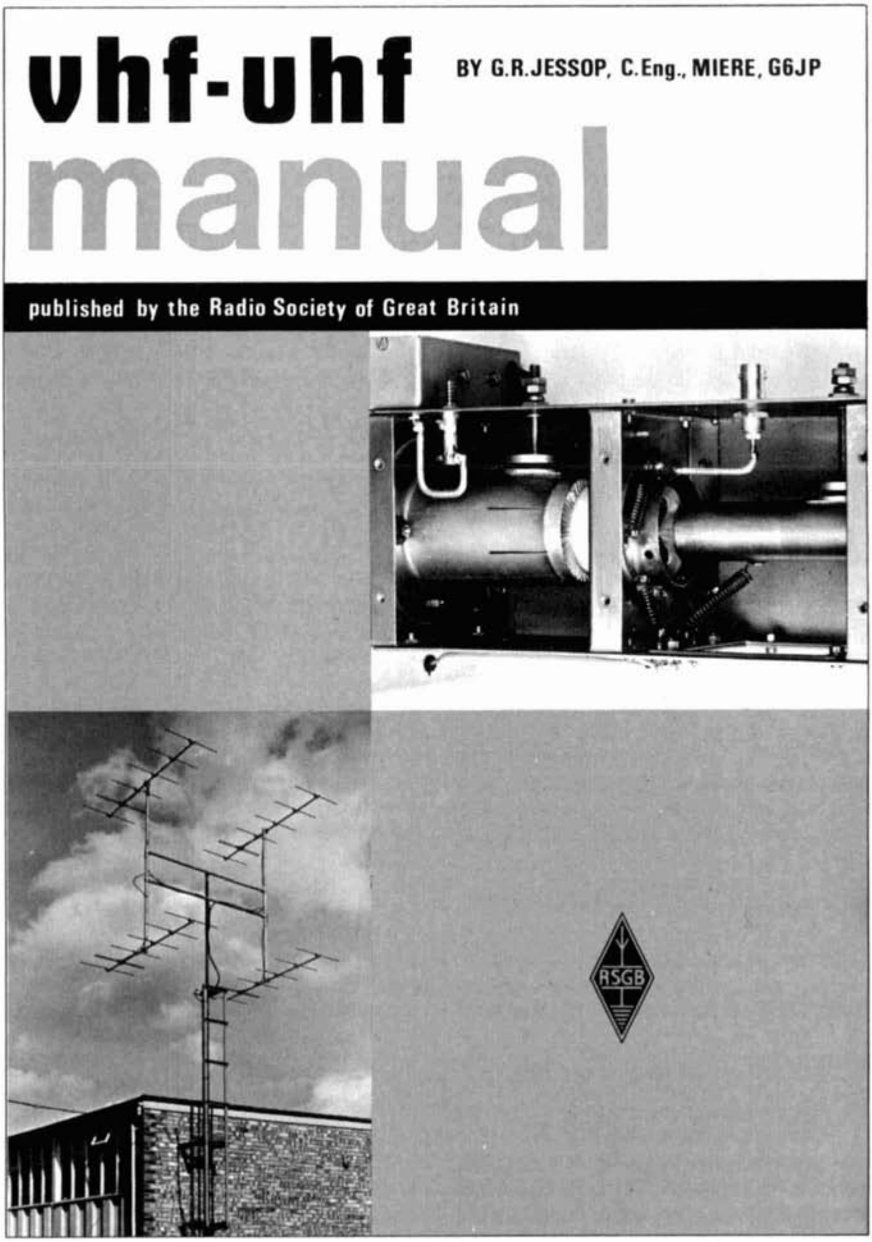

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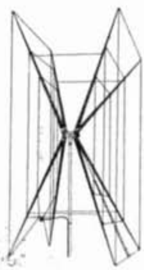

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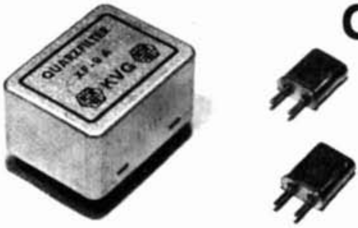

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WANTED: To buy or borrow for copy: Manuals and schematics for Collins receiver R390A/urr. M. Logan, Hilbrae House #5, Poughkeepsie, NY 12603.

FOR SALE — HY GAIN TH4 $50.00. Henry Ingwersen, Topsfield, MA 01983.

GREENE DIPOLE CENTER INSULATOR . . . see ad page 93, May, 1970 Ham Radio.

QSLS. SECOND TO NONE. Same day service. Samples 25¢. Ray, K7HLR, Box 331, Clearfield, Utah 84015.

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HAMVENTION: Albuquerque New Mexico on 18, 19 and 20 Sept. 1970. Lots of Door prizes. Gabfest, Flea Market, Technical Sessions. MARS, VHF, SSB, DX Meetings. For information and registration contact Ray Hill, W5SDM, 9016 Los Arboles Ave. NE, Albuquerque, New Mexico 87112, Phone Area Code 505 299-1719.

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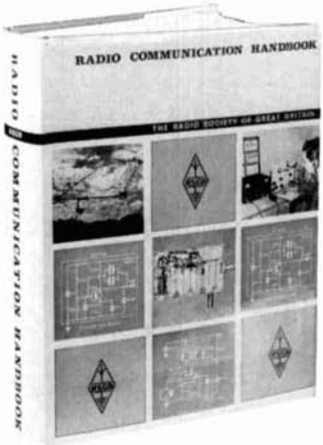

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SAROC, JANUARY 7-10, 1971, Flamingo Hotel Convention Center, Las Vegas, Nevada. Sponsored by Southern Nevada ARC, Inc., Box 73, Boulder City, Nevada. Advance registration $14.50 per person accepted until January 4, regular registration at door, includes Flamingo Hotel Late Show and drinks, Sunday Breakfast, Cocktail Parties, technical seminars and meetings, ARRL, DX, FM, MARS, QCWA, WCARS-7255, WPSS-3952 and WSSBA. Ladies Program. Flamingo Hotel SAROC room rate $12.00 plus room tax, per night, single or double occupancy January 3 thru 12, 1971. Mail accomodations request to Flamingo Hotel. Mail advance registration to SAROC. W7PRM, Club President. W7PBV, SAROC Convention Chairman.

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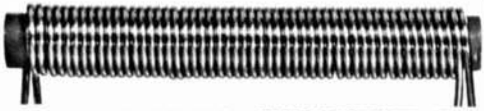

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THE ANNUAL SYRACUSE VHF ROUNDUP will be held at the Three Rivers Inn, Route 57, 10 miles north of Syracuse, N. Y., Saturday, October 10, 1970. Reservations and information from Charles Sellwood, W2RHQ, 902 1st North Street, Syracuse, New York 13208.

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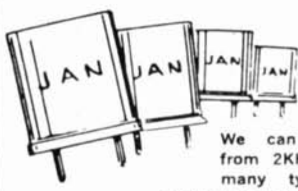

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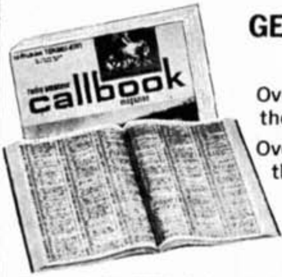

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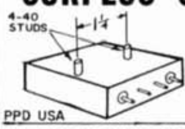

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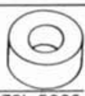

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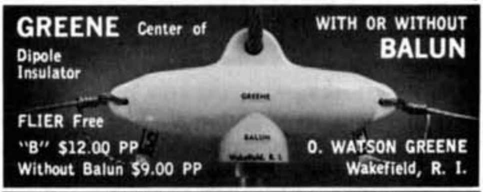

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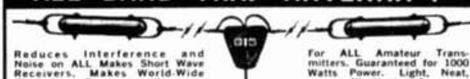

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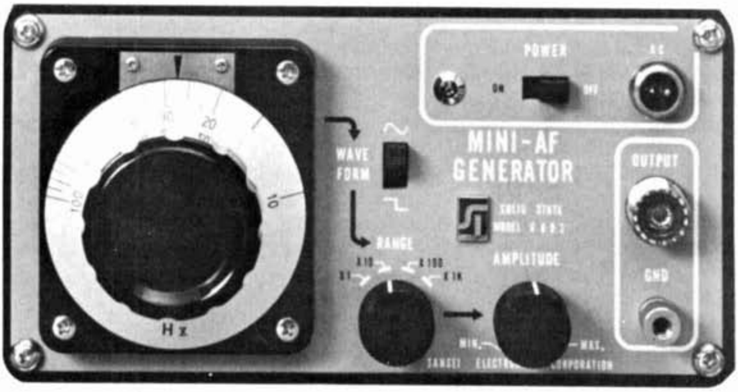

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