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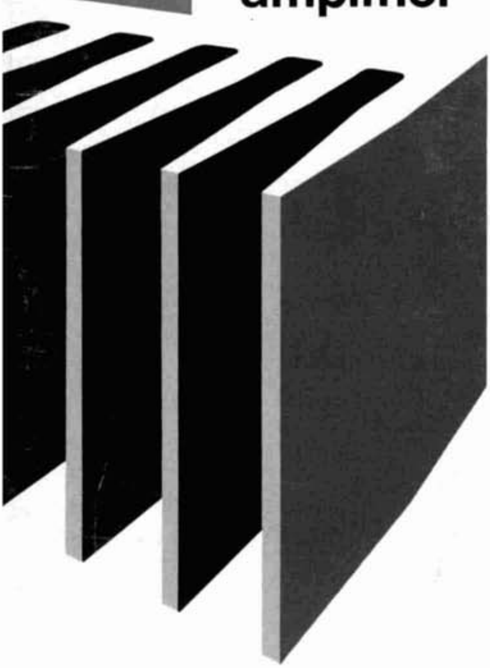
ham radio magazine

JULY 1972

conduction·cooled five·band linear amplifier



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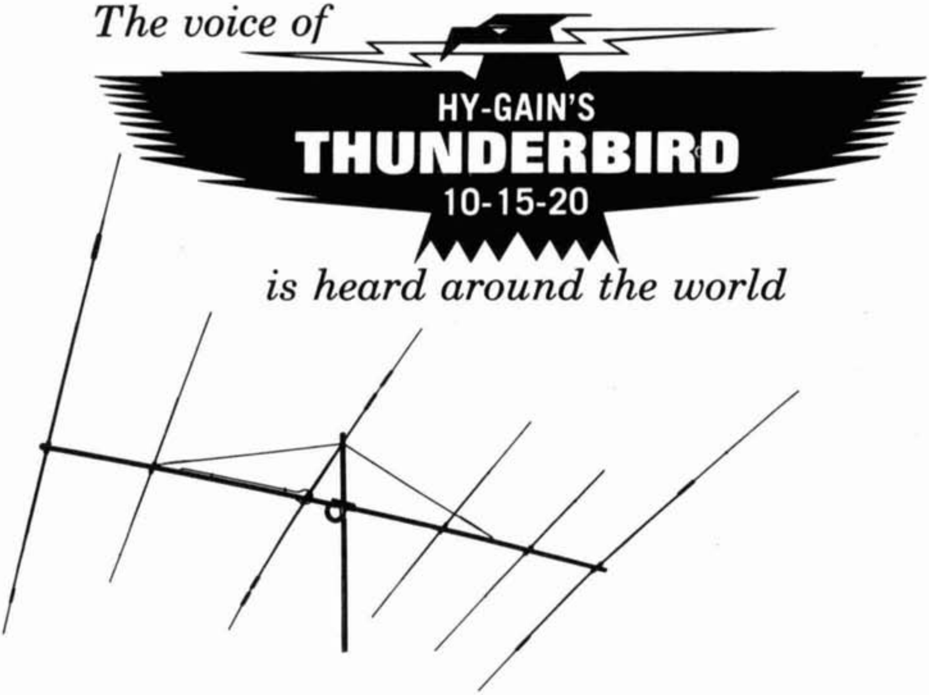

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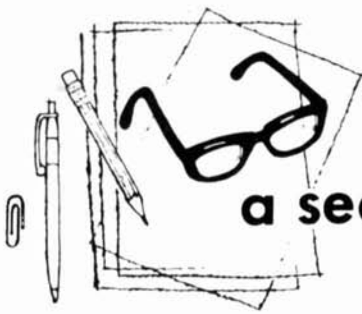

a second look

by jim fisk

Our society is a very mobile one, and when it comes to traveling long distances, most of us fly with one of the commercial airlines. It's only natural for the fm'er, with his portable fm rig, to question the possibility of using his equipment on commercial flights.

It is popularly believed that all we have to do is obtain the Captain's permission to operate; surely our little two-watt fm rig is not going to cause any interference with the high-powered radio equipment used on board the aircraft. However, this is not the case – according to the Federal Air Regulations, approval must be obtained from the air carrier (airline) and *not* the pilot in command. However, once approved by the air carrier, the permission of the Captain in command must also be obtained to operate equipment aboard a particular flight.

Shortly after World War II, portable Japanese fm broadcast receivers started appearing on the market, and passengers started using them aboard commercial flights. At the same time, aircraft navigation radios started doing funny things, and it didn't take long to determine that the interference was being caused by rf radiation from the portable fm receivers. The aircraft radios literally went wild, and at least two aircraft accidents have been attributed to interference of this type.

When it was determined that this interference was present, the FAA promulgated new regulations, paragraph 91.19 of the Federal Air Regulations. This paragraph states that *no* electronic device may be operated aboard a commercial airliner *except* heart pacemakers, voice recorders, hearing aids, electric shavers and electric watches, unless the device has been approved by the air carrier or operator. The regulation further states that *the captain of the aircraft does not have the authority to authorize such operation.*

Consider, for a moment, what might happen if such operation were allowed. Suppose you have been operating all across the country, and your plane is about to land. A passenger with a briefcase telephone sitting across from you has been watching you operate. About 10 minutes before landing, he decides to call his wife. Unfortunately, his telephone operates on a frequency right in the middle of the glide slope spectrum. As soon as his transmitter is keyed, the glide slope indicator cross pointer goes up or down, and the autopilot follows it. This could be disastrous.

As an airliner flies across the country, the pilot changes frequency every 5 minutes or so. If several fm operators are on the same flight, only one can talk at a time, so some may decide to switch to other frequencies. When you start to figure out all of the i-f and carrier frequencies of the aircraft radios, plus the

(continued on page 86)

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five-band conduction-cooled linear amplifier

A high performance table-top linear that provides noiseless operation on all bands from 3.5 to 28 MHz

Richard I. Bain, W9KIT, 4915 Ridgedale Drive, Fort Wayne, Indiana 46815

The idea of adding a linear amplifier to your station becomes quite attractive after fighting the DX dog-piles on 20, foreign broadcast QRM on 40, or QRN on 80. When I started planning a high-power linear amplifier for my station, I wanted a table-top unit that was quiet, lightweight, stable and efficient, and that would dissipate very little power on standby. I also wanted an amplifier as uncomplicated as possible; the fewer parts there are, the fewer there are to fail.

The conduction-cooled linear shown in **fig. 1** nicely fills all of these requirements. The conduction-cooled Eimac 8873 power tubes provide up to 1500 watts PEP input without a blower, so the amplifier is absolutely quiet.

The amplifier is housed in a table-top sized cabinet 24 inches long and 12 inches high. The unit dissipates little power on standby, and power output on 80, 40, 20 and 15 meters is 600 watts; on 10 meters power output is slightly less.

circuit

In the complete amplifier circuit shown in **fig. 1**, two vacuum relays are used to switch the linear in and out of the circuit. These two relays are controlled by a third, smaller relay which also shorts out the protective-bias resistor. The use of input and output switching allows use of the linear with transceivers. The relays can be energized only when the high voltage power supply is turned on.

The parasitic choke is located in th

cathode input circuit, and seems to be just as effective as if it were placed in the plate circuit.[1] The cathode and filaments are isolated from rf ground by a 3-winding rf choke wound on a ½- by 5-inch ferrite rod. The winding, consisting of two no. 16 wires for the filaments and one no. 20 for the cathode, fills the full length of the ferrite rod.

The 8.2-volt zener diode provides cathode bias to hold the two tubes to a total cathode current of approximately 50 mA. I used a 10-watt diode, which should be sufficient for up to 1500 watts input; for 2-kW operation, a 50-watt zener would be preferable. The resistor in parallel with the zener provides insurance against damage due to a floating ground in case the zener opens up. The 10k resistor in series with the zener diode provides cut-off bias during standby.

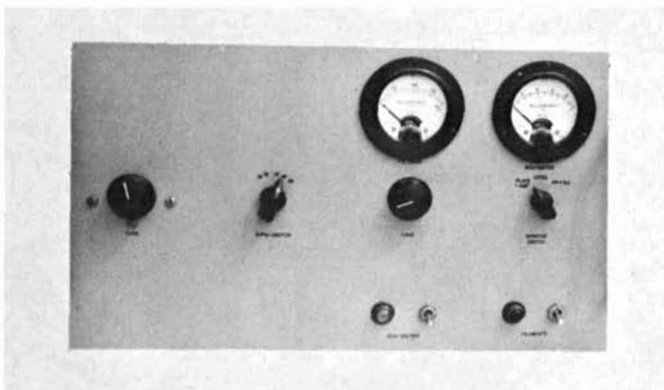
The rf plate choke I used is a surplus pie-wound unit. The National R-175A or B&W 800 are suitable substitutes. The plate switch has three sections, but only two are used. One switch section selects taps on the coil and adds extra loading capacitance on 3.5 MHz. The other section adds capacitance in the plate side in the 3.5-MHz position.

The plate-tuning capacitor is tapped down on the ten-meter coil. This reduces the effective minimum capacitance of the plate-tuning capacitor so more inductance is placed in the circuit when operating on the 28-MHz band.

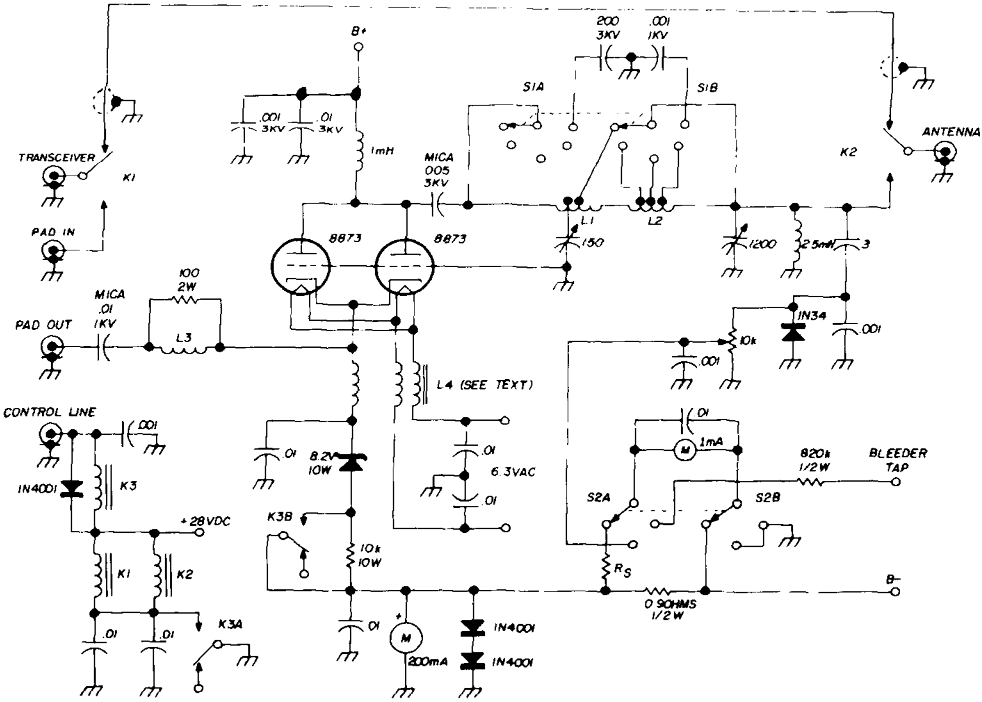
I wanted to run 1500 watts PEP on ssb, but this raised the question of obtaining reasonable tuned-circuit Q at both 1000 and 1500 watts input. However, I found that, by selecting a Q of 15 at 1000 watts, the Q was still 10 at 1500 watts; both fall within a satisfactory range. If you want to run 2000 watts PEP for ssb, the tank coil should have separate taps for CW and ssb.[1]

A second problem that had to be considered was the high fixed capacitance in the plate circuit due to tube output capacitance and stray capacitance to the heat sinks. Unfortunately, the total of this stray capacitance is sufficient to resonate the desired tank inductance on 10 meters. Therefore, I had to settle for a smaller coil, and consequently, higher Q. The sacrifice is small, however, and the 10-meter coil does not heat excessively during normal operation.

I debated about adding tuned circuits for cathode matching to gain the 1 or 2 dB advantage in intermodulation distortion, but decided to try the amplifier without them and see how it worked out. The cathode input impedance is in the 60- to 70-ohm region, a reasonable match for most exciters. A problem that popped up while using the linear was that exciter drive dropped off as a transmission progressed. I added a power-dropping attenuator pad so the exciter could be operated



The simple front panel of the conduction-cooled linear. The front panel of the finished amplifier has handles, facilitating moving, handling and servicing the unit.



L1	5 turns 3/16" copper tubing, 1¾" diameter, 2" long, tapped at 4 turns	K1,K2	28 Vdc vacuum relays
L2	Air-Dux 2408D4 tapped at 1, 4, 7 and 13 turns from coarse-wound end	K3	miniature 2-pole 28-Vdc relay

Heath sinks are 6" pieces of extruded stock, Thermal Associates no. S-511

fig. 1. Schematic diagram of the five-band 1500-watt conduction-cooled linear amplifier. Resistor Rs is chosen to give 1 ampere full-scale deflection.

at normal input levels where the output remains stable. The attenuator circuit is shown in fig. 2.

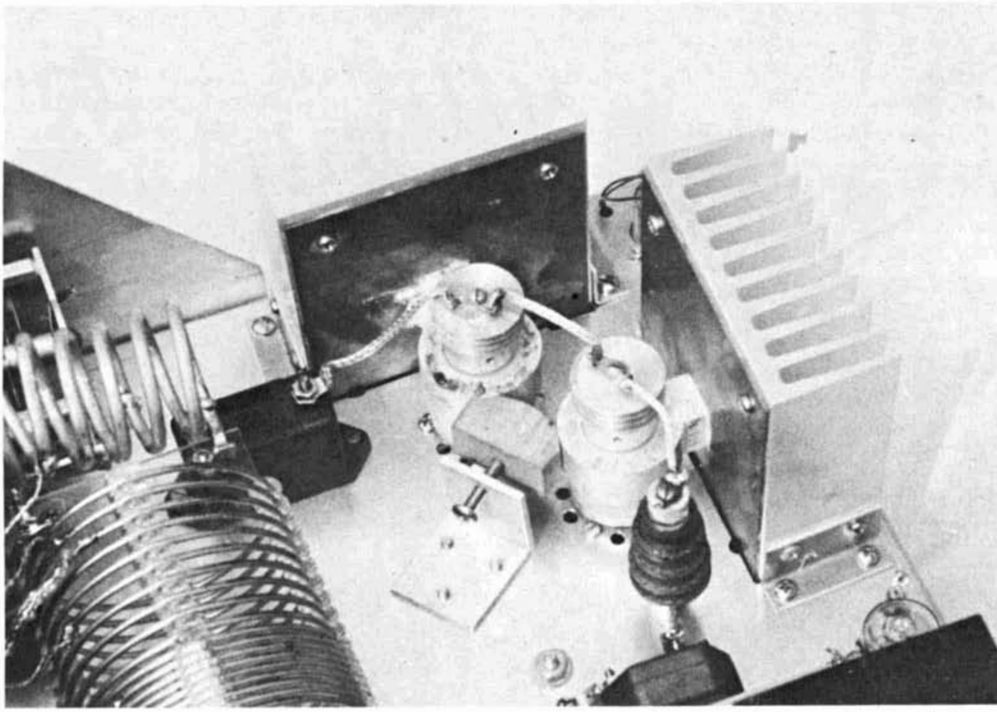
The attenuator is housed in a 2-by-3-by 5-inch minibox, with numerous ventilation holes. The pad could have been built on the amplifier chassis, but by being able to patch it in and out, it can be replaced by a coax jumper for operation with low-power exciters. The amplifier only requires about 16 watts drive for 1000 watts input on the lower bands, and slightly more on the higher bands.

One advantage of the pad is that the exciter sees a resistive 50-ohm load. Also, the pad tends to swamp out variations in cathode impedance due to plate tuning, so the exciter doesn't have to be returned each time the linear is adjusted.

construction

The amplifier shown in the photos was built on a 17- by 13- by 3-inch chassis with a panel 22 inches long by 11 inches high. The panel size was chosen to fit a surplus cabinet I had on hand. The chassis is supported by two side brackets to prevent excessive flexing.

The heat sinks used to cool the tubes are 6 by 4 inches with 1½-inch cooling fins. Thus, each heat sink provides 36 cubic inches of cooling volume. The heat sinks are faced with 1/16-inch copper to provide improved heat transfer from the vicinity of the thermal link blocks to the rest of the heat sink. The heat sinks are sub-mounted 1½ inch below the top of the chassis so the thermal links are



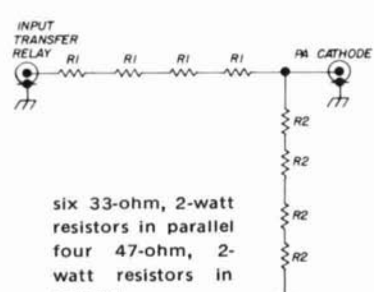
Close-up photograph shows the details of the final-tube mounting. Careful alignment of the tubes, heat sinks and clamping block is necessary to assure efficient and equal cooling of both tubes. To assure adequate ventilation of the heat sinks, drill plenty of holes in the equipment cabinet.

centered. Slots on the chassis sides beside the heat sinks allow free air movement up through the cooling fins.

The tube socket mounting holes are elongated to allow alignment of the flat face on the tubes with the heat sink surfaces. Also, the mounting holes in the aluminum angle brackets that hold the heat sinks in place are oversized – this allows some tilt adjustment of the heat sinks.

The alignment of the tubes and sinks must be nearly perfect, or the thermal link blocks will not sit flat on the sink, and heat transfer from the tubes to the heat sinks will be impaired. Thermal compound should be used liberally between the tubes and thermal links, between the thermal link and copper plate, and the copper facing.

The 8873s are held tightly against the heat sinks by a semicircular piece of diallyl phthalate, which is clamped in place by a number-10 screw through a hole tapped in a piece of aluminum angle bracket. The material used for the clamping block must be a good electrical insulator and be able to take temperatures up to 200° C. A tapered ceramic insulator might be used for this purpose. The chassis holes for the angle bracket are slotted so the bracket can be rotated to properly align the clamping block to



R1 six 33-ohm, 2-watt resistors in parallel

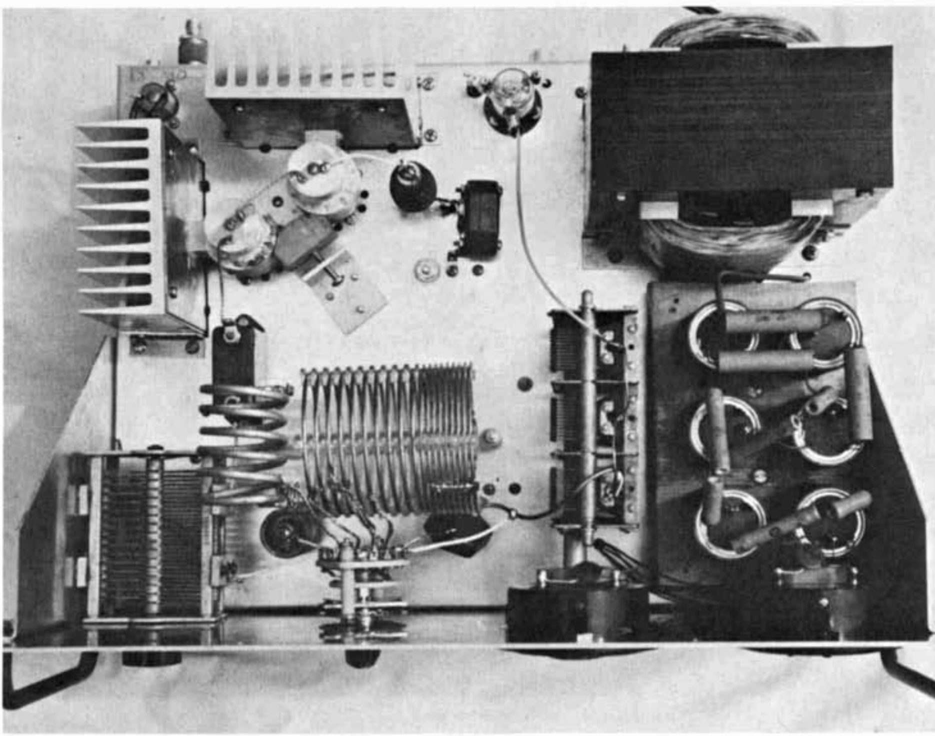
R2 four 47-ohm, 2-watt resistors in parallel

fig. 2. Power attenuator for use with high-power exciters. The amplifier requires only 16 watts drive for 1500 watts PEP input.

insure equal pressure on the two tubes.

It would, of course, be possible to mount the heat sinks in-line and clamp each tube individually with a toggle clamp.[2] However, regardless of the mounting you use, external air must be able to enter the cabinet, pass through the transformer. The high voltage drops about 400 volts with a 500-mA load when operating from 117 Vac. Operation from 220 Vac should result in better regulation due to reduced primary current.

My diode rectifiers generate a small



Overview of the complete amplifier and power supply show the clean layout and overall simplicity.

the heat sinks and exhaust, all with ease. I put several rows of ¼ inch holes in the cabinet above and below each heat sink. The cabinet must be mounted on at least ¾ inch feet to allow air to enter through the bottom.

power supply

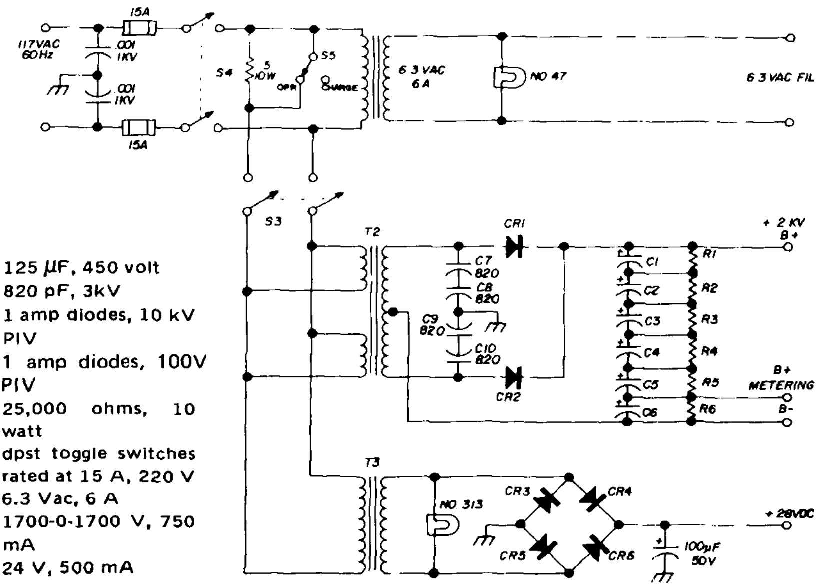
The high-voltage power supply has six series-connected capacitors for a total of 21 μF. This amount of filtering is adequate for up to 1500 watts PEP, but if you want to run 2000 watts input, more capacitance should be used. The rectifiers are 5000-volt bridges connected to give a 10 kV PIV diode array on each side of

amount of hash, probably due to switching transients. Capacitors on the primary and secondary side of the power transformer keep most of the hash out of the 117-Vac line. Filter chokes may be needed for tougher cases.

The bypass capacitors also protect the diodes from large amplitude, narrow spikes which may be present on the ac line. The 28-volt supply provides power to the high-voltage indicator and the relays (see **fig. 3**).

The high-voltage supply is metered by tapping off the top of the bottom bleeder resistor. Since the bleeder resistors are not precision types, the actual value of the high voltage should be measured by a more accurate means to obtain a correction factor; alternately, the value of the series-dropping resistor may be selected to make the meter read correctly.

The five-ohm resistor and S5 were added after six months of operation. I found that the surge current through S3 caused the switch contacts to stick closed. Closing S5 after S3 protects S3 and the high-voltage diodes.



C1-C6	125 µF, 450 volt
C7-C10	820 pF, 3kV
CR1, CR2	1 amp diodes, 10 kV PIV
CR3-CR6	1 amp diodes, 100V PIV
R1-R6	25,000 ohms, 10 watt
S2, S3	dpst toggle switches rated at 15 A, 220 V
T1	6.3 Vac, 6 A
T2	1700-0-1700 V, 750 mA
T3	24 V, 500 mA

fig. 3. Power supply for the five-band linear amplifier.

tuning and adjustment

The amplifier should be tuned to achieve maximum output at a given plate-current level while insuring that grid current does not exceed the normal operating value. When you have determined the normal settings for the tuning controls, it is easy to use these as starting points when changing bands. The amount of grid current at a given amount of plate current is determined by the setting of the loading control. With light loading, grid current will be excessive, while with heavy loading, grid current will be small, and efficiency will be poor.

There is some problem in determining just what the grid current should be under normal conditions. This depends upon the percentage of grid intercept. The manufacturer says that this is nominally 10%, but may be somewhat lower, or as high as 15%. This means that the grid current could run a bit higher than that listed on the manufacturers data sheet, and still not be excessive.

The best way to determine the correct grid current for a given power input is to check power output. My linear delivers 600 watts output on all bands except ten meters where it is a bit less. Allowing more than minimum grid current to achieve this power level does not result in more output, so this minimum grid current is my operating value. It is, of course, desirable to use a scope when operating and tuning up on ssb.

heat tests

Heat tests were run to determine the temperatures that the tubes and heat sinks reach during normal operation. The temperatures were checked with temperature-sensitive compound sticks known as *Tempilstiks.** The heat sinks were allowed to stabilize with just the filaments running before the heat-test runs were started. Static dissipation tests were made by reducing the bias on the tubes until each tube was dissipating 200 watts.

The anode temperature (on the side of the tube away from the sink) reached 200° C in 8 to 10 minutes. There was a 50° C drop from the tubes to the heat sinks. The tubes must not be operated above 250° C, so a limit of 200° C under normal operating conditions is reasonable.

Tube temperature seemed to slowly increase beyond 200° C. This indicates that key-down operation for periods longer than ten minutes would require some additional cooling measures, such as a thermostat on the heat sink to turn on a fan. The thermal link blocks used on my tubes are 3/8-inch thick, but thinner blocks are available for improved heat transfer. However, thinner blocks add more stray capacitance to the plate circuit.

Since the heat sinks are attached to the chassis, it receives conducted heat. Therefore it is wise to avoid mounting parts that could be damaged by the elevated temperatures close to the heat sinks. The hottest spot on the chassis is in the area around the tubes.

It was noted while running the static heat tests that one tube drew about 10 mA more than the other, so it was running much hotter. Diodes were added as shown in **fig. 4** to place more bias on the higher current tube. The reverse diode provides a current path during the negative half of the rf cycle. This simple addition seemed to achieve balance between the two tubes.

Tube operating temperatures were also checked during operation. Under extended operating times, such as a CW ragchew, tube temperatures did not reach 200° C. Likewise, ssb operation at 1500 watts PEP did not result in excessive tube temperatures. Even hot and heavy contest operation doesn't overheat the tubes.

**Tempilstiks* are manufactured by the Tempil Company, 132 West 22nd Street, New York, New York 10011.

operation

The linear should be placed so air can freely enter and leave the cabinet since cooling relies primarily on air moving by convection upward through the cooling fins. The cabinet should be mounted on feet, ¾ inch or higher, to allow free entry of air into the bottom of the cabinet.

The thermal links should be checked from time to time to insure that they have not changed position. It might be possible to place a ridge of solder or epoxy on the heat sinks around the location of the blocks to prevent creeping.

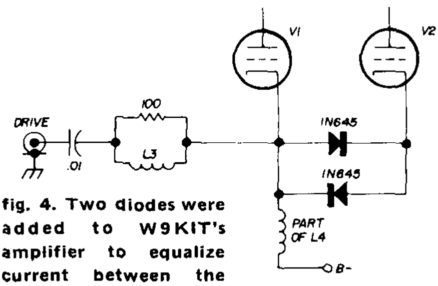


fig. 4. Two diodes were added to W9KIT's amplifier to equalize current between the two final-amplifier tubes (see text).

I have found this linear amplifier to be quite stable. The only problem I had was my homebrew exciter taking off on 80 meters. This was due to a poor ground connection between the exciter and the linear, and was cured with a ground strap between the units. I am well satisfied with the amplifier, and find that it fulfills all my operating requirements.

references

1. Douglas A. Blakeslee, W1KLK, Carl E. Smith, W1ETU, "Some Notes on the Design and Construction of Grounded-Grid Linear Amplifiers," *QST*, December, 1970, page 22.
2. Robert I. Sutherland, W6UOV, "Using the Eimac 8873 Zero-Bias Triode," *ham radio*, January, 1971, page 32.

ham radio

crystal-controlled AFSK generator

Using one oscillator with a variable frequency divider, allows crystal-controlled AFSK without envelope aberrations

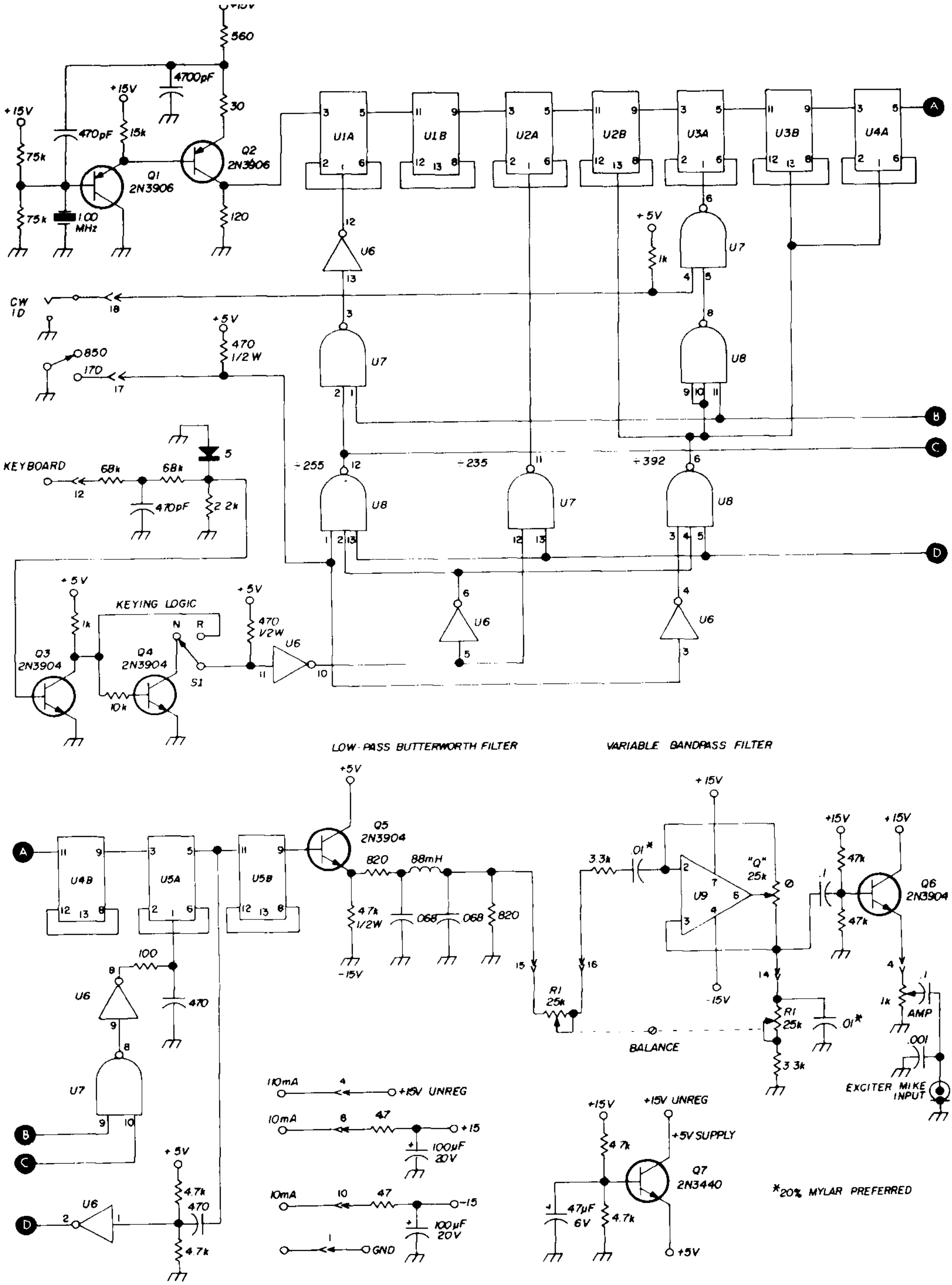
Chuck Barrows, K7BVT, 5541 SW Miles Court, Portland, Oregon 97219

Several crystal-controlled AFSK circuits have been described using two separate oscillators and a common digital dividing circuit.[1] The circuit presented here is unique in that it uses a single oscillator and varies the division factor of the counter circuit. This has the inherent advantage that transitions between tones always occur at zero crossings, thus eliminating envelope aberrations often generated when switching between two non-synchronous tones.

I used a 1-MHz crystal because these crystals are readily available on the surplus market. Dividing 1 MHz by 2.125 kHz yields a ratio of 470.59. Since the counter can only divide by an integer I use a compromise ratio of 470, resulting in a output tone of 2.1277 kHz; 2.7 Hz higher than the standard mark frequency. Starting with a 100-MHz oscillator and dividing by 4,706 would give a 2.12504-kHz output, but the cost of ICs operating over 10 MHz and the problems of 100-MHz circuitry would not justify the added accuracy.

I used a standard ripple counter for the frequency-divider circuit (see **fig. 1**). These counters give an output with a non-symmetrical period when dividing by ratios other than two, four, eight and so forth. Since a non-symmetrical period represents even harmonics, the filtering requirements are eased by designing the basic counter to yield twice the desired frequency and then dividing again by two in a simple toggle circuit.

The output at pin 5 of U5 is differentiated through the RC network at pin 2 of U6 providing a *clear* pulse for the appropriate stages to provide the different ratios. For instance, when 2.127 kHz is required each time pin 5 of U5 goes negative, a pulse is coupled through the NAND gate logic scheme to the 1, 4, 16 and 256 stages. The counter *will start* each cycle with a count of



Q7	2N3440, TO-5 case, 1W	U6	SN7404 Hex inverter
R1	dual 25k potentiometer, single shaft	U7	SN7400 Quad 2-input NAND gate
U1, U2, U3, U4, U5	SN7474 Dual D-type edge-triggered flip-flop	U8	SN7410 Triple 3-input NAND gate
		U9	µA741 Operational amplifier

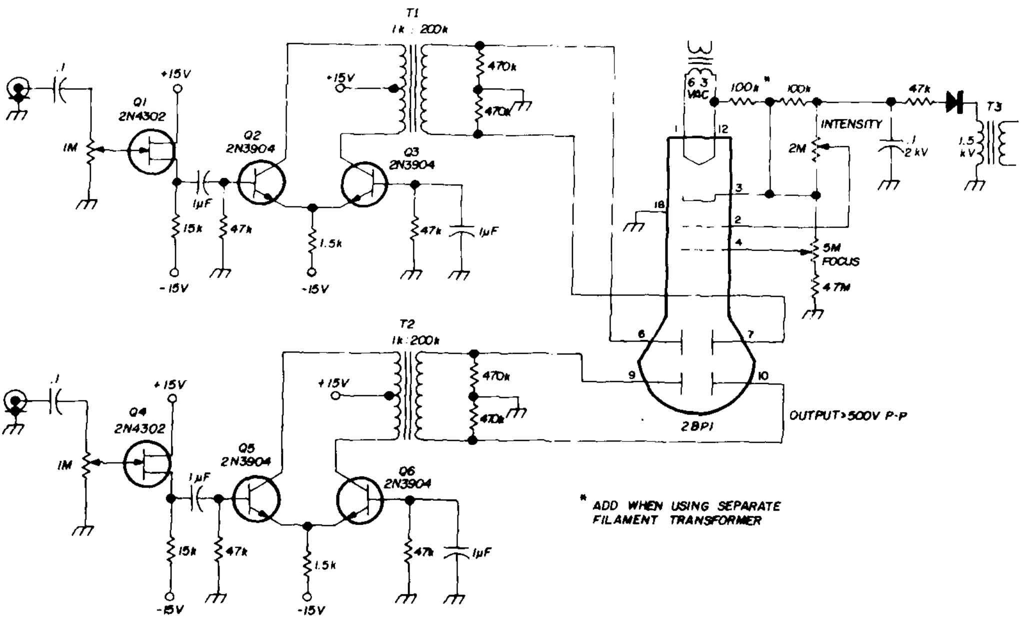


fig. 2. Transistorized, transformer-coupled deflection amplifier. The coupling transformers, T1 and T2, are inexpensive imported audio interstage types. T3 is a high-voltage scope transformer.

1+4+16+256=277, thus pin 5 of U5 will go negative at an actual count of 511-277=234. The output of U5 will be a non-symmetrical square wave with a frequency of 4.154 kHz providing a 2.127 symmetrical square wave at pin 9 of U6. Since the *clear* pulse is derived from the 256 stage, an RC network is placed ahead of the 256 clear input to insure enough delay for proper self-clearing action.

When a keying transition appears at the input (the network connected to the base of Q3), the counter ratio cannot change until the cycle in progress is completed.

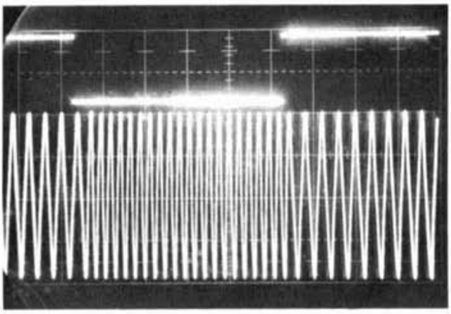
The output of the counter is fed to a 3-pole Butterworth low-pass filter with a −3 dB point of 2.5 kHz. The low-pass filter is followed by an active variable band-pass filter.[2] Tuning this filter between the mark and space tones provides amplitude compensation to counteract unequal attenuation of the two tones in the low-pass filter and the filter in the ssb exciter.

fig. 1. (opposite page) Crystal controlled AFSK generator. On every IC, U1 through U8, pin 14 is connected to +5 V and pin 7 is grounded. Divider resistors are chosen to give 5V at the emitter of Q7. Unless labeled, all resistors are quarter watt.

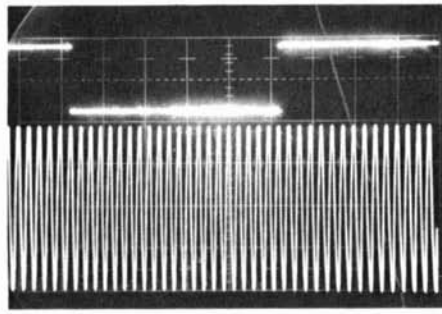
construction

The entire circuit was built on a 22-pin plug-in Vectorbord. I used a BC-929 cabinet to enclose the AFSK circuit as well as a version of the ST-6 demodulator.[3] The ST-6 ±15 V supply provides enough current for the AFSK, ST-6 and a transistorized transformer-coupled deflection amplifier (see **fig. 2**) that drives the original BC-929 3-inch CRT.

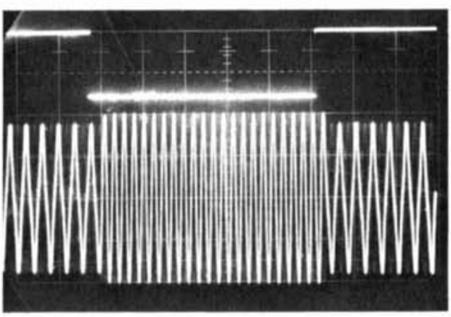
The circuit is fairly complex for the average amateur with nine ICs and seven transistors. The new Motorola frequency-synthesizer ICs (MC4318) would simplify the design, but at least two would be required at $7.50 each versus less than $10 for all of the TTL ICs through surplus outlets. The total cost of components through surplus houses is around $25. The 5-volt supply is marginal and was used for economy. An independent rectifier running from a 6-Vac filament transformer with a Zener diode controlling Q7 would be a definite improvement.



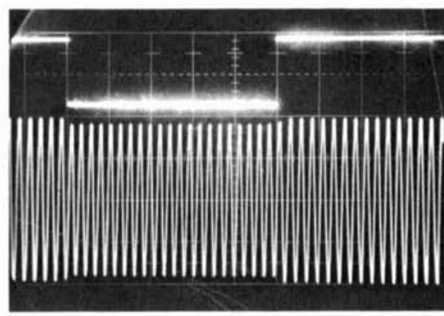
Wide shift adjusted for equal tone amplitude.



Narrow shift adjusted for equal tone amplitude.



Wide shift adjusted to boost high-frequency tone.



Narrow shift adjusted to boost low-frequency tone.

fig. 3. Performance of the AFSK generator. An electronic pulse generator was used to key the AFSK unit for these photos. All photographs were taken from a Tektronix type 547/1A4 50 MHz oscilloscope.

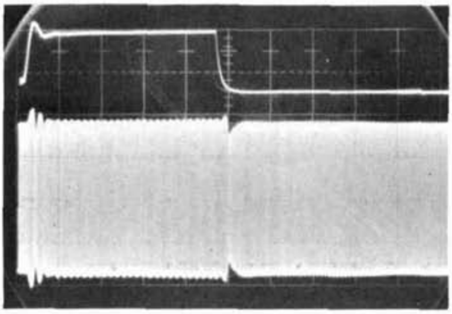
summary and criticism

The AFSK unit has been used to drive a HeathKit HW-100 for several months with very satisfactory results. While the output tones are not exact, they are more accurate and stable than typical LC or RC oscillators. For normal operation, I use upper sideband and I adjust my ST-6 tuned circuits to match the tones I use: mark–2125, space–1275, for 850 Hz shift; mark–2125, space–1955, for 170-Hz shift. The CW identification shift is 132 Hz. Since the change between narrow and wide shift requires only a ground closure, it is simple to control the AFSK shift and the shift of the demodulator with a single switch.

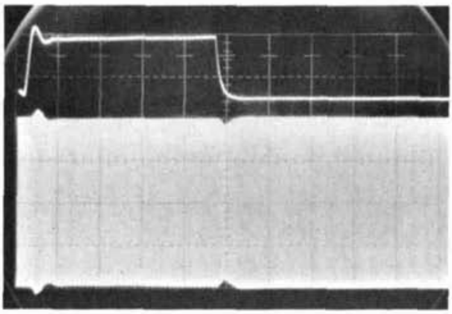
The low-pass filter does not remove the third harmonic of the 1.275 kHz tone completely, and this is compounded when the band-pass filter is aligned to boost the 2.125 kHz tone. An additional 88 mH toroid would provide a 5-pole filter which would reduce the third harmonic, but the ssb exciter filter seems to prevent radiation of the third harmonic.

While the audio envelope does not have any aberrations and tone transitions occur at zero crossings, there is some ripple on the rf envelope during transitions. This is probably due to attenuation of audio sidebands in the sharp ssb filter.

The speed of transitions between tones could be slowed, using an RC oscillator with a damping network in the frequency determining voltage loop for instance, thus reducing the audio sidebands and the rf envelope aberrations. This would increase the rise and fall times of the demodulated waveform, thus reducing the effective width of the mark and space



Rf envelope using wide shift.



Rf envelope using narrow shift.

fig. 4. Rf envelope pictures taken with HW-100 driving a Cantenna, 75 W at 3.6 MHz.

pulses, resulting in less immunity to noise, particularly on 100-wpm signals.

Thanks to W7GNI for many helpful circuit design suggestions, and to K7TBQ and WB6BZW/7 for help in evaluating the circuit.

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resistance-capacitance oscillators

Replacing the large inductors of the traditional L-C circuit with resistors and capacitors for today's miniature audio oscillators

Hank Olson, W6GXN, Box 339, Menlo Park, California 94025

The trend to microcircuits and its resultant reduction in size of electronic equipment has been accompanied by a move to eliminate inductors. This is not only because inductors (at least those with henries of inductance) are large and heavy, but because the inductor is one component that will apparently not be put on a silicon chip. (At vhf and uhf where very small inductances are useful, inductances *can* and *are* being made using etched circuit and metal-deposited-on-substrate techniques.)

Because of the engineering drive to get rid of the iron, new ideas for circuits using resistances and capacitances instead of inductances and capacitances are rapidly coming to the fore. One of the areas where R-C replaces L-C is in filtering. The techniques for systematic multiple-pole low-pass, high-pass, and bandpass filters using only resistances, capacitances and IC operational amplifiers are pretty well established.[1] Tuned transformers associated with ratio detectors and discriminators are also on their way out, having been designed around by IC phase-locked loops and other IC circuits for fm demodulation.

Another area wherein R-C circuits are widely used is oscillators; unlike the circuit designs mentioned above, R-C oscillators are not new technology. R-C oscillators have been with us for decades, at least for audio frequencies.[2] The phase-shift oscillator, bridged-tee oscillator, twin-tee oscillator, and Wien Bridge oscillator (with vacuum tubes) are good examples. Circuits of these are shown in **fig. 1** as they were originally used, with vacuum tubes as the gain blocks. Only the equivalent ac circuits are shown; that is, no blocking capacitors or biasing resistors are included.

basic circuits

The circuits presented in **fig. 1** have only one thing in common—all are R-C types. These circuits differ significantly in details. The phase-shift oscillator and the twin tee oscillator shown in **figs. 1A** and **1B** have no amplitude control system, and require their R-C networks to shift 180° at the oscillation frequency. The bridged-tee oscillator shown in **fig. 1C**, however, uses the R-C network as a null in the negative feedback path to increase the loop gain at the oscillation frequency. Positive feedback is *broadband* and provided by the resistance of I1 in **fig. 1C.**[3] The phase shift of the bridged-tee network is 0° at the null frequency. Like the bridged-tee oscillator, the Wien Bridge oscillator has an R-C network which provides 0° phase shift at the oscillation frequency. The R-C network, however, provides a peak instead of a null at the oscillation frequency, and so is placed in the *positive* feedback loop.

In **figs. 1A** and **1B** no attempt is made to automatically control the amplitude of oscillation. For this reason, these two circuits will generally produce a somewhat distorted sine-wave output.

The circuits of **figs. 1C** and **1D,** however, have a non-linear resistance (I1) which increases its resistance with an

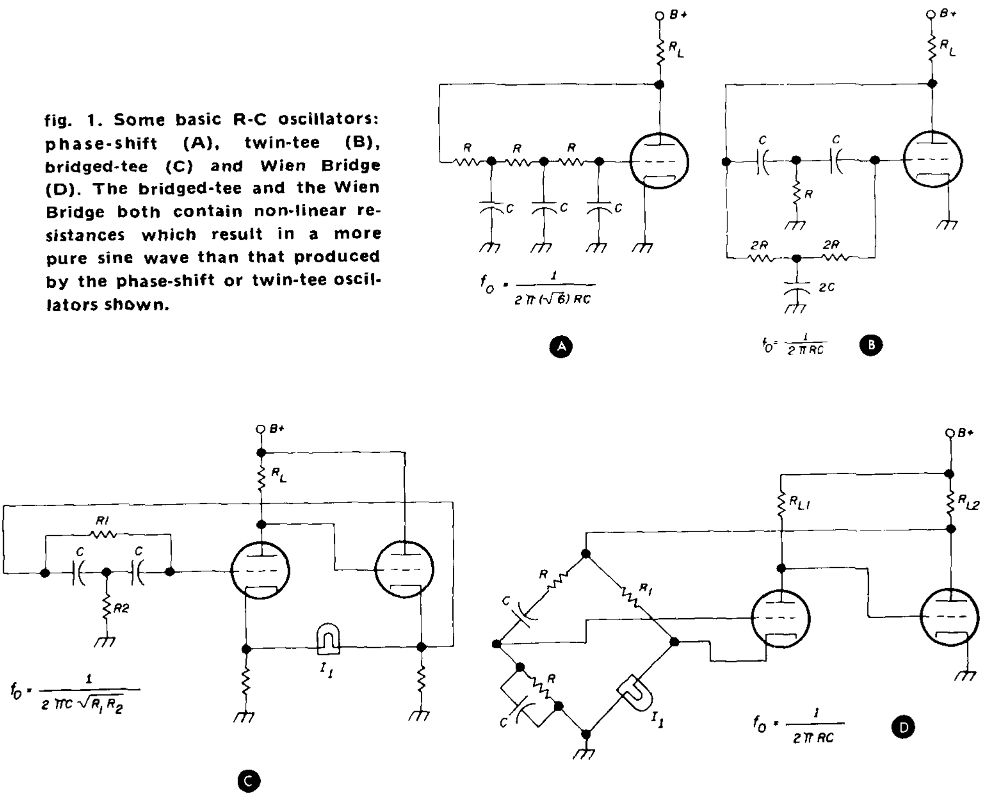

fig. 1. Some basic R-C oscillators: phase-shift (A), twin-tee (B), bridged-tee (C) and Wien Bridge (D). The bridged-tee and the Wien Bridge both contain non-linear resistances which result in a more pure sine wave than that produced by the phase-shift or twin-tee oscillators shown.

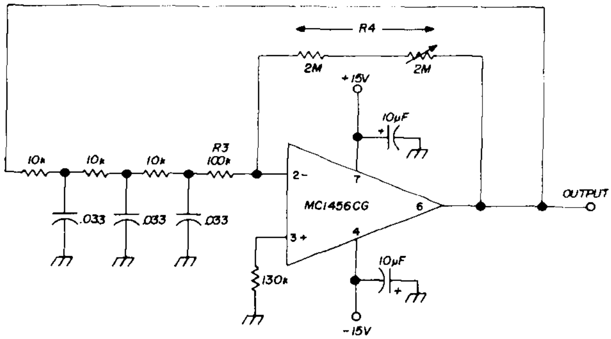

fig. 2 Phase-shift oscillator using monolithic IC op amp. The phase-shift section is composed of the three 10k resistors and the three 0.033 µF capacitors at left. All pin numbers refer to the TO-5 package.

increase in output level. The series position of I1 in **fig. 1C** assures that if the output level increases the effect will be to *decrease* the positive feedback factor. The parallel position of I1 in **fig. 1D** has the effect of increasing the negative-feedback factor upon an increase in output. In both cases the effect of I1, in increasing resistance with an increase in level, is to restore the operating condition that existed before the change.

The two oscillator circuits, with a nonlinear resistance for feedback control, produce very nearly sinusoidal waveforms, since the automatic feature holds them in the linear operating region. For this reason, these circuits have been widely used as laboratory audio generators; the famous Hewlett Packard 200 series is based on the Wien-Bridge circuit. Similarly, the bridged-tee is used in the Heathkit IG72 audio generator. It is also feasible, however, to apply amplitude control to the circuits of **figs. 1A** or **1B.**

op amps

Let us now redraw the four R-C oscillators using operational amplifiers as the gain-blocks. These are shown in **figs. 2, 3, 4** and **5**. The resistor R3 establishes the input impedance into which the R-C network must operate. Also, in *all* of these circuits, the ratio R4/R3 determines the closed-loop gain of the amplifier.

In the basic phase-shift oscillator, there are three R-C sections, each with 60° of phase shift at the operating frequency. These phase shifts all add up to 180°, causing the input and output to be 360° out of phase so that oscillation will start if the gain of the amplifier is adequate.

The twin-tee oscillator operates in much the same fashion as does the phase-shift oscillator. At the frequency of oscillation, there is 180° phase shift through the R-C network. The twin-tee network will be remembered by some readers as a notch network, which was often used in a-m radios to suppress the 10 kHz interstation whistle in early hi-fi tuners. In the oscillator use, however, it never quite operates in the notch, or no positive feedback could occur.

In the bridged-tee oscillator of **fig. 4**, the amplifier has an all-pass network (resistive voltage divider) in the positive-feedback path; so that except for the presence of negative feedback, it could oscillate at *any* frequency. However, the bridge-tee network is a *null* network, and so gives *minimum* negative feedback at its *null* frequency. This minimum negative feedback at the null frequency means that the amplifier has its maximum gain at that frequency, and so that's where the oscillations occur.

The Wien Bridge as shown in **fig. 5** is quite different in that the R-C network does not control the negative feedback but rather the *positive* feedback. The negative feedback is controlled by R3 and R4.

All the circuits of **figs. 2, 3, 4** and **5**

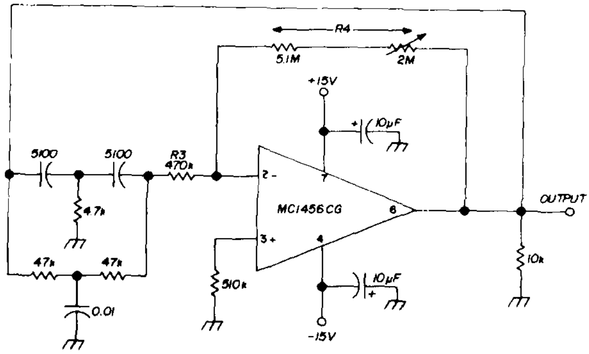

fig. 3. Twin-tee oscillator using an IC op amp.

use relatively low impedance values in their R-C frequency-controlling networks because of the low input impedances of the monolithic IC op amps they use. By using op amps with higher input impedances (such as the fet-input types), it is possible to make the C values smaller. This seemingly tiny advantage is really quite significant; it allows some important circuit possibilities: use of 15 to 468 pF dual- or triple-section variable capacitors for tuning, use of varactors for tuning, and use of on-chip capacitors. A few of the things that can then be constructed are: a capacitively-tuned continuously variable R-C oscillator, a voltage-variable R-C oscillator (for generating fm) and an R-C oscillator entirely built on a silicon chip.

variations

There are also some important variations on the basic oscillators that are worth looking at. The phase-shift oscillator, for instance, need not have the simple form of **fig. 1**, wherein all the Rs and Cs have equal values. A useful variation is that of tapering the R-C network.[4] This method uses a larger value of R and a smaller C in each successive R-C section. In this manner, the C of the first section is not loaded by the R of the following section (see **fig. 6A**). If R3 is much larger than R2, R2 is much larger than R1, C1 is much larger than C2, and C2 is much larger than C3, true tapering results. By "much larger than," I usually mean ten times as large. However, five times as

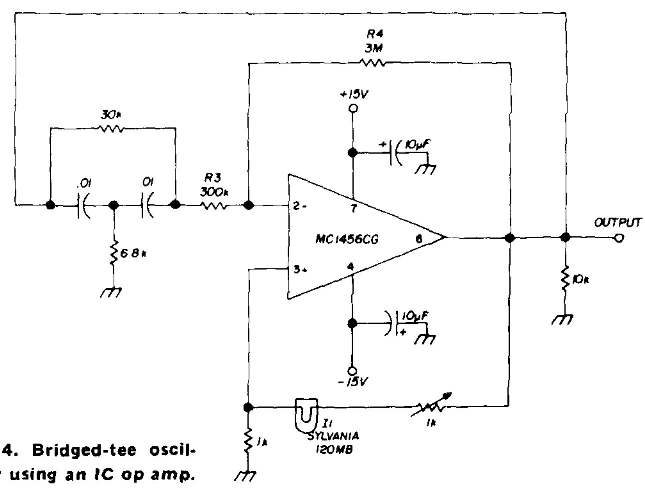

fig. 4. Bridged-tee oscillator using an IC op amp.

large is adequate for tapering in a phase-shift oscillator.

Fig. 6B shows another variation of the phase-shift oscillator network, using four sections. This variation can be extended to as many sections as desired.

Finally in **fig. 6C**, the Rs and Cs are exchanged in the circuit, and the phase-shift oscillator still operates. This works because it doesn't matter whether there is 180° lead or lag for oscillations to take place. This network may have some advantages in some circuits since it makes coupling capacitors and the amplifier input resistor part of the network. A concrete example of this is shown in **fig. 7.**

Fig. 7 also combines the techniques of *multiple sections (more than 3), and* tapering, to show that these several variations are all compatible. Tapering and increasing the number of sections both have the effect of reducing the *loss* in the R-C network, allowing us to use less amplifier voltage gain to sustain oscillation. A simple, three-section, equal R and C phase-shift oscillator takes a minimum voltage gain of 29 to make it oscillate. A four-section tapered network may require a voltage gain of less than 5 to oscillate.

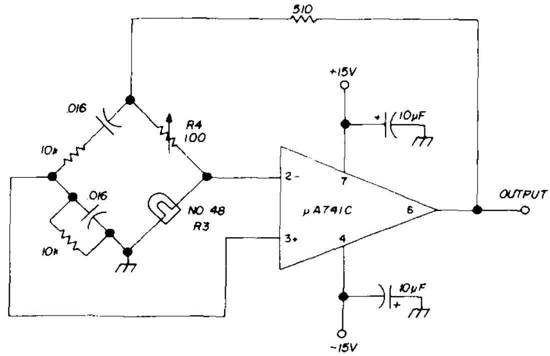

fig. 5. Wien-Bridge oscillator using an IC op amp. All pin connections refer to the TO5 case.

A rather interesting modification of the phase-shift oscillator uses a *distributed* R-C line as the network.[5] This

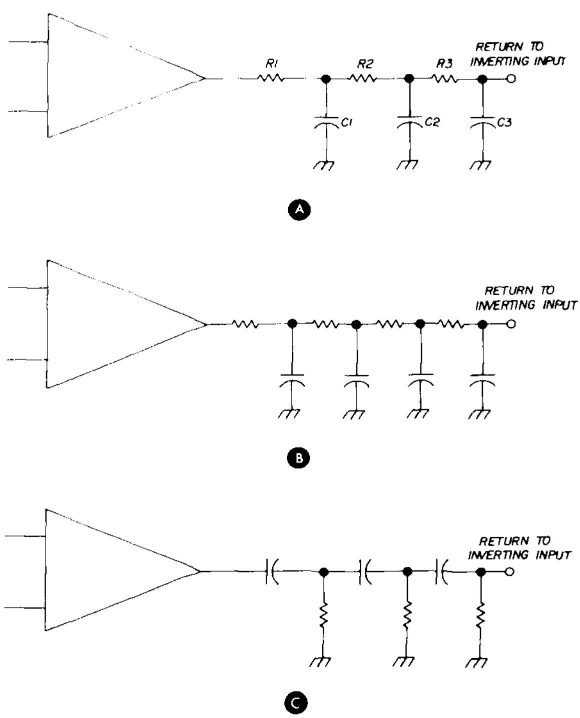

fig. 6. Variations of the basic oscillator. The tapered network is shown in A. C1 is much greater than C2 which is larger than C3. Similarly, R3 is greater that R2 which is greater than R1. B shows an extra section added to the R-C network. Any number of extra sections could be added. The usual location of the resistors and the capacitors has been switched in the oscillator in C.

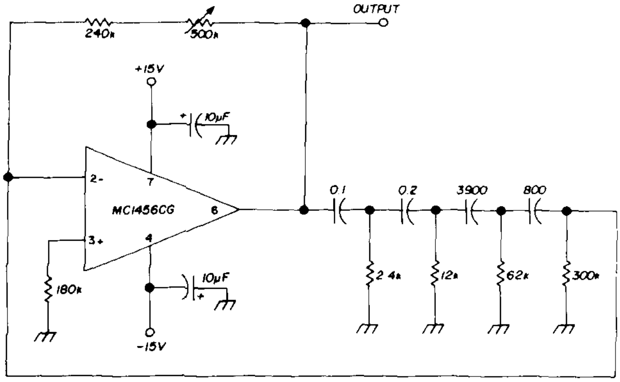

fig. 7. Phase-shift oscillator with series C and shunt R, multiple sections (more than three), and tapering.

could be thought of as a piece of coax cable with a high-resistance center conductor. It is then effectively a multisection R-C network with its shunt capacity to ground continuously in effect along the length of the line. Such a distributed R-C line phase-shift oscillator was actually built and tested as shown in **fig. 8**. The R-C line was made up of a deposited-carbon resistor (40K) of the older non-encapsulated style. A layer of ordinary kitchen-type aluminum foil was wrapped around its body and taped in place. This foil served as the outer conductor of the line, and the paint on the resistor formed the dielectric. The circuit oscillated at about 200 kHz, so it was necessary to use an IC op amp which had a high slew rate; I chose the Signetics NE531T because its slew rate is uniform at about 30 V/μsec regardless of the closed-loop gain. The NE531T is not internally compensated, however, and requires a single external capacitor for this purpose (100 pF between pins 8 and 6).

There are some other interesting ways of modifying the phase-shift oscillator. Combining tapering with a distributed line suggests itself, but is a bit difficult to implement using standard components. However, this is the sort of thing that may be very easy to do on a silicon chip in ICs of the future.

Wien Bridge variations

The Wien-Bridge oscillator has been well worked over; many variations have appeared. Early efforts to make a Wien-Bridge oscillator using bipolar transistors were inevitably tuned by a dual variable resistor, or used switched Rs and Cs. This,

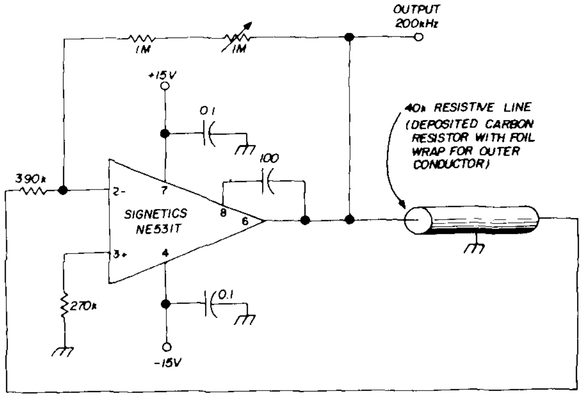

fig. 8. Phase-shift oscillator using distributed R-C line as the phase-shift network.

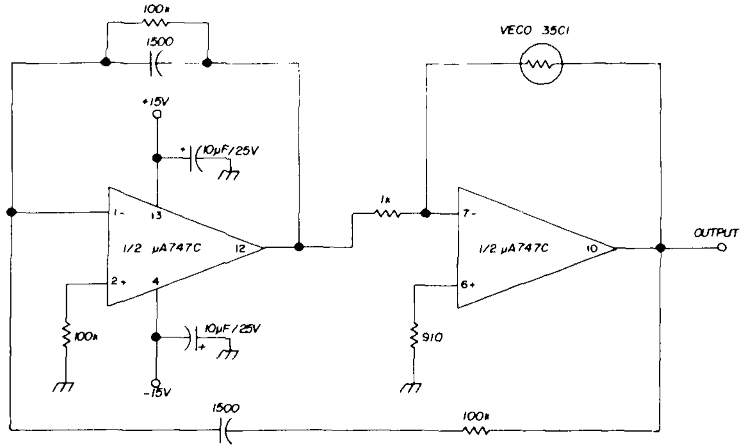

fig. 9. Baxandall version of the Wien Bridge. Notice the use of a thermistor as the nonlinear resistor in the feedback control circuit.

of course, was simply due to the low input impedance of the base of a bipolar transistor. An interesting variation of the Wien Bridge is the Baxandall modified Wien Bridge circuit as shown in **fig. 9**.[6] The main reason for using the Baxandall version of the Wien Bridge is to prevent the op amp input impedance from loading the parallel R-C branch of the bridge. A practical circuit before op amps used four 2N404s.[7]

Fig. 9 also demonstrates the use of a thermistor as the nonlinear resistor in the feedback control. Unlike the lamp bulb, the thermistor *decreases* resistance as it is heated; so it is placed in the position shown to increase negative feedback as it heats. This type of feedback control element is used in at least one laboratory audio oscillator, the General Radio 1311A.[8]

It is also possible to rectify the output of the oscillator and apply the filtered dc to some form of agc element. The agc element can be any one of a number of devices such as a diode, transistor or fet.[9]

An fet can also be used as a voltage-controlled resistance, and so is sometimes used in the negative-feedback voltage divider of the Wien Bridge oscil-

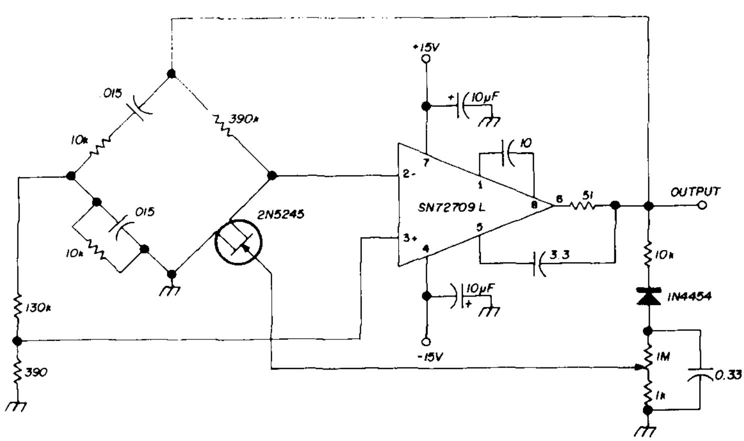

fig. 10. Wien Bridge using an op amp for gain and a fet for feedback control.

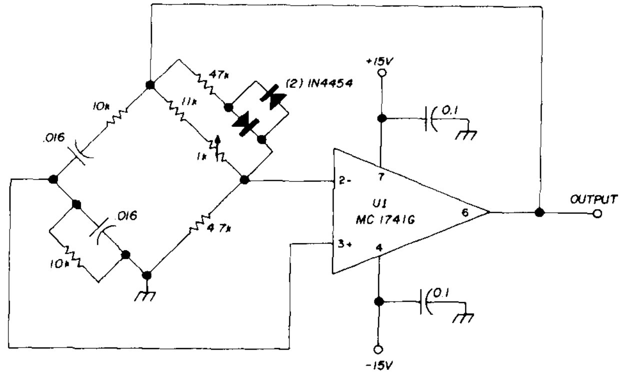

fig. 11. Wien Bridge oscillator using diodes for non-linear control element.

lator.[11,12] The source and drain terminals of the fet are treated as the two terminals of a resistor, and the control voltage is applied between source and gate. The fet used as a voltage-variable resistor should be operated with zero dc from source to drain and with very small ac voltage across it. Special fets characterized as voltage-controlled resistors are Siliconix.[13] An actual example of a Wien Bridge oscillator using an fet for feedback-control and an IC op amp as the gain stage is shown in **fig. 10.**[12]

There is one version of the Wien-Bridge oscillator that has no time constant inherent in the negative feedback control element.[14] This circuit is shown in **fig. 11.** As the nonlinear element is a pair of back-to-back diodes essentially operating instantaneously, there is no time constant to them. In fact, the negative feedback portion is very much like an operational voltage clipper. The 47-kilohm resistor in series with the diode pair and the fact that the diodes go into conduction symmetrically, however, allow this oscillator to produce fairly pure sine-wave output.

Of course, there are many other types of R-C oscillators besides the four basic circuits covered here. The ones which produce non-sinusoidal waveforms, such as astable multi-vibrators, would provide enough subject material for another article. Even in the restricted area of sinusoidal oscillators there are a couple more types that have not been covered here. These types are rarely used, and in some the reasons for their invention (low mu limitations of early tubes) no longer exist.

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ham radio

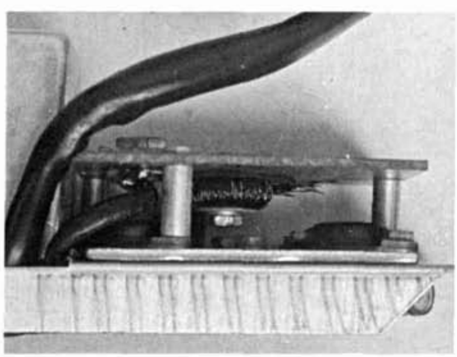
converting the Motorola Dispatcher to 12 v operation

Using an inexpensive homemade toroid to convert these surplus bargains from 6 V to 12 V operation

John Darjany, WB6HXU, 2347 Angela, Suite 4, Pomona, California 91766

There are a lot of Motorola Dispatcher 6- and 2-meter transceivers just waiting to be put into service by amateurs. These partially trasistorized radios started becoming available several years ago as police departments decided to replace their old equipment. Sometimes called motorcycle radios, the units are reliable, small, easy to tune to amateur frequencies and are relatively inexpensive. They are priced as low as $30.00, including accessories, mainly because of their not-so-popular 6 V power supply. Dispatchers with 12 V supplies are available, but not as readily or inexpensively.

The 6 V supplies can be modified for 12 V operation the conventional way, but the expense can easily equal the cost of the entire radio. The 12 V transformer alone is priced in the tens of dollars. For this reason, many hams resort to a series regulator to drop their 12 V system to 6 V for the radio. But this method is terribly inefficient, and the cost is about

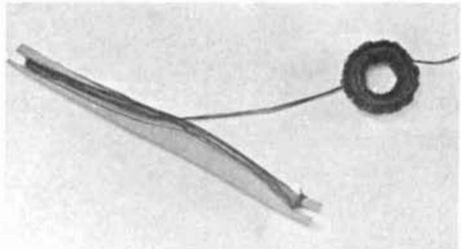


The new toroid is mounted on the phenolic platform when working with a newer model dispatcher.

the same as the method described here.*

theory

The principle behind the conversion is simple. In the original circuit, the transformer primary is a 6-0-6 V winding, **fig. 1A.** The switching transistors alternately connect the 6 V battery between the center tap and one 6 V winding, and the center tap and the other 6 V winding, **figs. 1B** *and* **1C.** The alternating field then induces various voltages into the other windings of the transformer. While Q1 and Q2 are conducting, **fig. 1B,** winding W2 is open and has a voltage induced in it. Likewise, while Q3 and Q4 are conducting, **fig. 1C,** W1 is open and has a voltage induced in it. Since W1 contains the same number of turns as W2, when 6V is switched across either winding, then 6 V is induced into the other. Furthermore, because the windings are in series aiding, the two 6 V potentials add to 12 V between A and C. During one half of the cycle, A is positive 12 V with respect to C, and during the other half, A is negative 12 V with respect to C. So, as it turns out, switching the 6 V battery in this fashion is equivalent to switching a 12 V battery across the entire winding (A-to-C). This is precisely what is accomplished by the modification described here.



The home-wound toroid and the cardboard bobbin.

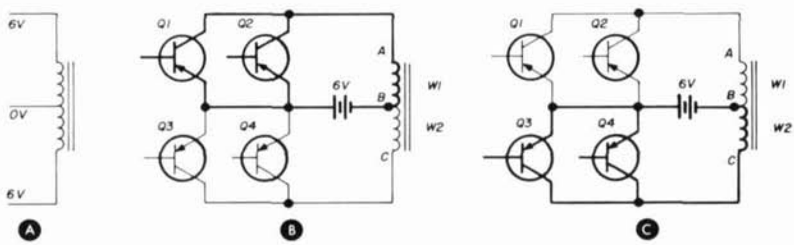


fig. 1. Original circuit for the Dispatcher power supply. B and C show current flow during alternate half cycles.

The original supply consists of four transistors wired in push-pull-parallel. Their configuration is changed to that of a quasi-complimentary bridge for the 12 V supply. During one half of the cycle, Q1 and Q4 conduct, connecting the 12 V battery between A and C with positive on A **(fig. 2A).** During the other half of the cycle, the polarity is reversed by the action of Q1 and Q4 turning off as Q2 and Q3 turn on **(fig. 2B).**

Drive is obtained by the addition of a

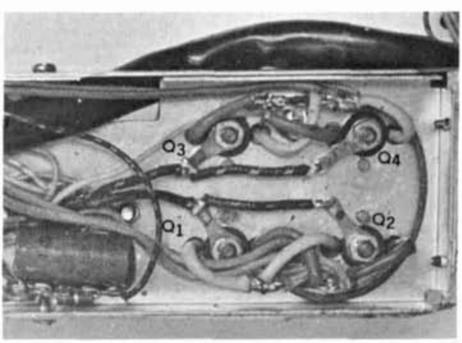
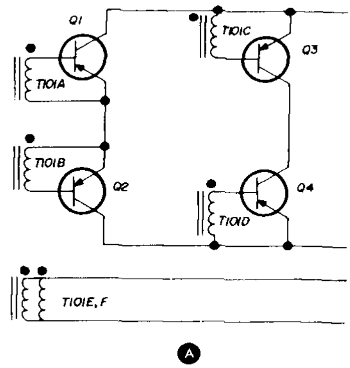


Photo of the power supply shows transistor location and chassis orientation for best parts access.

*A complete parts kit, including a wound toroid, is available for $6.50 postpaid from John Darjany, WB6HXU, 2347 Angela, Suite 4, Pomona, California 91766.



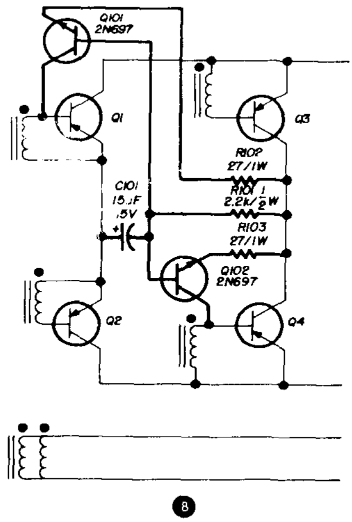


fig. 3. Drive circuit of the modified power supply uses resistance of the toroid windings for base-current limiting and balancing (A). The starting circuit is shown in B.

small home-wound toroidal transformer which utilizes the dc resistance of its windings for base-current limiting and balancing (**fig. 3A**). The start circuit, **fig. 3B**, generates a pulse rather than a continuous level so that receive current drain is held to a minimum (about 500 mA). The original 6 V circuit may be compared with the complete modified 12 V circuit for further clarification (**figs. 4** and **5**).

modification procedure

The first step in the modification is to wind the driver transformer. Cut a bobbin out of stiff cardboard and wind around it about 9 feet of six strands of number 30 enameled copper wire. Then, using the bobbin as a sewing needle, thread 80 turns of the six-strand bunch onto a Magnetics 55206-A2 toroid core, leaving about six inches of wire at each end.* Set the transformer aside until later.

Remove the power supply chassis from the bottom compartment and position it upside down in front of you, oriented as in the photographs. Note that there are several versions of this supply, and some may appear somewhat different than the ones shown. For example, one later version uses TO-3 transistors rather than TO-36 as pictured here.

Unsolder Q2 (see the photo for transistor locations) and remove it from the chassis. Install a mica washer with some thermal joint compound, reinstall and rewire Q2. Next, remove the jumper between the collector of Q1 and Q2. Remove the black and yellow wire from the transformer at the collector of Q3. Remove the emitter wires from Q3 and Q4 at the terminal strip. Connect the emitter of Q3 to the collector of Q1. Connect the emitter of Q4 to the collec-

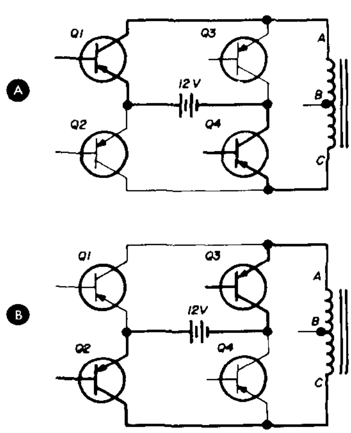


fig. 2. Basic operation of the circuit modified for 12 V operation. Again, A and B show current flow during alternate half cycles.

*Magnetics Component Division, Butler, Pennsylvania 16001.

tor of Q2. Extend the black and yellow wire from the transformer with some number 16 wire, and connect to the collector of Q2. Remove all four base wires from the transistors at the terminal strips and remove the jumpers between adjacent terminals. Remove the yellow and white and the red and white wires coming from the transformer at the terminal strip. Connect the base wires of the four transistors individually to the four empty terminal lugs. Disconnect the black and brown wire from the transformer and tape it so that it will not short to something.

The phenolic mounting platform for the toroid is part of the newer supplies. Avoid short circuits by mounting it so that no sharp or hard edges or surfaces contact the wires. Connect a wire from the collectors of Q3 and Q4 to the negative supply (heavy green wires coming from terminals 2, 3, 4 and 5 on the power connector). If an earlier version supply is being converted, mount the toroid below the chassis, held in place by its leads and the green wires.

Remove the 330 ohm, ½ W resistor. Remove the white and brown wire from the transformer and tape the free end. Remove the positive supply wire from the 25 ohm, 10 W adjustable resistor and connect the resistor to the white and red wire from the transformer. Change the 220 ohm, ½ W resistor to a 220 ohm 1 W resistor.

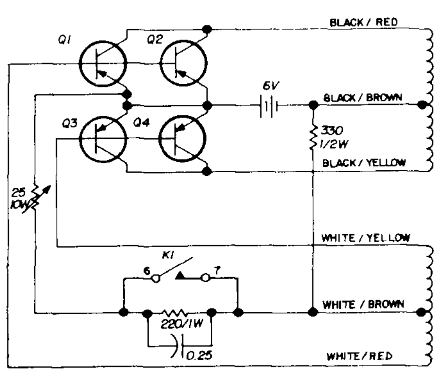


fig. 4. The original power supply circuit set up for 6 V operation.

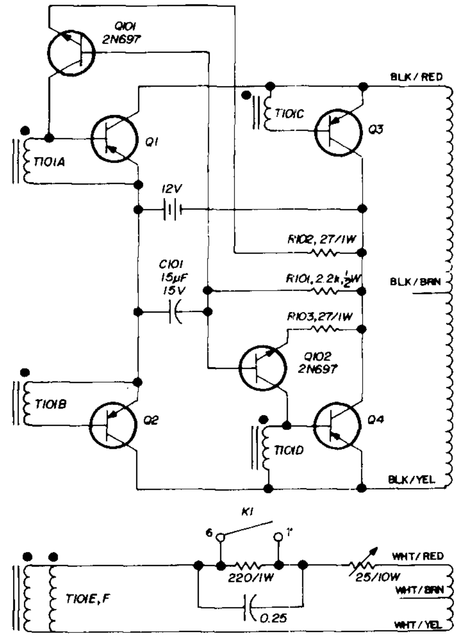


fig. 5. The completely modified power supply set up for 12 V operation. T101 is 80 turns, six strands no. 30 magnet wire on a Magnetics 55206-A2 toroid core. It is described in the text.

toroid wiring

For clarification, the six wires leaving the center of the toroid nearest the chassis will be called positive, and those leaving the center of the toroid away from the chassis will be called negative. This will help in connecting the windings in proper phases.

Connect any two positive toroid wires to the 220 ohm, 1 W resistor at the side not connected to the 25 ohm, 10 W adjustable resistor. Find the two corresponding negative wires with an ohmmeter, and connect them to the white and yellow wire coming from the original transformer. Connect any one remaining positive wire to the base of Q1, and its corresponding negative wire to the emitter of Q1.

Connect another positive wire to the base of Q4 and its corresponding negative wire to the emitter of Q4. Connect a remaining negative wire to the base of Q2 and its corresponding positive wire to the emitter of Q2. Connect the last negative wire to the base of Q3 and the last positive wire to the emitter of Q3. The wiring of the starting circuit is straightforward and can be done by following the diagram of **fig. 3B** and the photos.

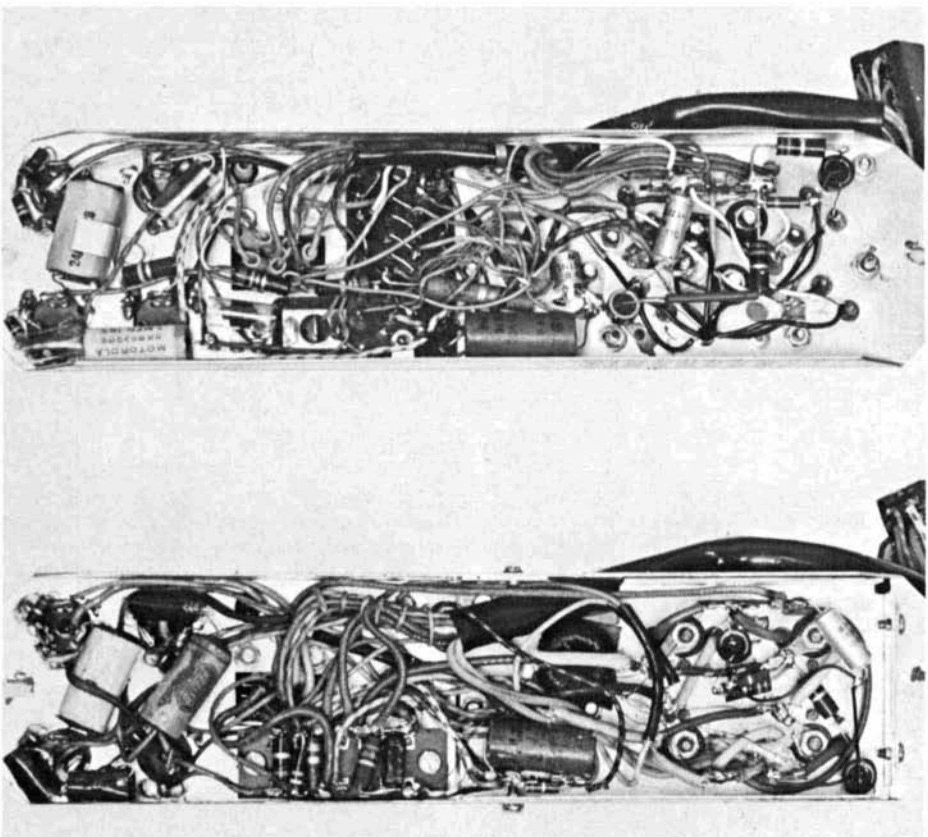
operation

The modification is now complete and should require no adjustments. If, however, the transmit voltages are low, careful adjustment of the 25 ohm, 10 W resistor should help. Of course, it is advisable to disconnect power while making adjustments.

Most of the parts used should be available in the average junk box. The 2N697 transistors can be replaced with any silicon npn switching transistor with similar characteristics. There are surplus toroids available which can replace the one called for – but the new, single-unit price is much less than a dollar, so surplus store hunting may not be worth the effort.

The results have been excellent. One unit has been in 12 V operation for nearly three years, and has not given any trouble at all. Another was completed along with this article, and looks as good as the first. Inquiries accompanied by a return envelope will be answered gladly.

ham radio



Underchassis view of the completed modification. The later model Dispatchers is shown in A, while B shows the earlier model. Notice on the earlier units how the toroid is mounted under the chassis, held in place by the surrounding wires. Tape under the toroid protects the enameled wires from the steel chassis.

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More Details? CHECK—OFF Page 110

optimizing the superregenerative detector

Hangover and receiver radiation can be minimized with these circuit refinements

Charles L. Ring, 294 Boulevard, Pompton Plains, New Jersey 07444

The superregenerative detector, known since the early 1920's, has seen limited commercial application but has appealed to radio amateurs and radio-control enthusiasts. Few circuits are capable of providing such high gain from a minimum of conventional parts, not to mention additional advantages such as low power requirements and good high-frequency performance.

The superregen circuit has been described in previous issues of *ham radio*.[1,2] The problems of hangover (blocking that limits sensitivity) and receiver radiation were shown to be minimized by adding a diode across part of the tank inductance.[3]

While many subtleties are involved in circuit operation, I've found that two simple factors are outstanding in achieving optimum effeciency: time constant and applied bias, which influence the blocking-voltage waveform.

The superregenerative circuit is a blocking oscillator[4] operating at radio frequency. The circuit contains an RC network with a time constant long with respect to the natural frequency of the circuit. After a few cycles of oscillation, a reverse bias develops on the RC network. The reverse bias becomes sufficiently large to cause circuit losses to exceed circuit gain. Residual oscillation then decays, and the remaining dc bias bleeds off until the conduction point of the circuit active device is reached; then the process repeats. The process is illustrated in **fig. 1**, a typical blocking- oscillator waveform developed at the input.

The rate at which the circuit goes into and out of oscillation is the quench frequency, which is made to occur at a supersonic rate so as not to interfere with the received signal. Detection occurs because any rf signal (or other electrical disturbance) will cause the bias slope to encounter the conduction point ahead of its natural time, thus increasing the quench frequency. Since a higher quench frequency means a higher average current through the circuit, a replica of the signal modulation is available.

The key to successful operation of the circuit is in the quenching voltage waveform, which influences circuit performance in two major ways. It determines:

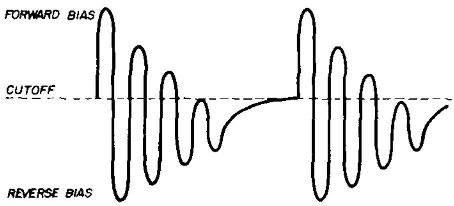


fig. 1. Typical blocking-oscillator wave-form developed at input of superregenerative circuit.

a. The amount of audio that will be developed from a given input signal.

b. The degree of regeneration a given signal will experience prior to the actual circuit oscillation burst.

It is this latter factor which controls circuit selectivity and sensitivity.

Consider now the effect of the quenching voltage waveform on circuit audio response, as illustrated in **fig. 2.** (In this figure the initial oscillation burst has been eliminated since it doesn't directly pertain to the following discussion.) In **fig. 2A** a relatively long time constant is used with a large value of forward bias. The forward bias increases the quench *frequency* by eliminating the shallow portion of the decay slope. The remaining portion of the slope crosses the circuit firing point (approximately equal to cutoff in the case of transistors) at a steep angle. If a signal having an instantaneous amplitude represented by the line A-B is applied to the slope, the circuit fires at T2 rather than at T1. The increase in *quench frequency* is small, resulting in a relatively low audio output.

The circuit conditions for **fig. 2B** are the same except that forward bias has been greatly reduced, preserving the shallow portion of the decay slope. If the same signal represented by line A-B is applied to this slope, a much greater change in quench frequency is produced, resulting in a much higher audio output. Unfortunately, the reduction in forward bias also results in a normal quench frequency well within the audio range, which destroys the usefulness of the circuit. What is important here is that (a) slope *shape* rather than quench frequency *per se* influences circuit efficiency, and (b) a shallow slope depends on the application of minimum forward bias.

Having established one condition for optimum performance, the second problem is to eliminate the low quench frequency. This condition can be achieved by reducing the value of the time-constant network: if the time-constant is made as short as possible, the net effect will be a usable, supersonic quench frequency with a shallow slope.

selectivity and sensitivity

The most significant advantage of the

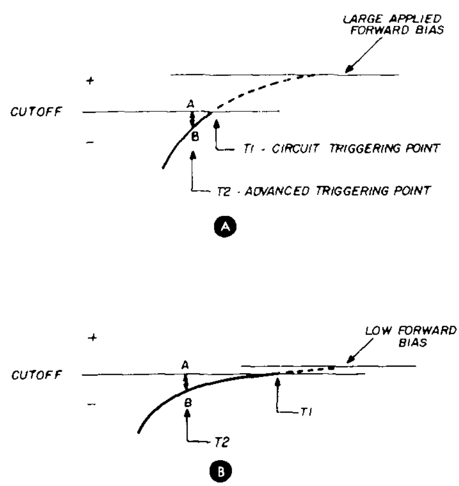


fig. 2. Triggering-point slope and quench frequency. In A a signal of amplitude A-B advances triggering time from T1 to T2. Increase in quench frequency is small since shallow portion of blocking waveform is lost due to large forward bias. Reduced forward bias, B, preserves shallow portion of slope and signal A-B causes a large change in quench frequency.

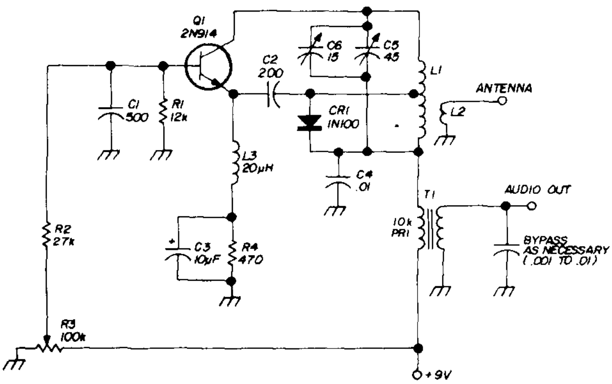
low forward bias, short time-constant combination is not the production of high amplitude audio but its influence with regard to circuit selectivity and sensitivity, which is related to the amount of pure regeneration present in the circuit. It is well known that simple regenerative receivers achieve a Q multiplying effect by operating close to the unity gain or oscillation point.

The superregenerative circuit is also regenerative when the decay slope is near *the unity-gain bias level.* Here, the advantage of a shallow slope, as compared to a steep slope, is that the shallow slope permits the signal to dwell in the threshold region for a longer period. This action allows a greater regenerative buildup with a resultant increase in gain and decrease in bandwidth. Since circuit losses are compensated by regeneration, the Q of the tank circuit doesn't greatly influence selectivity as it is sometimes believed. In fact, an excessively high Q tank is not permissible in the circuit.

hangover

In the superregenerative circuit, it is essential that no oscillations remain in the tank circuit after oscillation has terminated. Such persistent oscillation, sometimes referred to as hangover, can create spurious responses that block detection by forcing the receiver to listen to its own residual rf signal.

Hangover, which results from the large reactances involved, prevents efficient operation in the lower frequency ranges. In this case efficient operation can be obtained only at excessively low quench frequencies, which allow time for natural tank damping. Furthermore, attenpts at constructing a truly narrowband circuit, such as using a quartz crystal in conjunction with the tank, are doomed to failure unless some method of damping is developed.



L1 14 turns no. 24 on 5/16" diameter form, tap 5 turns from ground end

L2 4 turns no. 24 around cold end of L1

fig. 3. 28 MHz superregenerator.

The performance of conventional superregen circuits can be improved by adding a damping diode across a portion of the tank inductance. The diode is connected so that it doesn't prevent positive feedback. The diode dissipates energy immediately after the circuit oscillation burst, thus hangover is eliminated. Since the barrier potential (approximately 0.2 volt in the case of germanium diodes) exceeds any normal signal input, no tank loading will occur. The advantages resulting from diode inclusion are:

1. Residual oscillations are eliminated from the sensitive portion of the decay slope.

2. A more highly regenerative (shallow) slope can be employed since the tendency of the circuit to lapse into cw oscillation will be reduced.

3. Receiver radiation is greatly reduced: Diode damping lowers the amplitude and shortens the duration of the radiated pulse. (Preliminary measurements indicated a 12:1 reduction of radiated noise in the case of the 28-MHz circuit described below.)

practical circuits

Two circuits are presented, which exemplify the superregen operational principles. The first **(fig. 3)** covers the 28-MHz band and tunes from approximately 21 to 40 MHz. The second circuit, **fig. 4,** covers the 144-MHz band.

occurs, the control should be advanced to a slightly less sensitive setting.

A simple resonant circuit and diode detector were placed near the antenna of the 28-MHz receiver to detect receiver radiation. The output was fed to a scope and displayed as a vertical line. When the diode was inserted, the radiation-pulse

L1	3 turns no. 19 on 5/16" diameter form, tap at 1 turn from ground end
L2	2 turns no. 19 around cold end of L1

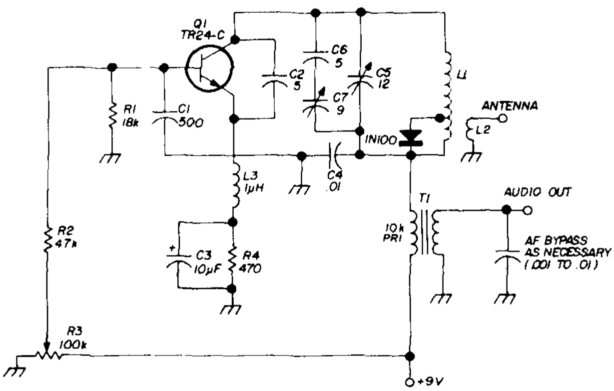


fig. 4. 144-MHz super-regenerator.

The difference between these circuits in the use of a Colpitts oscillator in the 144-MHz range to compensate for shunt capacitance.

In both circuits, the time-constant network consists of R1, C1. As a general rule, the time constant of R1, C1 should be as short as possible, the limiting factor being the point at which the circuit tends to lapse into cw oscillation. In this regard, one variable requiring compensation will be the beta spread within given transistor types, with high-gain units requiring a longer time constant. Here, the best method of adjustment consisted of holding C1 at 500 pF and adjusting the value of R1. When making this change isolating resistor R2 should have a value at least twice that of R1. Potentiometer R3 controls the forward bias and acts as a quench-frequency control. The most sensitive setting will correspond to the minimum bias consistent with a usable quench frequency – usually in the order of 20 kHz. Occasionally, this sensitive setting of R3 will result in oscillator pulling from strong local signals. If this

amplitude fell to 20 percent of its former value. The pulse width (scope sweep on) also decreased to about 70 percent of its former value.

The net effect on a nearby linearized receiver (no avc; no rf overload) was to reduce the audio noise voltage measured at the receiver output to 1/12th of its former value. This reduction amounts to about 21 dB. The 144-MHz circuit demonstrated the same effect; thus the principles appear to apply to every superregenerative circuit, allowing operation beyond the point at which a conventional circuit would be limited.

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ham radio

cooled preamplifier for vhf-uhf reception

An experiment into lowering solid-state device noise figure by cooling the unit hundreds of degrees

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One of the ingredients necessary for weak-signal reception at uhf is a receiver with the lowest possible noise figure. A lower noise rf amplifier in your frontend can make weak, fading stations easily readable as well as bring new signals up, out of the noise.

As a result of improved solid-state devices, noise figures have dropped substantially in the last few years. Today, a common-gate fet amplifier will readily provide a noise figure of 1.5 dB or better at 144 MHz. With this kind of performance from such a simple circuit, any improvement would be more or less academic. At 432 MHz, where low noise figures really begin to pay off, you can use a transistor with a possible noise figure of 1 dB. Fets are slightly inferior at this frequency but can give a 2 dB noise figure. Performance at 1296 MHz is similar, with state-of-the-art noise figures being approximately 2 dB for transistors and 3 dB for fets. In spite of all this, the parametric amplifier is still king. At 432 MHz, 0.5 dB can be achieved, and 1 dB is possible up to 2.3 GHz. However, the paramp represents a rather formidable project; at least it does to all of us who have never built one.

theory

In order to see how much these low noise figures can help your receiving setup, we'll review briefly a few expressions. These are:

$$(S/N)_o \propto 1/T_n \qquad (1)$$

$$T_n = T_s + T_e \qquad (2)$$

Where T_s = source noise temperature
T_e = effective receiver noise temperature
T_n = system noise temperature
$(S/N)_o$ = output signal to noise ratio

The equations tell us that the output signal-to-noise ratio is inversely proportional to the system noise temperature. Further, noise figure is related to receiver noise temperature by

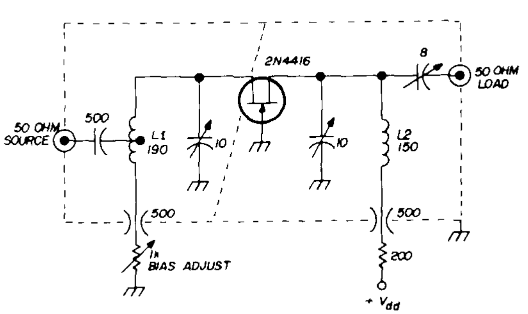
$$T_e = (F - 1)290°K \qquad (3)$$

Noise performance, therefore, is specified by either. It should be pointed out that F in **equation 3** is the ratio noise figure, not in dB. These are related by **equation 4.**

$$F_{db} = 10 \log F \qquad (4)$$

Now, T_S depends on several factors, but if everything is right, can be as low as 200°K at 144 MHz and 20°K at 432 MHz for a moonbounce link. If you're not familiar with this noise figure business, try reading K6MIO's article, and then plug in a few numbers and see what happens.[1] It shouldn't take long to convince yourself of the value of a low noise figure.

The approach to a low-noise amplifier used here is based on the fact that most noise added to the amplified input by an fet is thermal noise and proportional to



L1 6 turns, no. 16 enamelled copper wire, 3/8-inch inner diameter, input tapped 1¾ turns from ground

L2 5 turns, no. 16 enamelled copper wire, 3/8-inch inner diameter

fig. 1. 144 MHz common-gate fet test circuit. The inductances are in nanohenries.

the physical temperature of the device. Thus we should find that **equation 5** is true.

$$T_e \propto T_{device} \qquad (5)$$

test procedure

To test this hypothesis, I built the circuit of **fig. 1**. You may wonder why the common-gate configuration was used instead of the common-source. This is simply because a common-gate amplifier will give just as low a noise figure as a common-source circuit. This is so even though the *device* noise figure is lower in common-source. The input tuned circuit has a higher loaded Q and hence more loss than the corresponding circuit in the common-gate amplifier. Also remember that as we get rid of device noise, the lossy components preceeding the fet become important. Therefore, it is desirable to use the common-gate configuration with its low-Q input.

Before cooling the amplifier, I adjusted the circuit for best noise figure at room temperature. I adjusted the supply voltage, drain current, input coil tap, output coupling capacitor and input and output tuning capacitors. I found that the output coupling capacitor had little effect on noise figure. I finally set it at about 2 pF. The position of the input-coil tap was not critical, but a minimum was obtained with it about 30% up from ground.

Input tuning was broad. Again, however, there was no question that a dip in noise figure did exist. With the output tuning I obtained best noise figure at approximately the position of the capacitor that gave maximum gain. This point was, however, with the circuit tuned slightly on the high side of the signal frequency.

The results of varying the supply voltage and drain current differ somewhat from what is usually recommended as optimum. Most amplifiers use a supply voltage of 10 to 15 Vdc and a drain current of approximately 5 mA. I found with this amplifier that the best noise figure occured at $V_{gs} = 0$ and $V_{ds} = 5$ Vdc.

To clarify this somewhat, if $V_{gs} = 0$, a

table 1. Tabulated results of cooling a 144 MHz common-gate, fet preamplifier.

device temperature T_{device} (°K)	noise figure F (dB)	drain current I_{ds} (mA)	supply voltage V_{dd} (Vdc)
300	1.5	7.9	6
200	1.2	9.8	8
77	0.8	16.	12

rather high drain current results, which requires that the supply voltage be low so that power dissipation and hence noise figure stay down. Turning things around, if we start with a high supply voltage, then drain current must be low for the same reason. Even with I_{ds} = 7.9 mA (V_{gs} = 0) and V_{ds} = 5 Vdc, you would see an increase in noise figure as the device rose to operating temperature.

All the adjustments mentioned lowered the initial noise figure from 2.0 dB to 1.5 dB, and although there was no guarantee that the amplifier would remain optimized upon cooling, this seemed the best way to start.

results

The results of cooling the preamp are tabulated in **table 1** and the data plotted in **fig. 2.** The 200°K temperature was achieved by holding a small cube of dry ice directly on the case of the 2N4416 with a pair of small tweezers. Some slight retuning was necessary to minimize the noise figure. Liquid nitrogen at 77°K was used to get the lowest point in **fig. 2.** The amplifier was carefully lowered about a half inch into a Dewar containing the liquid nitrogen.

The three noise-figure readings are plotted in **fig. 2** and may be connected by a straight line. This shows that the noise temperature of the amplifier is proportional to the physical temperature of the fet. It is seen that at T_{device} = 0, T_e is greater than zero. This is due to circuit and transmission-line loss preceeding the fet and noise added by receiver stages following the cooled preamp. In this case the preamp was followed by two common-gate amplifiers and a transistor mixer into the noise-figure meter.

Because of the relatively high antenna noise temperature at 144 MHz, the low receiver noise temperature at 144 MHz, the low receiver noise temperature obtained cannot be fully appreciated in actual use. However, at 432 MHz this is not so, and if similar device behavior at this frequency is assumed, then a noise figure of 1.2 dB should be possible upon cooling with liquid nitrogen. This number is obtained by drawing a graph similar to **fig. 2** and taking the noise figure at 300°K to be 2.5 dB. While this is good, a parametric amplifier and *some* transistor amplifiers can do better.

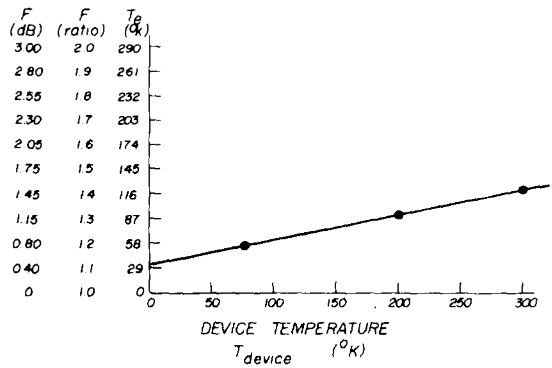


fig. 2. Noise figure and noise temperature are plotted against device temperature for the 144-MHz preamplifier shown in fig. 1. The results are tabulated in table 1.

The possibility of improvement with further experimentation should be evident. Barring some physical change in the operation of the device, cooling with liquid helium at about 4°K would give an extremely low noise temperature. Also, the possibility of operation above 1 GHz should not be overlooked since low-noise *microwave fets may be just around the* corner. I'd be very interested in hearing results from anyone who might try something along these lines.

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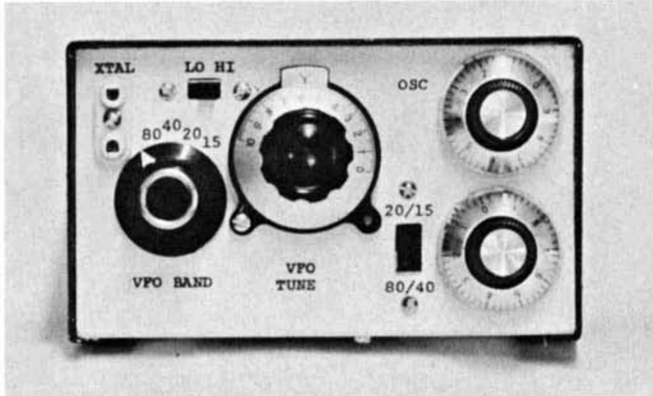


a multiband fet vfo QRPP transmitter

For the real low-power newcomer this rig covers the 80- through 15-meter bands with an inexpensive rf module and a versatile vfo

Ade Weiss, K8EEG, 213 Forest Street, Vermillion, South Dakota

Operating with less than five watts of power is lots of fun, but the newcomer to the sport of QRPP often faces many unexpected problems. This is particularly true when the newcomer chooses to build his own solid-state gear.

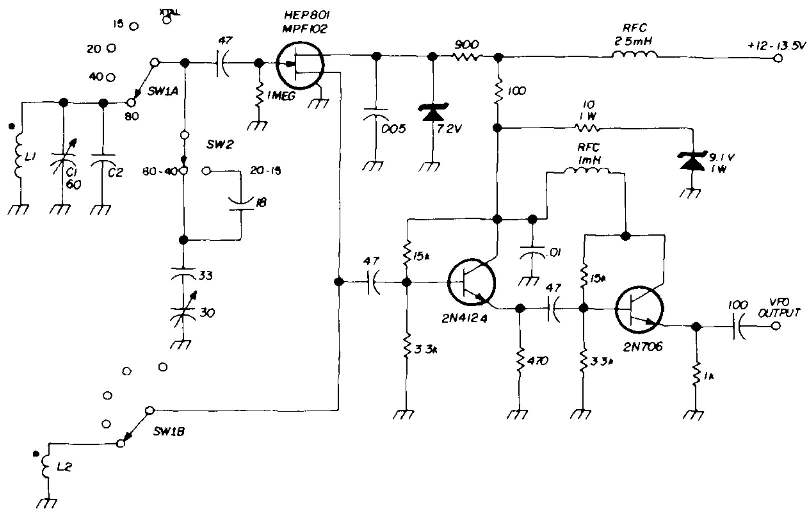
First, most of the published designs for solid-state gear are exclusively for crystal control. This is a serious drawback in actual QRPP operation where the weaker QRPP signal is often smothered by the higher-powered interference. The ability to change frequency to dodge interference is essential to successful operation. Also, the weak QRPP signal is less likely to work stations by calling CQ than by answering another station's CQ on the other station's frequency.

Second, most solid-state gear is designed for the 80- or 40-meter bands or vhf, totally overlooking the bands which provide the greatest opportunity for consistent success – 20 through 10 meters. This is unfortunate because the propagation and lower atmospheric loss factors of these bands foster low-power operation. For example, during the past year I have worked about 40 states including Hawaii and the Commonwealth of Puerto Rico on 20 and 15 meters while using only a simple antenna and an output never exceeding 120 milliwatts. Any QRPPer seriously interested in DX must be able to work these bands.

Third, the QRPP neophyte building his first solid-state rig often is baffled by the

factory wired and tested, has band-switching capability for 80 through 15 meters, is able to put out between 0.6 and 1.4 watts and has performed well in thousands of installations.

The multiband vfo (**fig. 1**) uses the HEP-801, available at most radio supply



C1	60 pF trimmer (Elmenco 404)	S3	double pole double throw switch or preferably two single pole double throw switches
S1	2 pole 5 position rotary switch	S4	double pole double throw switch
S2	single pole double throw switch		

fig. 1. Schematic of the multiband fet vfo. All capacitors given in picofarads are silver-mica dipped; all bypass capacitors are ceramic.

problems of duplicating rf circuits. This is certainly true of any attempt to construct a multiband driver-final circuit.

The solid-state transmitter described here provides the interested amateur with an economical, practical and dependable rig. Frequency flexibility is achieved with a bandswitching fet vfo that is stable and chirp free. I took advantage of a readily available Ten-Tec TX-1 driver/final module* to make it easy for the beginner to construct a solid-state transmitter without encountering the difficulties associated with rf power-stage construction. In addition, the TX-1 is

*Available from Ten-Tec, Inc., Highway 411 East, Sevierville, Tennessee 37862 for $7.95 postpaid.

stores; an MPF-102, or its recent derivatives – the 2N5668, 2N5669 or 2N5670 – will also work. I achieved isolation and buffering through two emitter-follower stages and using isolating

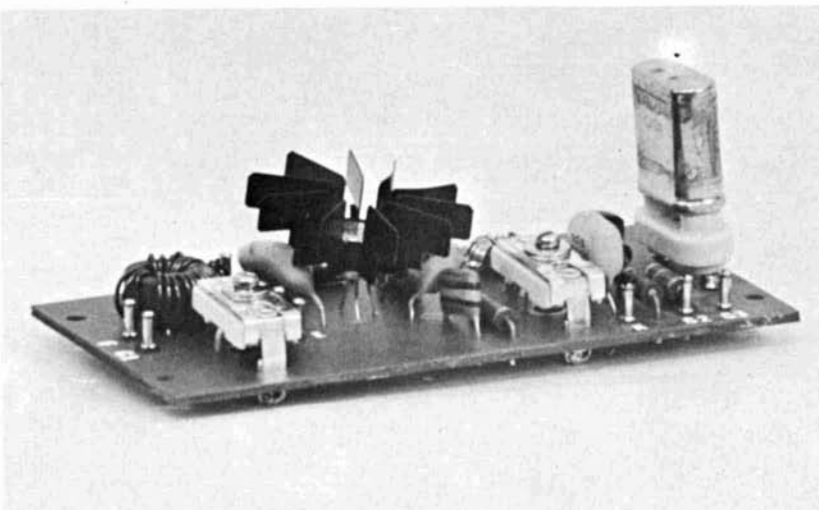
table 1. Tank circuit values for the fet vfo.

band	L1	L2	C2	coverage
80	50 t. no. 28	17 t. no. 24	47 pF	225 kHz
40	19 t. no. 28	9 t. no. 24	180 pF	110 kHz
20	8 t. no. 24	4 t. no. 24	150 pF	120 kHz
15	7 t. no. 24	3.5 t. no. 24	100 pF	340 kHz

All coils are wound on Amidon T-50-2 cores except for the 15 meter coil which is wound on a T-38-2 core. In all cases, L2 is wound over L1 over the full diameter of the core. Both windings start at the dots shown in **fig. 1** and are wound in the same direction! All wire is enameled copper.

rf chokes in the B+ leads to the oscillator and final buffer. Zener-diode regulation of the oscillator voltage insures frequency stability regardless of wide excursions in supply voltage. Although operation of the vfo and final on the same frequency frequently causes chirp, the buffering and

The circuit board is designed for standard size components. The trimmer capacitors are subminiature Elmenco 404 types (since some stores carry only 30- or 40-pF size of this type, these may be substituted if the size of C2 is increased accordingly); the coils are wound



The TX-1 board before any modification or installation.

isolation features of this design result in a chirp-free vfo. In fact, you can key the entire transmitter and vfo by breaking the B+ leads without any chirp at all! It is wiser, however, to allow the vfo to run continuously while keying the driver and final stages. Since the vfo emits a healthy signal, you will want to turn off the vfo while receiving by flipping the vfo bandswitch to the crystal position.

I change bands with a two-pole five-position rotary switch. I use four positions, while the fifth shuts off the oscillator for crystal operation or stand-by. The circuit board has provision for five bands if you wish to use it. You can get bandspread of about 125 kHz by placing a small 33-pF mica capacitor in series with the 30-pF APC vfo tuning capacitor; on 20 and 15 meters, an 18-pF mica is switched into the series circuit for more useful bandspread.

on Amidon toroid cores making for extremely compact high-Q tank circuits.

It can be an interesting task to design and make your own circuit board, as I have done. Besides actually etching a board there are many other construction techniques you can use like no-etch stick-on copper foil patterns, using copper-clad board with insulated stand-offs or insulated islands and the simple technique of using plain epoxy board, drilling holes for component leads and wiring the circuit with short insulated jumpers and the protruding component leads.[1,2,3]

Regardless of the construction method chosen, use care in soldering component leads and use a heat sink when soldering the semiconductors. A no. 60 drill*—the

*Available for 50c from America's Hobby Center, 146 West 22nd Street, New York, New York 10011.

right size for drilling holes for component leads—is invaluable.

To tune and calibrate the vfo, work on one band at a time. After installing the proper L/C combination, temporarily connect the tank to the proper oscillator ports with *short* wires. Find the vfo signal

At that time, cement the toroids to the circuit board. Finally, make an aluminum shield for the vfo and move on to the transmitter module.

transmitter module

The two-stage TX-1 (**fig. 2**) uses high f_T high-gain devices (MPS6514, 2N4427,

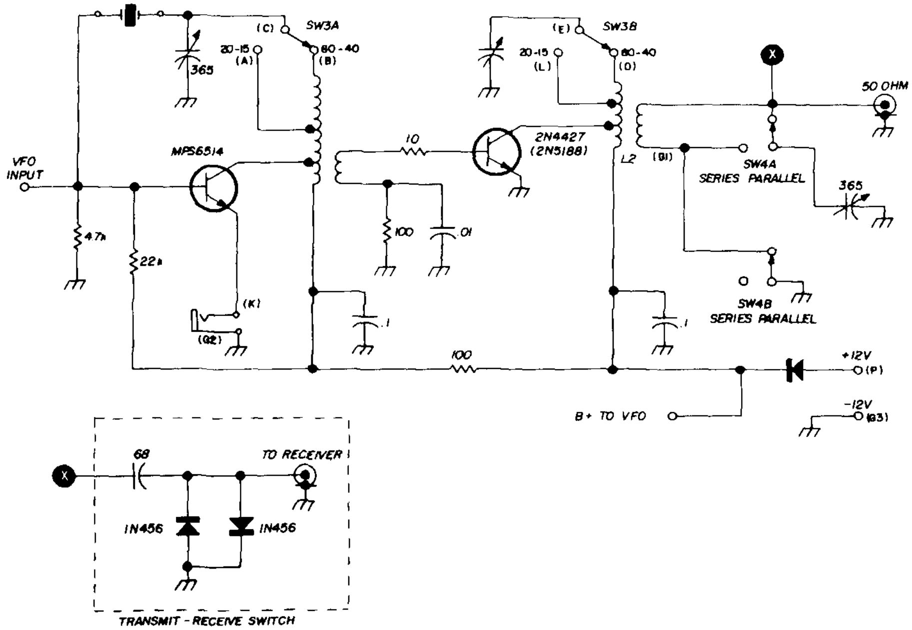


fig. 2. Schematic of the Ten-Tec TX-1 transmitter module. The letters given in parenthesis correspond to the proper port on the TX-1 circuit board.

by listening on the station receiver while adjusting C1. The best way to tune a toroid is to use a variable capacitor, but an adequate compromise tuning method would be to squeeze the turns closer together or spread them further apart. Remember though, this type of adjustment will upset the symmetry of the toroid and can lead to some slight losses and stray radiation. However, some coil tuning will probably be necessary on 20 and 15 meters.

Once you initially calibrate all bands, mount the board, connect the B+, the driver base and the tuning capacitor. Calibrate the vfo for final bandspread *only* after the transmitter is completed.

which can be replaced by the cheaper 2N5188 at 60c), with their collectors tapped down on toroids for proper impedance matching. Preparation of the TX-1 requires two minor modifications. First, remove the trimmer capacitors used in the two tank circuits. To do this, insert a knife blade between the board and one end of the capacitor, melt the solder, and twist the knife blade until the capacitor lead is clear of the board. Second, if you wish crystal control capability, remove the crystal socket from the board and mount it on a panel. A third possible modification is the addition of an antenna tuning capacitor across the output link (see schematic). Simply break the common (ground) copper strip at the bottom of the circuit board before it reaches the ground side port (G1) of the antenna link. Attach the antenna link directly to the dpdt switch, thus allowing you to *shift between series and parallel* tuning to accomodate any inductive or capacitive reactances your feedline may present. However, this luxury is not essential and may be omitted without limiting the rig's effectiveness.

To mount the TX-1, simply connect the key leads, B+, vfo output, crystal socket and bandswitch. A word of caution about the bandswitch. Keep the bandswitch leads as short as possible. The capacitance added by four inches of leads, for example, will make it impossible to peak the driver on 15 meters. I recommend mounting the TX-1 flush against the panel where the switch is to be mounted. Two spdt switches instead of a single dpdt will insure the shortest possible leads.

tuning

If you haven't worked with a solid-state transmitter before, a few words of advice are in order. First and foremost, *do not attempt to operate the transmitter without a proper load*! To do so is the easiest way to zap transistors. Further, if your swr is over 3:1, do not hold the key down for long periods – three seconds on is a good rule of thumb. Mismatch can cause thermal runaway – the puncturing of internal transistor junctions because of the heat resulting from the excessive current. Normally the 2N4427 should be comfortably warm. Also watch for self-oscillation, shown by the transmitter continuing to put out rf after the key has been opened. If you encounter this problem, immediately remove the B+. It is usually caused by improper final tuning. Transistors can be tricky devices and they have suicidal tendencies; self-*oscillation inevitably leads to self-destruction* in the transistor world.

Become familiar with the tuning of the TX-1 before making modifications. Install a crystal and jumper wires for the proper band, and tune the driver trimmer until

table 2. Measured outputs from finished rig: (R = 49 ohms)

band	voltage	watts
80	13	1.87
40	12	1.4
20	10	1.02
15	7.5	0.56

Measured using the circuit of figure 3; R = 49 ohms. P = V2/2R

the crystal breaks into oscillation. Peak tuning is accompanied by pulling of the crystal frequency. With a 50-ohm non-inductive load and rf indicator **(fig. 3)** connected across the output posts, close the key and tune for an output indication. Retuning the driver and final will be necessary for peak output. Remember the 2N4427 is not a 4-1000; it should be

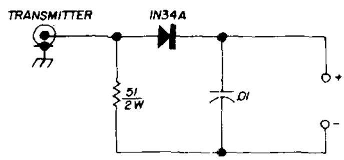


fig. 3. Simple rf output indicator for use with almost any QRP transmitter. Voltmeter should read at least 15 volts full scale for use with this rig.

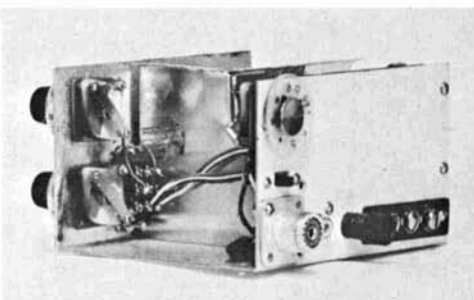
respected and not pushed unnecessarily. Tuning with the vfo is about the same as with a crystal. The final tuning with the vfo will result in pulling the vfo frequency several hundred hertz when the key is closed. This is normal, and peak output should occur with a pulling of about 500 hertz.

construction

The photos show my approach to construction. I housed the transmitter in a 3 x 4 x 5-inch box, making for some tight fits. The only critical aspect of mounting is the drive/final bandswitch. I used cheap mica-dielectric variable capacitors, and they work well. The switch used to insert series capacitance to the vfo tuning capacitor should be a good toggle switch – slide switches usually exhibit intermittent contact, a detriment to vfo stability. A small vernier and 30-pF APC capacitor provide smooth tuning on all bands. Since all critical parts are mounted on the circuit boards, you can work out your own mounting setup.

operation

It is hard to define what is so fascinating about QRPP. I suspect it is accepting the challenges of QRM and propagation, and with a dry cell and a transmitter that fits into a lunchbox, working the world. In a sense, it is getting back to the basics that characterized most of our Novice experiences. The story told by converted appliance operators is always the same: the KWS-1 gathers dust while the QRPP rig is worked to death providing all the excitement that ham radio can offer. And the thrill never seems to leave – you can work a KH6 on 40 meters three nights in a row (as I did this spring) with a couple hundred milliwatts, and you still tremble as he comes back to your call, you still feel the exhilaration of man stripped to the bare essentials confronting and overcoming nature. It's a great experience!



Rear view of the completed transmitter shows use of inexpensive bc type capacitors and the shield around the vfo compartment.

Let me try to give you a realistic idea of what to expect from QRPP operation and how to go about it. QRPP operation requires skill, patience and an understanding of how everything in the transmitting system works together. Propagation is important – what to expect of your rig and antenna in terms of distance and signal strength during a given season at a particular time of day on a particular band. Likewise, the QRPP operator attempts to be as efficient as possible in his overall system – good matching, efficient feedline, accurately cut antenna, clean signal. In short, he attempts to offset his power disadvantage by knowledge, skill, and above all, patience.

Every newcomer to QRPP is amazed at what can be done with a few hundred milliwatts; for example, newcomer WA8WWS, during the first ten days of operation at 2.5-watts input, managed to work 54 stations in 18 states on 40 meters. Some guys turn into fanatics upon seeing what QRPP can do – WA8DDI was so excited that after six months he has WAS and over 50 countries with his 1-watt output rig! But for most of us, QRPP is just the way we approach ham radio for daily enjoyment of the hobby – ragchews, casual contacts and the like. Here are some brief suggestions to help you master QRPP operating skills.

You can assume that most stations will be running high power, and if a station calling CQ is weak, it is very likely that propagation between his location and yours is bad. Instead of calling him, find a stronger signal that you can copy well. Experience has shown that calling CQ with QRPP is futile – although, it doesn't hurt to try as long as you don't expect a logjam of answers! Third, select your transmitting frequency carefully. Many hams use transceivers with ±1-kHz offset tuning, so this is normally the limit of deviation for a calling station. But even in a crowded situation, this is usually enough to put your signal into the open. If interference appears after making contact, suggest changing frequency, but let him call you so you can zero beat him. Cultivate a good clean fist and send at medium speed until asked to speed up – it is easier to copy a slow signal than a fast one. Don't be afraid to repeat things two or three times if the going is rough – the other guy will appreciate it. Inform the other station that you are QRPP immediately upon making contact – this inevitably causes him to turn up the rf gain and notch the filter a bit deeper. Most amateurs find it exciting to work a milliwatt station; take advantage of it. Similarly, always afix QRP to your beginning and concluding signature, at the end of a contact, give a short QRZ – it's amazing how many fellows will read the mail when they hear someone sign QRP, and will gladly come back to a QRZ to give you a report. Don't waste all night calling the same station – if he hasn't come back on the third call, he probably has the rf gain turned back waiting for a blockbuster to shake his shack. Don't be shy – usually a station calling CQ DX will gladly accommodate a milliwatt station, and the psychology is perfect – he's tuning for weak signals in the first place, and that's you! Above all, be *patient* and know what to expect from your gear and propagation phenomenon. You will find that after getting the hang of it, it will be possible to do almost as much with QRPP as with high power. In populous areas of the country, QRPP stations have as high as an 85% call-to-contact ratio. The main point is the QRPP operator voluntarily imposes upon himself a power limitation, but after getting the hang of operating QRPP, wonders why he ever thought he needed the kW amplifier in the first place.

Whether you are a newcomer or oldtimer, this is a good rig for the investment of time and money, and it has performed flawlessly for me. Give yourself a break – try QRPP.

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ham radio

using Y parameters in rf amplifier design

By using Y parameters in rf circuit design, the designer can determine stability and gain before building a breadboard

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On the data sheet for the RCA 40673, a dual-gate mosfet, is the statement, "The reduced capacitance allows operation at maximum gain *without neutralization.*" This is a comforting thought when looking for a suitable rf amplifier. But then you stumble onto a circuit in the 1970 ARRL *Radio Amateur's Handbook* for a dual-gate mosfet rf amplifier, and it is pointed out that neutralization is usually required.

The truth is, both assertions may be correct. The secret of the stability or instability of a device is locked up in its *y parameters;* the purpose of this article is to give some explanation of them and practical instruction in their use. Let's define some basic notions.

The quantities resistance, r, reactance, x, and impedance, z, should be generally familiar. If not, there is a good primer in the ARRL *Handbook.* The reciprocals of these quantities are very useful and have the following nomenclature and symbology:

$$\text{Conductance} = g = 1/r \quad (1)$$

$$\text{Susceptance} = b = 1/x \quad (2)$$

$$\text{Admittance} = y = 1/z \quad (3)$$

Note that these quantities are measured in mhos, derived from ohm spelled backwards!

Understanding the relationships between these six quantities is not too difficult, however. The standard way of handling them mathematically is to use the complex plane and complex algebra. It's not too much fun on a slide rule, but quite necessary to get the answers. Since the y parameters are all complex, and the stability calculations involving them must be carried out using complex arithmetic, I have added an appendix on complex numbers with an example or two. Don't let the word "complex" scare you. High-school math books should have additional information on them.

y parameters

In this discussion, I will not deal with all the whys and wherefores that design engineers use in the rf amplifier problem. Suffice it to say, that although impedance could be directly used in design, and although other parameters (like the s and h parameters) could be used, the y parameters make the work the easiest, and are commonly found on device spec sheets. Let's begin by defining them.

Assume for discussion that the transistor is a "black box" sketched in **fig. 1** to which is connected an rf signal source and an output load. In a real circuit you must apply power and bias, etc., but these are ignored for this simple analysis. Likewise, you might, in real life, have an antenna for the signal source, or generator, and an output tuned circuit coupled to a mixer for the load. But let's be completely general and simply specify that at the input of the black box there is a generator of voltage e_1, which sends a current i_1 into the box.

Seemingly a bit strange at first, the load is also viewed as a *generator.* Why not? A voltage e_2 appears across it with its associated current i_2. To a design engineer, the "generator" of voltage e_2 producing a current i_2 is a perfectly logical analysis tool. You may get a bit more feeling for this view if you remember that coupling energy from the load to the input (the load now becomes a signal

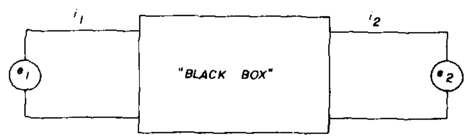


fig. 1. Y parameters of a transistor are defined in terms of the input and output voltages and currents.

source or *generator*) is precisely what is needed to make an *oscillator.* The truth is, there *are* coupling paths through the transistor itself — completely divorced from your external circuit layout — which will make your rf amplifier an oscillator, and that's precisely what this article is trying to help you avoid!

Let us now simply state some results which can be derived mathematically from ac circuit theory, namely

$$i_1 = P_{11}e_1 + P_{12}e_2, \text{ and} \quad (4)$$

$$i_2 = P_{21}e_1 + P_{22}e_2 \quad (5)$$

The mathematical coefficients or "P" parameters in these equations are characteristic of what's in the black box (i. e., the transistor) and are constant except for changes in bias, temperature or signal frequency.

Let's explore the first equation a little. Suppose we *short circuit* the output. The voltage e_2 must then be zero. (Would you hear that VU2 with a jumper across your rf stage output?) The equation then reduces to $i_1 = P_{11}e_1$ under these circumstances. We could also write this as $e_1 = i_1(1/P_{11})$ which is nothing more than Ohm's law in slight disguise. What usually goes in place of the $(1/P_{11})$ is be measured in *mhos,* and it is one of four similar parameters (including P_{12}, P_{21} and P_{22}) in the equations.

Since the parameters are all in mhos – that is units of admittance – they are called admittance parameters or y parameters. Having thus identified them the equations may be written:

$$i_1 = y_{11}e_1 + y_{12}e_2 \quad (6)$$

$$i_2 = y_{21}e_1 + y_{22}e_2 \quad (7)$$

When a spec sheet gives you a number value for a y parameter, it must also specify the operating point, frequency, etc. Curves of the y parameters against frequency or bias are often given.

How does the transistor manufacturer get values for the y parameters published on the spec sheets? Several manufacturers make admittance bridges which do the job and are capable of measuring over a wide range of frequencies – even into the GHz region. Some use capacitors for output ac short circuit, others use tuned lines. Neither short upsets the dc operating point, and, of course, a large number of devices must be measured to get the so-called typical values.

Why should the y parameter be measured by shorting the device? Mathematically, either e must be eliminated by short circuit, or *i* by open circuit to make the equations solvable. At radio frequencies a true rf short-circuit is easier to make than a true open-circuit, so making measurements becomes more practical this way. As hinted before, there are other parameters such as z parameters for impedance, h for hybrid, etc. However, if you have a complete set of one kind of parameter, it can be mathematically transformed into any other kind.

Therefore, the definitions for the y parameters can be written using **eqs. 6** and **7**

$$y_{11} = \frac{i_1}{e_1} \text{ when } e_2 = 0 \text{ (output shorted)} \quad (8)$$

$$y_{12} = \frac{i_1}{e_2} \text{ when } e_1 = 0 \text{ (input shorted)} \quad (9)$$

$$y_{21} = \frac{i_2}{e_1} \text{ when } e_2 = 0 \quad (10)$$

$$y_{22} = \frac{i_2}{e_2} \text{ when } e_1 = 0 \quad (11)$$

These equations apply to any "linear active two-port (that is, input and output) network," (LAN) and are good for bipolar transistors as well as fets and IC rf amplifiers. However, when applied to fets, the number subscripts have yielded to descriptive letters which I will use since the example design will use an fet. **Table 1** cross references and names the parameters. The s in all the designations refers to a common-source configuration.

Note that the y parameters are complex quantities, that is, $y_{is} = g_{is} + b_{is}$ or, input admittance is the sum of input conductance and input susceptance.

Let's see what we can make of these. Looking into the fet amplifier (i.e. gate) you would see both resistance and capacitance. The same would hold true for a circuit looking into the fet output (i. e. drain). These impedances (resistances and capacitive reactances) can be expressed as admittance according to **eq. 3** and are exactly what y_{is} and y_{rs} are when appropriate short circuits are made as outlined before.

Y_{fs} is similar except that it relates to *output current* and *input voltage* and is therefore transadmittance. Remember transconductance in vacuum tubes? Y_{fs} is essentially the same but includes the reactive part too.

Y_{rs} involves output voltage and input current and is the path for signals back through the device. Remember plate-grid capacitance and neutralization of your rf amplifier? Y_{rs} is much the same but includes the non-reactive (i. e. conductive) part of the feedback path as well.

You see, a properly designed vacuum tube normally has no conductive path for electrons from plate to grid, only capacitance. But a transistor does have a conductive path back through the semiconductor material, although it is negligible for many purposes. Y_{rs} then includes both the resistance and capacitance, expressed in terms of conductance and susceptance.

stability

Several years back, J. G. Linvill[1] devised a method of determining the stability of a device using the y parameters. The Linvill stability factor C is given as

$$C = \frac{|y_{12}y_{21}|}{2g_{11}g_{22} - Re(y_{12}y_{21})} \quad (12)$$

Absolute values and real parts are discussed in the appendix (g_{11} is the real part of y_{11}, etc.). When C is less than 1, the device is unconditionally stable. If it is *greater* than 1, it is potentially unstable.

Since the Linvill stability is taken for the worst possible case, that of infinitely large source and load resistances (i. e. open circuit), there arises the possibility of rendering the potentially unstable device tractable. Note that I say *device.* This stability refers to signal paths through the transistor, not to paths due to stray circuit capacitance, etc. A. P. Stern[2] has defined the Stern stability factor K which includes the effects of input and output loads as

$$K = \frac{2(g_{11} + G_S)(g_{22} + G_L)}{|y_{12}y_{21}| + Re(y_{12}y_{21})} \quad (13)$$

G_S and G_L are the conductances of the source and load impedance respectively. If the value of K is less than 1, the amplifier is unstable. Values of K around 2 to 4 should be satisfactory for a well laid out amplifier to be stable. For some devices at some loads K values over 100 appear.

The first step in amplifier design then, is to compute C. If it is less than 1, stable

table 1. Y parameters for field-effect transistors. The subscript s indicates a common-source configuration.

$y_{11} = y_{is}$	input admittance
$y_{12} = y_{rs}$	reverse transadmittance
$y_{21} = y_{fs}$	forward transadmittance
$y_{22} = y_{os}$	output admittance

design in easy, and only external feedback paths must be eliminated by proper layout. If C is greater than one, you must compute K, adding in the source and load conductances which are frequently derived from the resonant impedance of a parallel tuned circuit.

The tuned circuit is designed for inductance, capacitance and loaded Q values compatible with the selected goals in impedance matching, bandwidth, etc. The tuned circuit design may lead you into conflict, however. Maximum gain occurs when the source and load are matched to the transistor. However, for many purposes quite wide mismatching affects the gain by few enough dB that gain can be sacrificed to achieve other objectives.

Indeed, deliberate mismatching is one way to achieve a large K and a stable amplifier. You can see in **eq. 13** that as the source and load resistances decrease, G_S and G_L increase, and, being in the numerator, increase K. So, smaller source and load impedances make for a more stable amplifier.

To see how badly mismatching affects gain, you can compute the gain from the y parameters:

$$\text{Gain} = \frac{|y_{21}|^2 G_L}{|Y_L + y_{22}|^2 \, Re(y_{11} - (y_{12}y_{21}/(y_{22} + Y_L))} \quad (14)$$

For a tuned circuit at resonance Y_L equals G_L.

feedback

A second method may be used to achieve *stability – feedback. There are two kinds provided by proper feedback networks.* The first is unilateralization which reduces y_{12} to zero. Since there is then no input/output communication, the amplifier is stable if well laid out, and a bonus is that tuning the output can't detune the input (good input/output isolation).

The second feedback scheme is neutralization which reduces y_{12} to some value other than zero. The common circuit with a neutralization capacitor wipes out the reverse transsusceptance, but not the reverse transconductance, so y_{12} is not zero. Maximum input/output isolation is not achieved, but perfectly sufficient stability may be.

In summary then, mismatching is the easiest way to achieve stability since it requires no additional circuitry, but is achieved at the expense of gain. This loss may be small, and quite tolerable, however. You have seen reference to detuning to make an rf amplifier stable; this is pretty crude mismatching. It is likely to be quite costly in gain, and is no substitute for selection of optimum resonant circuit impedances which will meet the various stability, bandwidth and other criteria.

Finally, before some examples, there are several other useful expressions which involve y parameters and are used by the sophisticated design engineer. I will mention only two in passing, the input and output admittance equations:

$$Y_{in} = y_{11} - \frac{y_{12}y_{21}}{y_{22}+Y_L} \qquad (15)$$

$$Y_{out} = y_{22} - \frac{y_{12}y_{21}}{y_{11}+Y_S} \qquad (16)$$

Normally, in a tuned rf amplifier, the reactive (imaginary) parts of Y_{in} and Y_{out} simply become part of the tuned circuit reactances and disappear when tuning for peak response. The real parts G_{in} and G_{out} load the tuned circuits and enter into their design for proper bandwidth; that is, they contribute to the value of loaded Q. They are expressed as equivalent resistances which appear in parallel with the tuned circuit and figure into the calculations for circuit Q.

practical design

I will now give a design example with some numerical values with which you may check your understanding of complex calculations. A set of y parameter curves is given for the MPF121 dual-gate mosfet in **fig. 2.** The values are in millimhos; the answers will be also. Let's try the design of converters for 2 and 6 meters with a 10- to 14-MHz tunable i-f provided by the station receiver. Reading off the values at 50 and 150 MHz, for 6 meters we get:

$$y_{is} = 0.08 + j1.2$$
$$y_{rs} = 0 - j.0065$$
$$y_{fs} = 12.9 - j1.4$$
$$y_{os} = .08 + j.7$$

and for 2 meters:

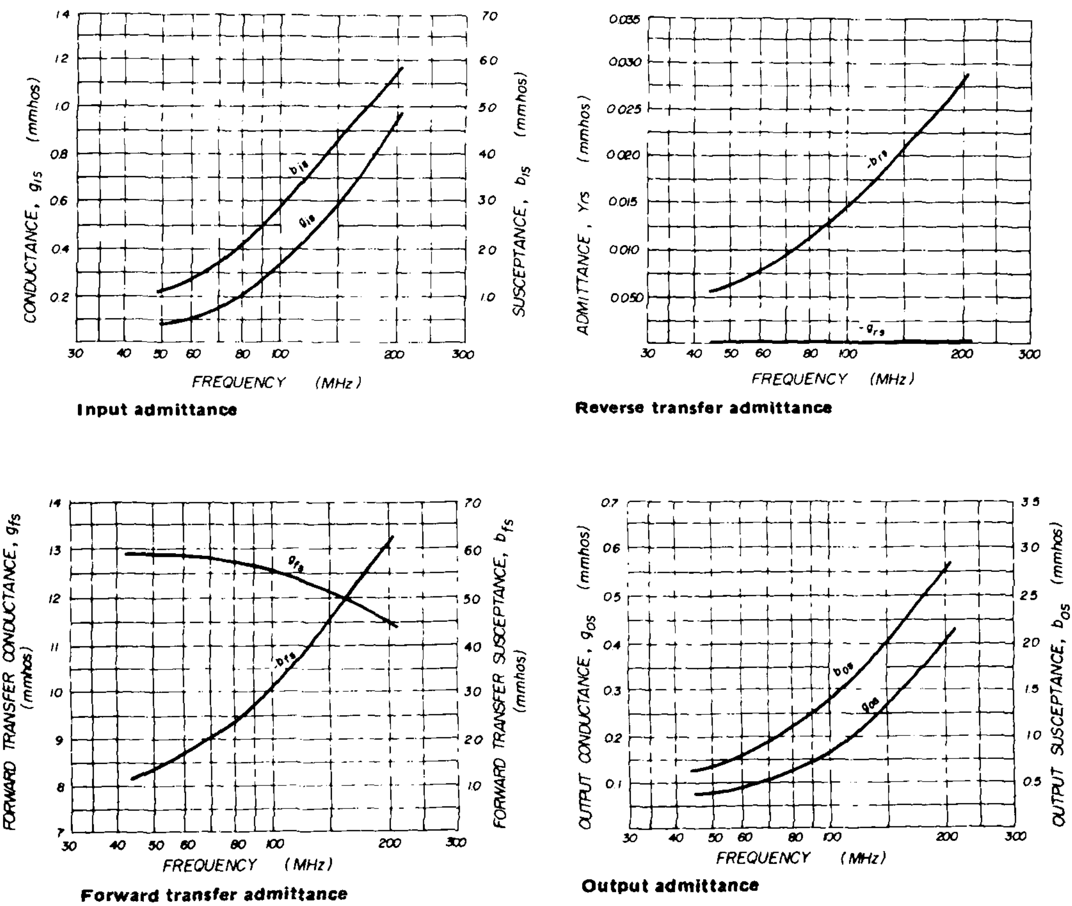
$$y_{is} = .69 + j4.5$$
$$y_{rs} = 0 - j.023$$
$$y_{fs} = 12.1 - j4.7$$
$$y_{os} = .28 + j2.1$$

First compute the Linvill stability for these two cases. It turns out to be 0.60 for 2 meters and 3.9 at 6 meters. Thus, the MPF121 is unconditionally stable at 144 MHz and conditionally unstable at 50 MHz. Let's proceed with the simpler design first, for 2 meters.

two-meter design

Using normalized tuned-circuit response curves such as those in the *Radio Amateur's Handbook,* and starting somewhat arbitrarily with a reactance of 150 ohms, you find that the required inductance is 0.16 μH, with a capacitance of 7.3 pF. For a response across the 144-148 MHz band within 1 dB down, the required loaded Q is 16, and the parallel resonant impedance of the tuned circuit is 2400 ohms.

This last fact means that the total effective parallel resistance across the tank must be 2400 ohms if our criterion of 1 dB response is to hold. This resistance includes that contributed by the input resistance, while the reciprocal of the susceptance yields the input capacitive reactance. (Remember that millimhos yield answers in kilohms.) The susceptance may be converted to a capacitance



Input admittance

Reverse transfer admittance

Forward transfer admittance

Output admittance

fig. 2. MPF121 dual-gate mosfet common-source admittance parameters (V_{DS} = 15 Vdc, V_{G2S} = 4.0 Vdc, I_D = 6.0 mA dc).

circuit components, that transformed from the antenna, and that due to the input resistance of the fet. The latter may be computed from the y parameters according to the formula for Y_{in}.

Y_{in} of course consists of a real part, the input conductance, and an *imaginary* part, the input susceptance. Taking the reciprocal of the conductance gives the using $C = 1/(2\pi f X_c)$, and turns out to be 4.9 pF. This is part of the 7.3-pF tank capacitance, a goodly part indeed; you may need to recalculate for a higher C circuit.

The input resistance is quite low at this frequency, only 1200 ohms. Suddenly you realize that this will degrade the Q if the fet is connected across the whole tank. This may not hurt; however, you should look for possible poor image response.

At Q = 16, the response curves show that an image at 122 MHz would be 15 dB down. Using a second identical tuned circuit at the output would yield a ± 1 dB response across the band and image response of -30 dB. You would most likely want to tap the fet input down on the tank to preserve the Q under these circumstances.

Similar procedures apply to the output circuit with fet output resistance of 2900 ohms and output capacitance of 2.3 pF. *Incidentally, the Stern stability for 2400 ohms source and load resistances turns out to be 8.1 and the calculated gain 12.4 dB.* The fet is thus very stable, but it is up to the builder to avoid external feedback paths in his layout.

six-meter design

Let's look at the 6-meter case next. Starting with a reactance of 150 ohms, and ± ½ dB response across 50 to 51 MHz (i. e., 1 dB down at the band edges), you find that L = 0.47 μH, C = 21.0 pF and loaded Q must be 22.2. This gives a resonant impedance of 3333 ohms, and an image response at 30 MHz of -25 dB for the single circuit. If you use identical input and output tanks, the response across the band segment would be ± 2 dB and the Stern stability 3.8, which is adequate. The calculated gain is 26.5 dB. Suppose the slug-tuned coils had unloaded Qs of 80. R = QX = 80(150) = 12k which is the resistance contribution of the coil alone. The input resistance of the MPF121 is 5620 ohms, so the following parallel resistance relationship must hold:

$$\frac{1}{5620} + \frac{1}{12K} + \frac{1}{R_{ant}} = \frac{1}{3333} \qquad (17)$$

Solving, R_{ant} must be 26k. Thus, the antenna (perhaps 50 ohms) must be transformed to 26k by the input circuit in order to preserve the loaded Q of 22.2. Frankly, this makes an antenna coupling of rather poor efficiency.

The design has several alternatives at this point, some of limited usefulness, including: tap the fet down on the tank; use a higher Q coil; accept a lower Q through tighter antenna coupling with its wider bandwidth and poorer image rejection; use a coupled circuit to gain selectivity and largely separate the fet and antenna loadings. Choices like these must be weighed to arrive at a final good design.

Suppose you wish to use a transistor for a high-frequency receiver, but no y values are given below 30 MHz? After inspecting a number of data sheets, a few of which had values for lower frequencies, and assuming that dual-gate mosfets have similar general characteristics, even though they are advertised for a given frequency range, I conclude that the following rules-of-thumb would be better than using single values for the high-frequency region:

1. g_{rs}, g_{fs} and g_{os} are constant throughout the hf region.

2. The four susceptances may be approximated assuming they are proportional to frequency. That is, a b_{is} of 1.1 at 40 MHz would be close to 0.4 at 14 MHz. Use the parameter value at the lowest frequency given on the spec sheet as a starting point.

3. g_{is} appears to fall off by a factor of four per octave. For instance, a value of 0.24 for 10 meters would reduce to 0.06 for 20 meters, 0.015 for 40 meters, etc.

Note that y_{is} means *input* admittance in the common-source configuration. What about common-gate and common drain? Formulas to convert the common-source parameters to the other configurations, Y_{ig}, Y_{id}, etc., have been worked out and are given in reference 4. All equations for stability, gain, etc. are applicable to the common-gate and common-drain configurations as they stand, simply by using the appropriate y parameters for the configuration of interest.

summary

I may summarize then, with some

rules-of-thumb which may prove beneficial even if you make no calculations:

1. A typical dual-gate mosfet will be unconditionally stable if the frequency is high enough.

2. A potentially unstable device may be stabilized by feedback or mismatching.

3. Mismatching may be preferable for stabilization since it requires no extra components, no adjustment, is not frequency sensitive, and the cost in gain is usually tolerably small for amateur work.

4. The smaller the source and load impedances, the greater will be the stability. Thus, at parallel resonance, smaller values of both reactance and Q give smaller parallel impedance and greater stability.

In conclusion, I must point out that there are techniques in rf amplifier design used by the professional design engineer which have not been mentioned here. Some of these methods require access to a computer since the number of computations may be vast. In addition, I have not pursued the design examples through to final optimization, but rather pointed out procedures and design choices. As you can easily see, there are tradeoffs between many factors such as selectivity, high gain, optimum noise figure and reasonable cost which a designer must evaluate.

Ultimately, the stability criteria are of key importance since no one can accept an unstable oscillating rf amplifier! Although I was privileged to program and run my calculations on a computer, I hope that someone may be brave enough to tackle the calculations and possibly avoid generating his own input signals in that new receiving system!

appendix

complex numbers

The square roots of negative numbers are designated *imaginary numbers* and there are definite algebraic rules for dealing with them. The square root of minus one is the basis for complex operations, and it is generally given the symbol i in mathematics while engineers use j. Thus, $i = j = \sqrt{-1}$, and from the rules governing square roots, you can verify that $j^2 = -1$, $j^3 = \sqrt{-1}$, $j^4 = 1$ and so forth. Ordinary algebraic rules also apply such as $(j)(j) = j^2$, $j^3/j = j^2$, etc. Any nonimaginary number is a *real* number.

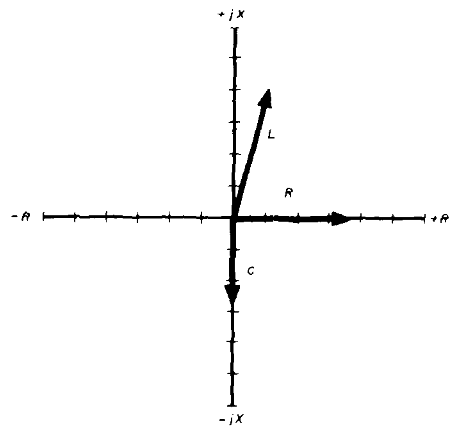


fig. 3. Circuit elements L, R and C plotted on a complex coordinate system. Each division is 100 ohms.

A *complex number* is the sum of two numbers, the first real and the second imaginary, such as $3 + j2$, $2.5 - j7$, or in general, $a + jb$. The properties of j and complex numbers make them suitable for describing the characteristics of resistors, capacitors and inductors plotted on coordinate axes on the *complex plane.* These plots make it easy to visualize the behavior of these circuit elements, and the associated complex algebra makes it possible to determine their behavior with slide rule or computer.

The basic reason that the complex plane is needed is that the voltage is 90° out of phase with the current in any pure reactive element. Thus, in **fig. 3,** the complex coordinate system is shown with three circuit elements plotted thereon, all at a frequency of 60 Hz. A pure 350 ohm resistor is shown at R. It lies on the R axis and has no component along the jX axis. A pure 10 μF capacitance is labeled C and since $X_c = 1/(2\pi fC)$, it has 270 ohms capacitive reactance at 60 Hz. The voltage across the capacitor is 90° behind the current, and the capacitive reactance is along the negative jX axis.

The 1.06-henry inductor is not pure since the wire has a resistance of 100 ohms. (Any *real capacitor has some resistance* associated with it too!) The inductive reactance is $+jX = +j(2\pi fL)$ (positive since the applied voltage leads the current by 90°) and has a value of 400 ohms. The inductor then has two components, 400 ohms of pure inductive reactance and 100 ohms of pure resistance (plus some stray

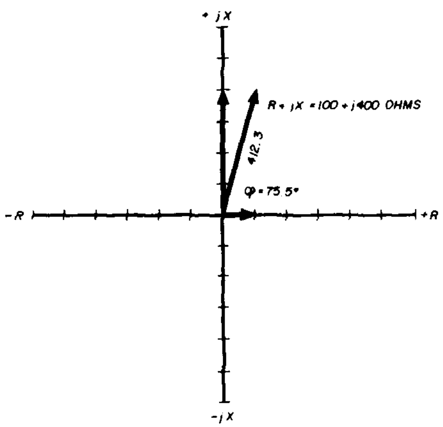


fig. 4. R + jX characteristics of 1.06-H inductor discussed in text. Each division is 100 ohms.

capacitance in any real inductor which we ignore). These two quantities have the properties of vectors, and are shown plotted in **fig. 4**.

The two components R and jX lie along the axes while the representation of the inductor R + jX = 100 + j400 ohms is the resultant of the two, and is a complex number. We say that the device has 100 ohms resistance plus 400 ohms inductive reactance. The sum of these (and it is a *vector* sum) has a magnitude obtainable by the Pythagorean theorem as follows:

$$\sqrt{100^2 + 400^2} = \sqrt{170{,}000} = 412.3 \text{ ohms}$$

It also has a phase angle θ which is 75.5°.

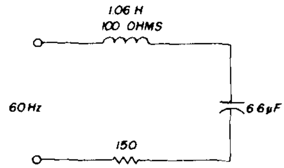


fig. 5. Simple RLC circuit. Complex characteristics of this circuit are plotted in fig. 6.

Both the complex representation 100 + j400 ohms and the vector form 412.3 + ∠75.5° ohms are equivalent and are simply different ways of stating the *impedance* of the inductor Z_L. We may then state the general relationship that Z = R + jX.

As an example of how the behavior of a circuit can be visualized, plot the characteristics of the circuit in **fig. 5** on the complex plane. The capacitor has a reactance of −j400 ohms. All components are shown in **fig. 6A**. All components are added vectorially together to find the result; you can see that *−j400 cancels* +j400 while 150 + 100 = 250 ohms. This can be done by computation alone as follows:

$$Z_{total} = Z_R + Z_C + Z_L$$
$$= 150 - j400 + 100 + j400$$

Real and imaginary parts are grouped together giving

$$Z_{total} = (150 + 100) + (j400 - j400) = 250 \text{ Ohms}$$

If you change the frequency to 120 Hz,

$$Z_{total} = Z_R + Z_C + Z_L$$
$$= 150 - j200 + 100 + j800$$
$$= 250 + j600 \text{ ohms}$$

and at 30 Hz

$$Z_{total} = 250 - j600 \text{ ohms}$$

These values of L and C are series resonant at 60 Hz since the inductive and capacitive reactance cancel at series resonance; the current is determined only by the resistor and the resistance in the inductor. Incidently, you may relate the negative R axis to *amplification.*

In the preceeding manipulation of complex numbers, no special techniques were required. But multiplication and division must be added to the list. Let's try an example by multiplying the following complex numbers:

$$(2 + j3)(3 - j5)$$

This operation is carried out just as the algebraic multiplication of two binominals:

$(a+b)(c+d) = ac+ad+bc+bd$
$(2)(3) = 6$
$(2)(-j5) = -j10$
$(+j3)(3) = +j9$
$(+j3)(-j5) = (+j)(-j)(+3)(+5) = -(-1)(15)$

Adding all terms, we get $(21 - j1)$

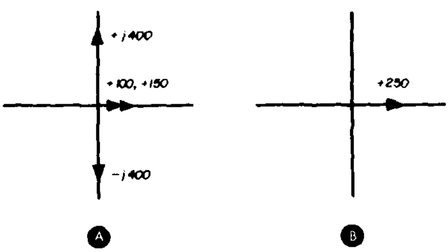


fig. 6. Complex characteristics of the circuit in fig. 5.

Division of complex numbers requires the use of the *complex conjugate* to get rid of the imaginary parts in the denominator. For any complex number $(a + jb)$ its complex conjugate is defined as $(a - jb)$, and vice versa. Let's see what happens when a number is multiplied by its conjugate, remembering that $(a + b)(a - b) = a^2 - b^2$:

$$\begin{aligned}(a+jb)(a-jb) &= a^2 - (jb)^2 \\ &= a^2 + b^2\end{aligned}$$

The j's drop out. Division is performed by multiplying the denominator by its complex conjugate, and also multiplying the numerator by the same quantity so that the value of the fraction remains unchanged as follows:

$$\begin{aligned}\frac{4 + j8}{2 + j6} &= \frac{(4 + j8)(2 - j6)}{(2 + j6)(2 - j6)} \\ &= \frac{8 - j24 + j16 + 48}{4 + 36} \\ &= \frac{56 - j8}{40} = 1.4 - j0.2\end{aligned}$$

You thus get rid of j's in the denominator and can easily get a simple answer.

Two additional complex manipulations will be necessary. The first is commonly designated $Re(a + jb)$, and means the real part of $(a + jb)$, or just a. Thus $Re(5 - j2) = 5$. Secondly, $a + jb$ designates the absolute value. It is always positive, and by definition, is the magnitude of the quantity when expressed in vector form. Thus

$$|5 - j2| = \sqrt{5^2 + 2^2} = \sqrt{29} = 5.4$$

For the inductor in **fig. 4**, $(100 + j400)$ ohms, you may write $Re(100 + j400) = 100$ and $|100 + j400| = 412.3$

In summary, addition, subtraction, multiplication, division and the evaluation of real and absolute values just about covers the needed techniques to get started with complex numbers in a useful way.

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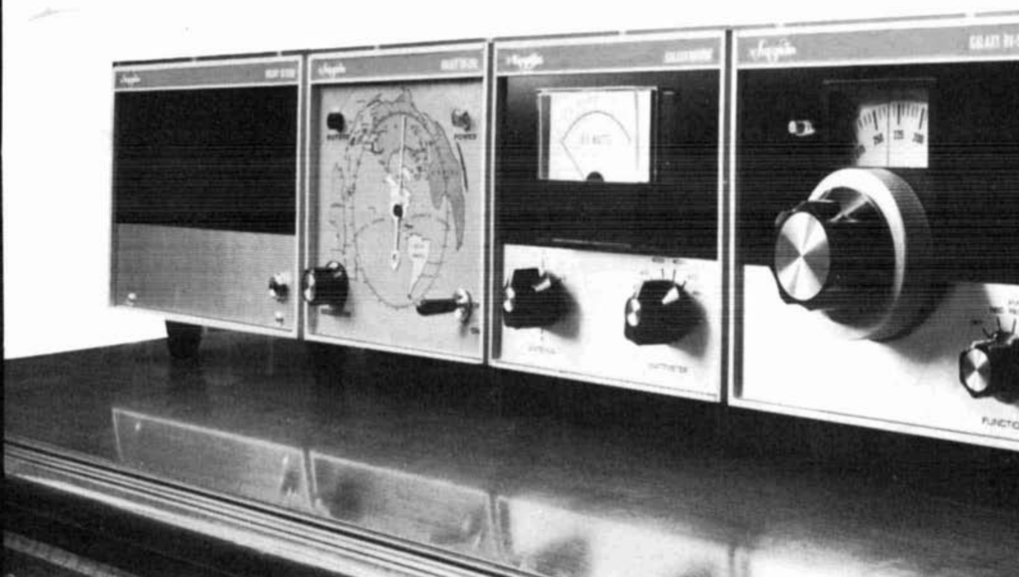

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ham radio sweepstakes winners

W3PTG wins grand prize, puts all new Drake station on the air; WAØKKC proudly using new Standard fm station on two meters

Skip Tenney, W1NLB, Publisher, Ham Radio Magazine

Ham Radio Magazine's 1972 Sweepstakes is now history, and it was a busy bit of history at that. After many months of hard work by our staff here in Greenville (have you ever tried to open and read over a thousand pieces of mail in one day?), the big drawing finally took place as scheduled on May 18th. As you can see, we had quite a box full of entries. In fact, it was a large job just to mix them up thoroughly so that everyone had an even chance.

Many of the people who bring you HAM RADIO each month gathered to watch Pat Hawes draw the name of this year's Sweepstakes winner—Gus Haak, W3PTG.

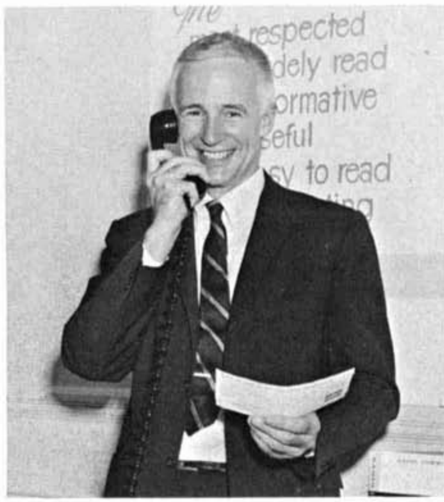
winners of
Radio Communications Handbook

W1VLD	W5BL	W9IQN
WA1FXA	W5RDE	W9JQY
WA1MFI	K5SZH	W9JYY
W2PKY	WA5OQR	K9ALD
W2VLS	W6CTY	K9GET
WB2QKQ	K6CMV	K9RYW
W3BSE	K6SEQ	WA9GAY
W3GWM/1	WA6PQG	WA9SCD
W3MCL	WA6TMQ	WB9ADL
W3TOL	W7MUG	WN9FIJ
W3UZN	WA7BKW	WØCOE
WA3PJL	W8WRJ	WØZLO
W4KQK	W8YBM	KØKKK
K4EHP	K8ZPR	KØOJR
K4SHJ	WA8ODV	WAØNUG
WB4VGR	WB8BOT	WAØUEN
WN4ZCF	WB8LAT/9	

The winner of the grand prize was Gus Haak, W3PTG, who by now has had his new Drake TR-4, RV-4 and L4-B in operation for over a month. Gus was a perfect choice for this gear as he is active on 10, 15 and 20 meter sideband and was running a considerably more modest station before now.

This new station provides him with an outstanding signal on all the hf bands. The Drake TR-4 is one of the most respected ssb transceivers on the market while the RV-4 allows split frequency operation. The L4-B linear permits 2 kW PEP ssb operation and 1 kW a-m, CW and RTTY on a continuous-duty basis.

Our second prize went to Dick Mollentine, WAØKKC. Again, the prize was ideally suited to it's winner's needs. Dick has primarily been active on 6-meter a-m until now, and he had been eyeing 2-meter fm with great interest.



Ham Radio publisher Skip Tenney, W1NLB, telephoning W3PTG to tell him he has just won the 1972 GRAND PRIZE.

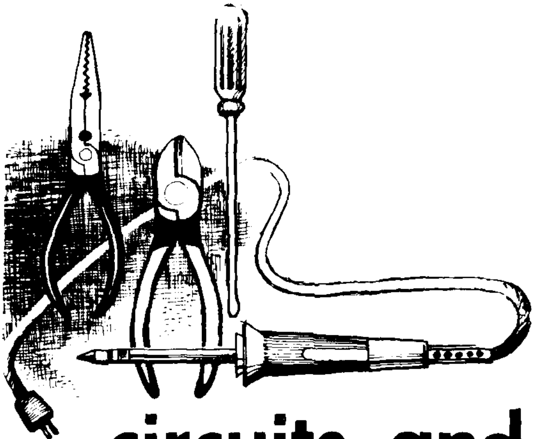
Assistant editor Doug Stivison, WA1KWJ, enjoys a contact before sending off Standard's versatile SR-C146 handheld fm transceiver to this year's winner.

With his Standard trio consisting of a SR-C146 hand-held 2-meter fm transceiver, SR-C826M mobile rig and the all-new SR-C14 base station, he can claim to have the perfect piece of fm gear for every operating requirement. Of particular interest is the base station. When the prize was awarded it was one of the first of this model to arrive in this country!

The 50 third prizes of Radio Communications Handbooks pretty well covered the whole country as you can see from the list of winners. We were glad to find that the list included three YLs and two novices, along with a wide variety of calls, both old and new.

Again, we would like to thank the many thousands of you who entered. Someday we hope to figure out a way so that everyone can win.

ham radio



circuits and techniques

ed noll, W3FQJ

ic flip-flops

Last month this column contained a series of experiments that permit you to observe the operational characteristics of various gates and a basic flip-flop multivibrator. Data on the construction of a 100-kHz calibrator oscillator and the use of a universal flip-flop as a two-to-one counter was presented. In this month's column, specific experiments permit you to take a more detailed look at the operational characteristics of a universal (JK, R-S and clocked) digital multivibrator.

universal flip-flop logic

The very simple circuit arrangement of **fig. 1** is appropriate for checking the logic voltages at the various terminals of the SN7472 J-K master-slave flip-flop. Use Fahnestock clips or binding posts as a convenience in making circuit changes. The unit has three sets of J and K inputs; only one set need be used. Other inputs are R, S (clear and set) and clock. At each of these inputs, logic 1 is the no-connection condition. Logic 0 is established by shorting any one of the inputs to ground. This simple procedure permits you to check out the dc logic 0 and logic 1 operating conditions.

The R and S inputs are overriding. When they are used the unit operates as a simple R-S flip-flop. A basic flip-flop logic diagram is given in **fig. 2.** Generally when the set input is logic 1, the Q output is logic 1 and the $\bar{Q}$ is logic 0. In the case of the SN7472, there is an inversion in the S and R line as indicated by small circles in the logic diagram, **fig. 2B.** This means that with input S set to logic 1, the Q output is logic 0 and the $\bar{Q}$ output is logic 1.

step 1:

Connect a voltmeter to the Q output. Leave the S terminal at logic 1 (no connection) and set the R terminal to logic 0 by connecting a short between the R terminal and ground. The Q output is zero and the $\bar{Q}$ output is positive logic 1 voltage (positive logic device). Reverse the logics of inputs R and S. Note the output logics flip over. What are the output logics with both the R and S terminals at logic 0? Do the same with both the R and S inputs at logic 1. The output logic for the latter condition can be 1,0 or 0,1 depending upon the previous logic of the R and S inputs. The above logic conditions hold regardless of the logic at the J, K and C inputs; R and S are overriding.

step 2:

Momentarily apply logic 0 to the R input and logic 1 to the S input to establish the *set* state of the Q and $\bar{Q}$ outputs with $\bar{Q}$ at logic 1. Keep the R and S inputs on logic 1 (no connections). Set the K input on logic 0 and the J input on logic 1. Momentarily apply logic 0 to the clock input. Note that the output logic changes and holds whether the clock input is kept on zero or one. Momentarily apply logic 0 to the R input. This returns the output logic to its set state.

Now connect logic 0 to J and logic 1 to K. Again, momentarily apply logic 0 to the clock input. Note that the output logic again flips and holds. Momentarily apply logic 0 to the S input. Note that the output logic goes back to its reset state.

signal operation

Additional understanding of the operation of the universal flip-flop can be gained by using a signal from an audio sine- or square-wave generator and the 100-kHz crystal calibrator detailed in last month's column.

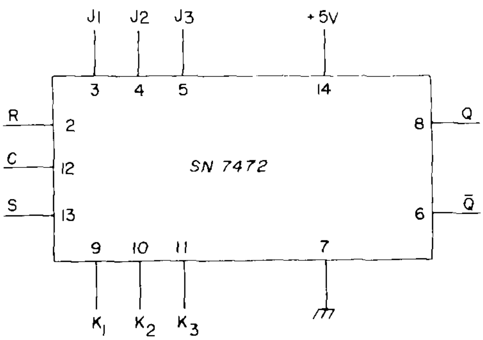


fig. 1. Connection diagram for SN7471.

step 3:

Set all inputs to logic 1 (no connection). Apply the 100-kHz output of the crystal calibrator to the clock input. Observe the signal at the Q and $\bar{Q}$ outputs. Compare the input and output frequencies. The unit is operating as a two-to-one counter. Disconnect the 100-kHz clock.

step 4:

Apply the output of the audio generator across the R-S inputs. To activate the unit it may be necessary to increase the audio gain above that required at the clock input. Output and input frequencies are now the same, but the waveform has been improved and the output is steep-sided compared to the input signal.

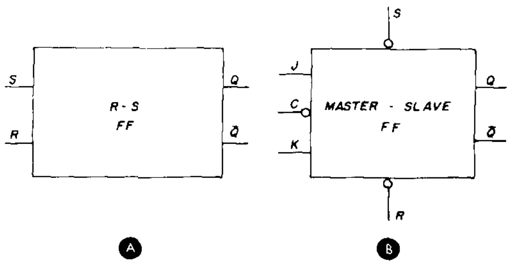


fig. 2. Basic and universal flip-flop.

step 5:

Disconnect the audio generator. Connect the output of the 100-kHz crystal calibrator to the clock input. Observe the output waveforms at Q and $\bar{Q}$.

Connect the output of the audio oscillator to the J input of the flip-flop. Set the audio oscillator frequency to 1000 Hz. Display one cycle of the 1000-Hz repetition rate on the oscilloscope screen. During approximately one-half of its period, the clock signal appears in the output while the second alternation contains no clock component. Momentarily connect 0 logic to the R input. Output is removed and the flip-flop is returned to the set state.

inhibiting

In an inhibiting circuit a certain logic at one of the input circuits prevents or inhibits the transfer of logic information between another input and the output. A basic inhibiting circuit is shown in **fig. 3.**

The transistor is normally non-conducting because the emitter junction is not forward biased. This represents logic zero condition at the base. When a positive pulse of sufficient magnitude (logic 1 level) is applied to the base, the transistor conducts, and its inversion produces a logic 0 output. This is true provided no logic 1 voltage is applied to the emitter.

If a logic 1 pulse arrives at the emitter at the same time a logic 1 pulse is applied to the base, the transistor remains non-conducting. Therefore, the output of the transistor remains normal at logic 1 potential.

Likewise, the two other possible conditions of X and Y both being at logic 0 or both being at logic 1 result in a logic 1 output. The truth chart for the circuit is given with **fig. 3.**

In considering the universal flip-flop you learned that the R and S inputs were overriding. One might state that these inputs inhibit the clock's, J and K inputs just as input Y in **fig. 3** is able to inhibit input X. The inhibiting process is used to advantage in counters because they permit a binary counter chain (all even count) to also function in an odd-count manner. In our next experiment a divide-by-eight counter consisting of three binary 2-to-1 counters in series is made to divide by 5. This is done by feeding back inhibiting information from the output.

even and odd counting

The final construction experiment last

X	Y	Z
1	0	0
0	0	1
1	1	1
0	1	1

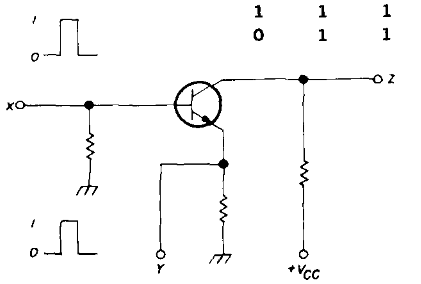


fig. 3. A basic inhibitor circuit and its truth table.

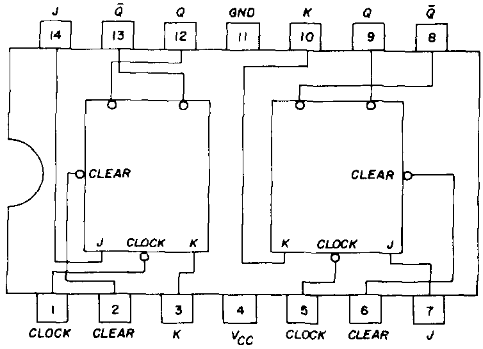


fig. 4. The basic 7473 dual flip-flop containing two binary counters.

month consisted of a crystal-controlled multivibrator and a 7472 two-to-one counter. The versatility of this calibrator can be extended with the addition of still another integrated circuit. In this case it is the inexpensive 7473 dual J-K master-slave flip-flop. Two binary counters are included in the same case with separate inputs, outputs and clear terminals, **fig. 4.** They can be operated separately or joined together to obtain an overall count of four-to-one. Combined with the previous counter, a total division of eight (2 x 2 x 2) can be obtained.

step 1:

Connect the circuit as shown in **fig. 5.** The output of the previous counter is joined to the new dual flip-flop by connecting its $\bar{Q}$ output to the clock input, pin 1. The two counters in the 7473 are joined by tying together the $\bar{Q}$ output at pin 13 to the second clock input at pin 5. A divide-by-eight output is made available at either pin 8 or pin 9.

step 2:

Turn on the calibrator circuit, connect the oscilloscope to the output of the crystal oscillator and display eight cycles on the scope screen. Now connect the oscilloscope to the output of the last counter. Note that only a single waveform is displayed, indicating a division by eight. Two cycles will be seen at the junction between the two dual counters and four cycles at the output of the first counter. Waveforms will be those of **fig. 5.**

step 3:

Now connect the circuit of **fig. 6.** A 2000-pF capacitor is used to connect the Q output of the last counter to the reset or clear inputs of the first and second counters. Capacitor C1 provides a feedback path for a reset pulse. Pulse polarity is such that it cancels out or inhibits the activities of the first two counters at the proper time to start a new five-pulse sequence.

Note in **fig. 5** (divide-by-eight counter) that the edges of all five waveforms are coincident at the leading edge of the fifth clock pulse. Furthermore, the trailing edge of the Q output of the last counter is swinging from logic 1 to logic 0, a polarity which can be used to clear and reset the first two counters. This very activity is shown as coinciding with the leading edge of the fifth clock pulse in **fig. 6.** The leading edge of the sixth clock pulse will now switch all three counters just as the leading edge of clock pulse one at the very beginning. The counters see everything as starting anew and they go off to try and count eight again only to

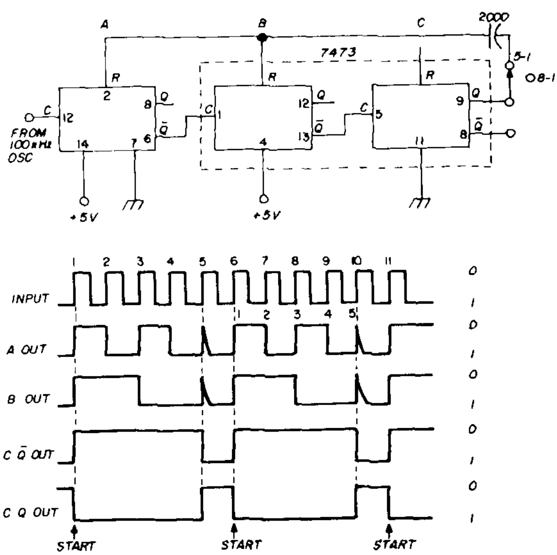


fig. 6. A 5-to-1 divider and corresponding waveforms.

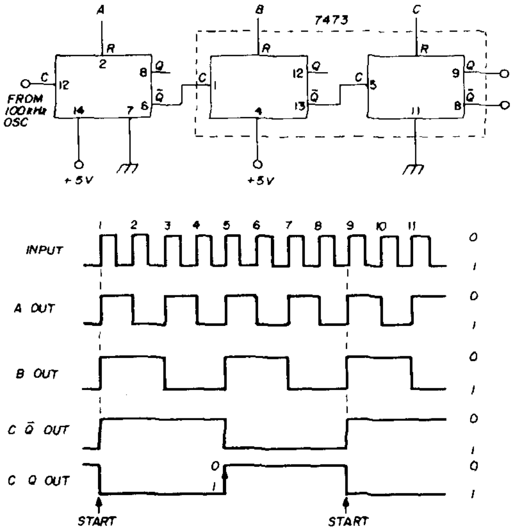


fig. 5. An 8-to-1 divider and corresponding waveforms.

be met by another reset pulse coincident with the leading edge of clock pulse ten.

The feedback spikes shown in the counter 1 and counter 2 outputs coincide with the leading edge of the fifth clock pulse. Observe these spikes on the oscilloscope. Note how the entire cycle of events repeats itself. Actually, the counter chain is never permitted to go through its eight-count cycle but is interrupted in a manner that continues a five-count sequence.

step 4:

Turn on the counter and observe the waveforms at the outputs of each counter. Look for the inhibiting spike in the outputs of the first and second counters, exactly as shown in the waveforms of **fig. 6.**

step 5:

Attach the oscilloscope to the output of the last counter and adjust it until eight cycles are displayed. Note also that the output is now asymmetrical (pulse of shorter duration then the intervening spacing). Recall that the outputs for the even-count activity were symmetrical square waves. Momentarily disconnect the capacitor from the Q output of the last counter. How many cycles are displayed now? The number has dropped to five, indicating a change back to an even eight-to-one count.

recommended by H. Beverage in his original paper. The correction to the 200-ohm terminator can be multiplied by quite a large factor due to the somewhat non-linear instantaneous current distribution in any terminated wire. The value of radiation resistance may differ slightly from the characteristic impedance owing to the termination along the wire due to radiation in transmitting or losses in the wire in receiving. In these cases, the values for the feedline and the termination resistance would be different. Therefore, both input and output values should be raised or lowered. These values should be set somewhat toward those values necessary for the center wavelength to be used. Variations exist as to a particular height and location and according to the principal height of the electrical ground.

"Due to the high ground losses, the Beverage can never produce the high signal results as the same antenna if it were remote from earth, but all the same, the practically complete attenuation, (in the earth) under the Beverage of most *minor lobes gives an* apparent boost to received signals arriving off the terminated end."

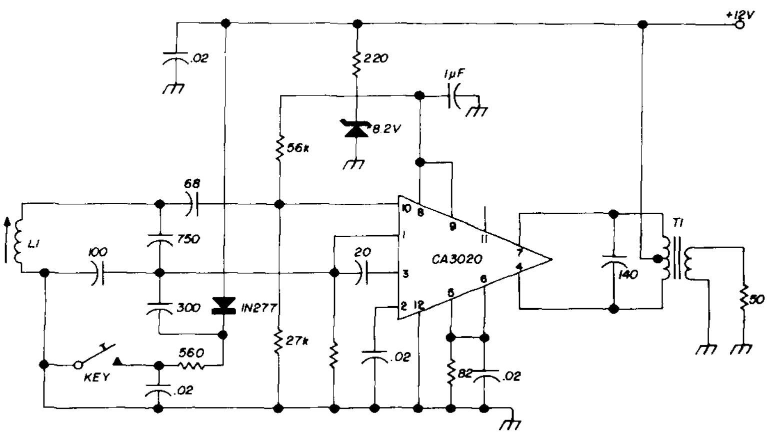


fig. 7. Seiler vfo with CA3020 developed by W9ZTK. T1 is a toroidal transformer.

step 6:

Operate the counter in the divide-by-five mode. Connect the output to the input of your receiver. As you tune over the band you will find a calibration point every 20 kHz. This pulsed output is again rich in harmonics with marks to be found over the 6-meter band, and even on 2 meters with proper low-loss coupling.

Beverage termination resistance

Here is an instructive note from Robert N. Morris, W7ALU. It has to do with the normal 200-ohm termination value for a Beverage antenna,

" . . . this is a close to theoretical value as per the radiation resistance of a non-resonant terminated antenna about three and one-half wavelengths long and being remote from earth. The Beverage however, *not being remote from earth, has a* very high earth loss for both transmitting and receiving. A much higher order of termination would be indicated as being necessary due to the very low height of the Beverage, ten to twenty feet being

keyed vfo

Cal Sondgeroth, W9ZTK, *modified the*

CA3020 QRP unit described in the August, 1971 issue of *ham radio*. His stable and clean keying circuit is shown in **fig. 7**. His description is as follows:

"... as you can see, I used a slug-tuned coil for L1 with the slug attached *to a shaft for tuning from the front panel.* Grounding the slug was important to eliminate hand-capacitance effects. The shaft is fitted with a little bearing near the panel, and this tuning arrangement works out to be very stable even though it's not calibrated.

"The main drawback of the oscillator at first was that it drifted quite badly every time it was turned on. Putting a soldering iron on the IC seemed to indicate that it was pulling the frequency as it warmed up. Attempts to let it run all the time and open the emitter lead of the oscillator transistor to shut off the oscillator really didn't seem to help much, and this doesn't provide good keying for CW either. So, I came up with the diode-keying circuit shown. This allows everything to be on all the time to level out the temperature of the IC. With the key open, oscillation is prevented by the fact that there is no capacitance (or 100 pF as *in my circuit)* from emitter to ground on the oscillator. Oscillations can be stopped in this way, or the oscillator can be shifted off the oscillating frequency by moving about 100 kHz during standby. The frequency shift can be varied by the amount of capacitance switched in and out of the circuit. For CW keying, of course, it would be necessary to completely stop the oscillator. Frequency-*shift keying for RTTY can be obtained* by the same sort of circuit with smaller values of shift capacitance.

"My output transformer is a toroid coil with bifilar primary. The secondary is matched to about 50 ohms, and I did not get as much output as I expected. However, this is a secondary consideration for a vfo, anyway. With the keying circuit shown and the zener regulator, the keying is now *really stable and after all,* stability was what I was after."

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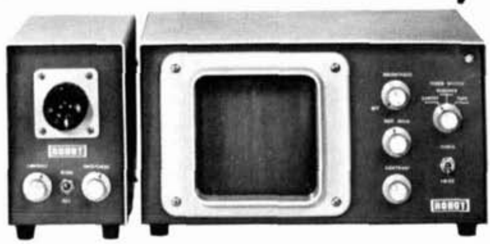

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the ham notebook

simple intercom

Here's a simple intercom that can be put together in a couple of nights and is guaranteed to pacify the wife when you're in the basement and dinner is waiting on the table. It uses a half-watt audio IC, Motorola MC 1306P ($1.10). This IC was designed for use in portable a-m/fm radios and tape recorders and takes a regular 9-V transistor radio battery for power. Since these circuits usually have a high-impedance output, this IC was made with a high-impedance input. The output, however, is low impedance and will work directly into an 8-ohm speaker without the usual output transformer.

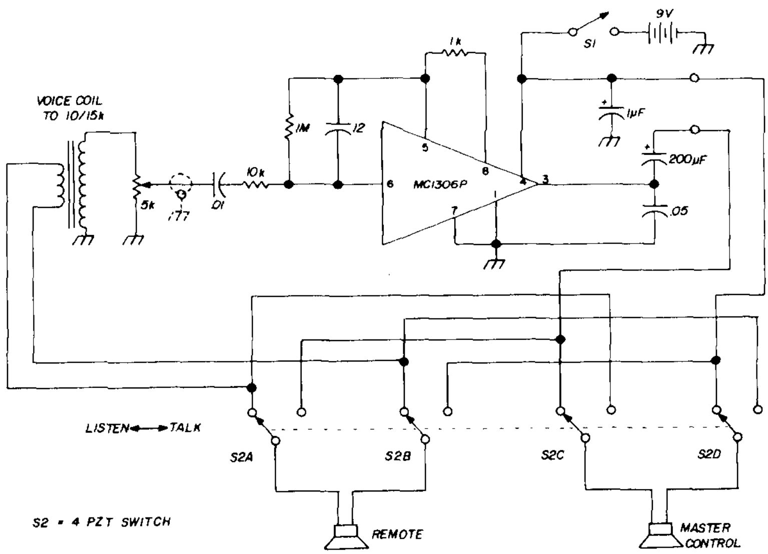

fig. 1. Schematic of the simple intercom.

The circuit, shown in **fig. 1**, is very simple. I made a small etched-circuit board for the IC and associated components. For the master control, I used an old ac-dc radio, with all parts removed execpt the speaker and its output transformer. The output transformer is necessary to step up the input (talking) speaker to a high impedance. The etched-circuit board and battery were mounted inside and the *talk* switch (Burstein Applebee 18A1309) was mounted in the dial hole with the volume control in the regular volume control hole. Shielded cable from the volume control to the input may be necessary. The remote can be any 8-ohm speaker in whatever enclosure is available.

Once installed, the remote is on all the time and any activity or talk can be heard by the control. By pushing the switch to the *talk* position, the master control can talk to the remote. In my setup the remote picked up so much noise from machinery that it was distracting. To remedy this I ran a 4-wire cable instead of the usual 2-wire. I used the two extra wires for a buzzer at the control end and a push-button switch at the remote end. I

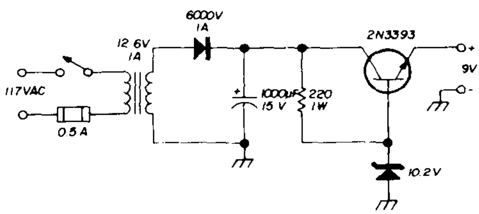

fig. 2. Power supply to operate the intercom from a 117 Vac source.

now leave the unit off (unless I want to listen in or call the remote) and the remote can buzz when someone wants to talk. Since there are no tubes to heat up, the unit will work as soon as it is turned on. This also conserves battery power. An ac supply can be used, and one is shown in **fig. 2**. This will probably have to be mounted outboard to avoid hum pickup from proximity to the circuit board.

When this unit was first turned on it picked up the local broadcasting station, but a 0.002 bypass capacitor from one side of the remote line to ground cured the trouble. In extreme cases of this type, try different values of capacitors on either or both sides of the line to ground. Perhaps chokes in both lines would help.

The IC will take up to 12 Vdc, but more than 9 V will make it run hot and

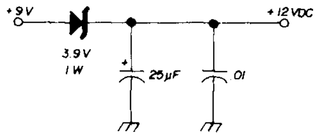

fig. 3. Voltage dropping and regulating circuit to operate the intercom on 12 Vdc.

high voltage isn't really necessary. If you have 12 Vdc available and want to get 9 V, try the circuit in **fig. 3** using a 3 V, 1 W zener. A dropping resistor will not work as the current varies with speech input and a stable 9 V would not result. Also note that neither side of the remote line is grounded because the positive voltage is on the line for the output speaker.

Nat Stinnette, W4AYV

miniature power supply transformers

When powering really compact equipment such as electronic keyers and digital devices, obtaining a suitable tiny power transformer may prove difficult and expensive. This fact came to light while I was designing a sixteen IC keyer which was to be no larger than a conventional bug, yet completely self-contained including squeeze paddles, monitor and power supply.

The requirement for a miniature power transformer was met by using an audio output transformer intended for service in pocket transistor radios. A bridge rectifier of small glass diodes across the voice coil winding and a simple resistor-capacitor filter provided a full 3 V at 200 mA. The compact keyer, made possible by the tiny transformer, has been in service for four years without failure.

Gene Brizendine, W4ATE

meter safety

Amateurs often overlook the fact that high-voltage meter multipliers are dependent upon the meter coil for their return to ground. Should this coil open, the entire high voltage will appear at one terminal of the meter! To preclude this risk, a very high value resistor, Rx, is connected across the meter as shown in **fig. 1.** This resistor must be large enough to introduce only a small error in the meter reading, which may be compensated by a

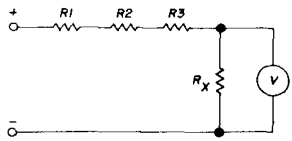

fig. 4. Use of a shunt resistor for safety.

corresponding decrease in the value of R2 to correct the meter.

As a further note of safety, the multipliers, R1 through R3, generally should be 2 W or larger, because a 500 V maximum drop across each resistor is an acceptable value. Otherwise, the resistor drift due to heating may be excessive and the voltage breakdown of the multipliers will be exceeded.

M. H. Gonsior, W6VFR

electronic fence interference

I live in a built-up area which permits keeping horses. I have had very strong noise interference on one bearing on the higher bands, and on all 3.5- and 7-MHz signals when I use a nondirectional antenna. I found the source of the noise very quickly with an inexpensive vhf aircraft-band portable a-m radio.

The interference consists of about one buzz per second, somewhat more regular than thermostatic devices. It is caused by the less expensive *Sears weed and stock-control charger with pulser timer,* which is described more extensively in the *Suburban and Farm Catalog* than in the general catalog.

Bill Nelson, the radio interference expert with The Southern California Edison Company and author of a very good article on the subject several years ago in *QST,* says that some electric fences produce only a series of clicks, while others have the buzzes.

During a 21-MHz contact with VK3AKB, Bill turned on the charger that was located near his ssb equipment and heard no interference at all.

The Sears general catalog did not mention any radio interference filter in the lower-priced charger, but did for the more expensive one. The farm catalog, in addition to mentioning the filter in the $40 unit, says that the $24 unit will not interfere with radio and tv. The noisy one here puts intermittent show on a tv screen near the noise source, and can be heard a hundred yards away on 140 MHz.

Sears has written extensively to assist in eliminating the radio interference and has provided factory comment. They say that both of their chargers are shielded and filtered to prevent radio, tv and telephone interference. When the interference is present, it is usually caused by one or both of the following conditions:

First: A current leak to ground somewhere along the fence line. This could be caused by very dirty or broken insulators, the wire touching against the side of a post, tree or building or heavy vegetation growing up against the wire.

Second: A loose connection somewhere along the fence line. This could be due to poor splices (usually just twisted connections in a rusting wire), a gate opening or very badly rusted spots in the wire.

Either of the two conditions above would cause a gap across which sparks could jump, exciting the fence wire as an antenna, thus resulting in very broad interference to radio, television or telephone. A careful check of the fence line should disclose the trouble — and may require some soldering for permanent connections.

Sears also says that you can determine definitely whether the controller (charger) itself is at fault by disconnecting it

Why won't Don Wallace listen to anyone else?

W6AM doesn't have to, since he began using our professional quality VHF-FM 2 meter transceiver with exclusive *Astropoint*. The precise and powerful Astropoint system actually blocks out all interference. So when one of the nation's number one hams is having a 2-way conversation he won't have unwanted interference.

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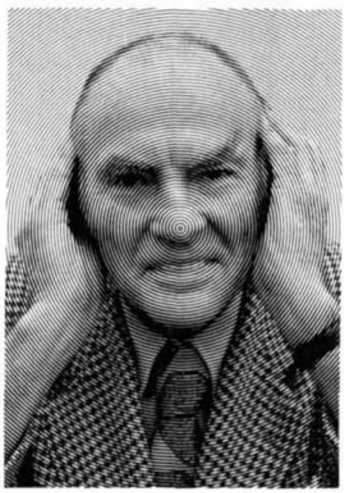
Which is why Don Wallace hasn't been listening to anyone else lately. Not that he's choosy about who he listens to. Just whose equipment he listens on.



For complete specifications and the name of your nearest dealer, write:

STANDARD COMMUNICATIONS CORP.
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(213) 775-6284

from the fence and allowing it to operate. Of course, the removal of the antenna will reduce the interference markedly in any event. However, if a nearby receiver shows that the interference disappears entirely, the fence rather than the charger is at fault.

If the noise continues, the charger unit – possibly due to an open filter condenser – is at fault and may be returned to a service station for repair. First, however, you may try a new pulser in the charger, as this possibly could be the source of the difficulty.

E.H. Conklin, K6KA

low-band vertical antenna

The 4BTV four-band vertical antenna requires frequent readjustment when you are changing frequency and using the 75-meter resonator or loading coil. The same problem is present when using base loading. If a remote motor drive is used, additional cabling is required. There is a solution available that allows the operator to tune the 4BTV over the entire 80- and 75-meter band from his station location and still maintain a reasonable bandwidth. The technique uses a coaxial quarter wave transformer tuned to 3.750 MHz (43.3 feet long) and a broadcast type ganged variable capacitor connected at the remote end as shown in **fig. 1**.

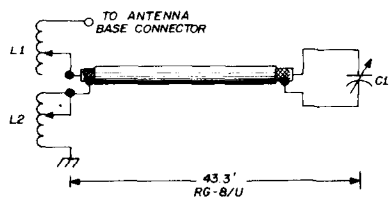
In my case, the tap on L2 for proper impedance matching occurs at the top of the coil. L1 requires exactly half of the 30 available turns when C1 is half meshed. The vswr is 1:1 at the mid-band resonance frequency of 3.750 MHz. The inductive loading for quarter-wave resonance and a near 50-ohm resistive impedance is found by varying the tap on L2 while L1 is varied for quarter-wave resonance (L1 and L2 are mounted with zero mutual coupling).

After L1 and L2 are adjusted for a minimum vswr at 3.750 MHz with C1 at half mesh, then C1 need only be adjusted at any other frequency in the 80-meter band for a vswr better than 1.3:1.

The arrangement and components used have been exposed to a 300-watt power output level. At 500-watts output, the base antenna connection of the 4BTV arced across to ground. Therefore, 30-watts output is used with the 4BTV on 80 meters with this arrangement.

The stub may lie on the ground since it is connected at the low-impedance point (shield of coax to top of L2 and center conductor to bottom of L1). Negligible loss and capacitance-to-ground effects are experienced and the quarter-wave transformer inversion properties present a variable series inductance and capacitance between L1 and L2. The coils, once set at 3.750 MHz, require no further adjustment over the band.

A 4BTV can be replaced by a full quarter-wavelength 40-meter vertical or an all-band vertical with equally good



L1 30 turns, no. 12 wire, 8 tpi, 2.5-inch diameter

L2 6 turns, no. 12 wire, 2.5-inch diameter. Turns spaced approx. 5/8- inch apart

C1 five gang broadcast type, 365 pF per section

fig. 5. Remote antenna tuning arrangement.

performance on the 80- and 75-meter bands.

One word of caution: When using base loading of a quarter wavelength 40-meter vertical antenna on 80 meters, a healthy rf voltage will appear at the base of the antenna, a hazard to unsuspecting people or animals.

The values of L1 and L2 may vary at different installations and must be determined experimentally. Those given in **fig. 1** should be about right for those installations using one ground rod and mounting pipe with the 4BTV.

Fred M. Griffee, W4IYB

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*Above prices do not include shipping costs. Please add 75¢ on parts orders, $2.00 on larger kits. Shipping via UPS whenever possible; therefore, street address required.

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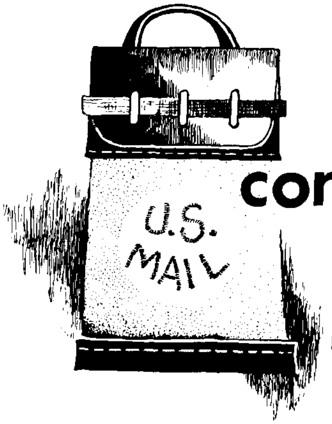

comments

speech clipping

Dear HR:

I write this letter to question the general understanding and use of rf clipping circuits as given in many schematic diagrams; namely, I question the conventional explanation given for clipping at an intermediate frequency such as 9 MHz.

Squires and Bedrosian ("The Computation of Single-Sideband Peak Power," *Proceedings of the IRE,* page 123, January, 1960) have used a Fourier analysis, coupled with the concept of "frequency incommensurability" to show that audio clipping results in an *increase* in ssb peak amplitude rather than the expected reduction. This, of course, arises because a single-sideband signal does not preserve relative phase information; hence, there always exists a point in the ssb envelope where all frequency components add for an instant to produce a peak (the more frequency components present, the higher the peak). Thus, because audio clipping produces additional frequency components, the rf peaks become more intense, contrary to expectations.

However, the mathematics of the subject article also appear to be applicable to a situation where one clipped rf wave modulates another rf sine wave (heterodyning), and one of the sidebands (say, the difference frequency) is rejected. In this case, as in the audio case (which is not really a special one), phase information is lost, and one can use Squires' and Bedrosian's mathematics to show that the final signal should be degraded too. I therefore feel that this is a matter which ought to be clarified, because this particular article has become the *raison d'etre* for rf clipping.

It appears that if Squires' and Bedrosian's analysis is correct, then rf clipping can only be expected to produce optimum results when it is used at the *operating* frequency. If that is true, as my own brief inspection seems to indicate, then someone who has the time to pursue it further should discuss in detail this apparent negation of rf clipping. After all, how do you filter the out-of-passband products at the operating frequency?

Furthermore, if my interpretation of the mathematics is corrent, clipped dsb should be far more effective than unclipped dsb, whereas clipped ssb will supposedly be a disappointment.

Oscillograms appearing in the 1969 "Radio Amateur's Handbook" appear to support the notion that rf clipping at intermediate frequencies works, but one can again take note that "spikes" (many of them) appear in the output of a clipped and heterodyned, ssb waveform, and that they do not follow the i-f envelope faithfully, all in accord with the subject article's predictions. Perhaps clipping at the i-f is effective, but if it is, someone must look more closely for the reasons and, perhaps, determine if Squires and Bedrosian produced an analysis that was too simplified. (It never really concerned itself with speech – merely distorted "sine" waves – and it totally ignored filtering.) As a matter of fact, I suspect that the analysis *is* over-simplified and that amateurs may have succeeded in regions not really covered by theory. Perhaps there are later papers which clarify the situation, and one of your readers who keeps up with IEEE's journals can help out. I hope so, because in my mind the case for conventional rf clipping does not seem to be on solid ground.

Richard R. Slater, W3EJD

Dear HR:

I was pleased to get the reference to the Proceedings of the IRE note by Squires and Bedrosian, since it nicely supports my simplified explanation of why speech clipping and the ssb mode are incompatible. (See my article in HR *for February, 1971) I was not aware of the material though I have the Proceedings issue in my collection!*

Let me start by assuring you that rf (or i-f) clipping is indeed well understood. I believe your difficulty arises from the fact that in the IRE article an infinite bandwidth is assumed. This is why the authors get even worse results for clipped speech in a ssb system than I, because I assumed practical limits for bandwidth (3 kHz) and the lowest audio frequency (400 Hz). An infinite, or to be more practical, a large bandwidth, is a purely relative term and must be viewed against the signal frequency. A 3-kHz bandwidth is indeed nearly infinite when you consider the distortion products or harmonics of a clipped 300-Hz tone. However, a 3-kHz or even a 30-kHz bandwidth is small when you consider a 100-kHz (or 9-MHz) clipped signal.

Since the distortion products or harmonics (multiples of the frequency of the clipped signal) are sufficiently filtered out by a single i-f transformer after the i-f clipper, the problem which is causing your concern does not arise. The subsequent mixers and amplifiers are dealing with an amplitude-limited signal without *the distortion components.*

Some time ago, there was a widespread belief that when you filtered a clipped signal so that all harmonics were removed, you got back the unclipped original. This of course is a fallacy; inspection of the Fourier series for a square wave shows that the output variation is 2 dB for inputs between 1 (the clipping level) and infinity!

Within the context of your letter, clipping at audio is a special case, since the lower harmonics generated by the clipper fall into the band of interest. Clipping at i-f or rf avoids this through the selectivity inherent in most designs so that there is no phase information to lose. (I agree of course that later mixers are in effect ssb mixers.) For example, when clipping a 300-Hz tone, the distortion products are at 900, 1500, 2100, 2700 Hz etc. When you clip a 100-kHz ssb signal, the distortion terms show up at 300 kHz, 500 kHz etc. Obviously, it is no major task to get rid of these, as the band of interest is still only 3 kHz wide.

Walter Schreuer, K1YZW
Ipswich, Massachusetts

mosfets maligned

Dear HR:

An article in the March, 1972 issue, Gerald Vogt's writeup on a two-meter preamplifier, inspires this missive, since I am of the opinion that he leaves some misrepresentations in his article.

I am pleased to see that Mr. Vogt

recognizes the ease with which the classic cascode amplifier circuit can be directly implemented with fet devices. The excellent results that may be obtained with this circuit arrangement without the need for neutralization are recognized advantages of the cascode amplifier. The 2N4416 or other jfets can produce a fine two-meter preamplifier.

However, I do feel that the author has been inclined to dismiss the dual-gate mosfet devices in much too casual a manner, and without a sufficient understanding of their true virtues. The argument that the mos junctions are subject to destruction by electrostatic discharges is archaic in this day of the diode-protected device. Certainly the use of dual-gate mosfets in the Heath SB-303 shows the practicality of the protected device. I am presently involved with a receiver design which is in production, using RCA dual-gate 40822 fets, and we have used some 4000 devices on the assembly floor without a failure due to electrostatic problems. Incidentally, I am using this type fet for rf, mixer, i-f, audio and oscillator stages and find it admirably suited for all applications.

The inherent cascode internal connection of the dual-gate mosfet makes it ideally suited for cascode preamplifiers, and it is regrettable that its superiority was not recognized.

The dual-gate device offers better intermod and crossmod performance than any other device in the solid state museum, and permits the introduction of an rf gain control or agc function that actually *increases* the signal handling capabilities as the gain is reduced. This is even better performance than that offered by the pentode variable-mu vacuum tube!

I would suggest that Mr. Vogt read over some of the informative application notes produced by Motorola and RCA, for example, on the use of dual-gate mosfets in rf applications. Data on the RCA 40673, RCA 40822, or Motorola MFE121, for instance, will reveal that these devices will give noise figures in the 2- to 5-dB range at two meters with stage gains near 20 dB as cascode amplifiers at 150 MHz.

RCA application notes give particular insight into uses of these fets. RCA publication ST-3233 on small-signal rf amplification of mos devices; AN-3435 on cross modulation effects; ST-3486 on receiver applications of dual gate mosfets and AN-4431 on rf applications of the dual-gate mosfet up to 500 MHz should be particularly informative.

Finally, I am surprised that Vogt's circuit does not include any local feedback, such as a source resistor with suitable rf-bypass capacity. Since the parameters of fets vary considerably from device to device, it is very important to incorporate dc feedback to make the circuit less critical of device parameter controls. Since Idss varies from device to device, and Gm does also, it is better to smooth out these variables with feedback than to accept large performance variations in the circuit as devices are changed.

I fear that the circuit given by Mr. Vogt will function best only with low-level signals, since nothing is done to optimize the large-signal capability. Further, an rf gain control can be readily introduced to a dual-gate cascode circuit by controlling the voltage applied to gate number 2. If this gate is biased initially at say 30 to 40 percent of the (drain) supply voltage, rf gain can be easily reduced by merely lowering this gate number 2 voltage, which at the same time actually *increases* the signal-handling ability while reducing the stage gain. Certaintly this is a virtue not to be so lightly dismissed.

The above comments are not intended to be hyper-critical of Mr. Vogt's article, as certainly the use of fets as a cascode rf amplifier is far superior to usual bipolar circuits. However, I do feel that the mosfet, especially in the diode-protected dual-gate form, should not be overlooked as the best of the solid-state devices for small-signal circuit designs.

Maurice P. Johnson, W3TRR
Randallstown, Maryland

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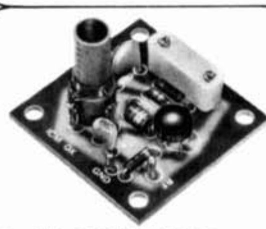

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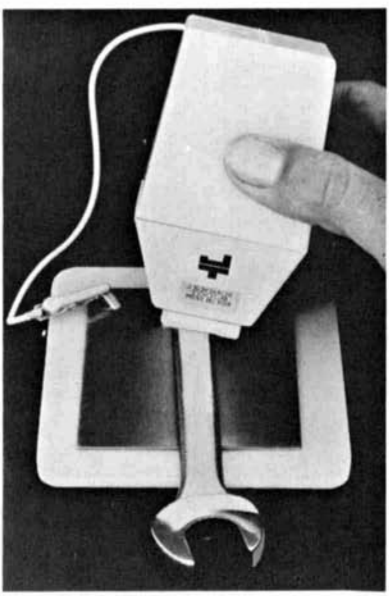
More Details? CHECK—OFF Page 110

new products

metal-etching tool



You can mark tools, metal components, chassis as well as camping gear and household items with a new metal-etching instrument from Jensen Tools. The Model 17 Marking system can electronically etch flat or round metal surfaces and can make two-color metal labels. The tool is safe, silent, vibration free and powered by ordinary 117 Vac.

Simple to use, all that is required is to type, write or draw the mark to be etched on the special stencil included in the kit. A wetting solution is applied to the head of the marker, stencil attached, and the tool is ready for impression making. Chrome plated and solid brass labels with strong adhesive backing are available for etching with a black, copper, or brass-colored mark for a striking two-tone effect.

Model 17 is packaged in an economical kit which includes the marking tool, electrical cord, ground plate, electrolyte solution in plastic bottle, an adapter clip for deep etching and full instructions. The kit includes supplies sufficient for marking up to 2000 items. It is priced at $24.95. Refill supplies are readily available.

Further information is available by using *check-off* on page 110 or by writing to the manufacturer and requesting the descriptive folder on the Model 17 Marking system from Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018.

great circle map

A new edition of the Radio Amateur's Great Circle Chart of the World has just been published by the Radio Amateur Call-book. Measuring 29" by 25", the six-color map is centered on the geographical center of the United States and includes a chart giving great circle bearings to all parts of the world from Boston, Miami, Seattle, Los Angeles, San Francisco and Washington, DC. The chart, an azimuthal equidistant projection, shows the great circle course from its center point to any other point on the earth as a straight line. Additionally, distances along the straight line can be

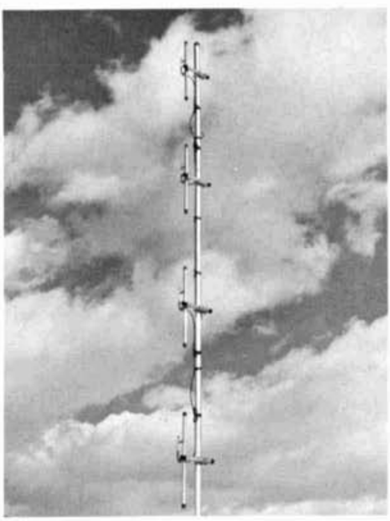
More Details? CHECK—OFF Page 110

measured accurately against one standard scale.

The chart, besides dressing up a radio shack, is helpful for orienting antennas and comparing long and short paths to distant spots. The map amazes visitors to the station as it shows some interesting facts such as the shortest path to Singapore, Vietnam or Burma from Greenville, New Hampshire right over the North Pole.

The map is available for $1.25 from Comtec Books, Greenville, New Hampshire 03048.

four pole antenna



Cush Craft announced the addition of two new models to their Four Pole antenna design. The Four Pole is a series of four stacked dipoles for amateur fm and commercial use.

Four Poles are supplied with the dipoles, mounting booms, harness and all hardware. The center support mast is not supplied, allowing the user to custom select a mast for his installation or to tower mount the antenna.

Gain figures for the antennas show 6-dB omnidirectional and 9-dB semidirectional pattern. The three models now available are the AFM-4D, 144 - 150 MHz, $42.50; the AFM-24D, 220 - 225 MHz, $40.50 and the AFM-44D, 435 - 450 MHz, $38.50.

More Details? CHECK—OFF Page 110

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Write for more information or use READER SERVICE

More information is available from Cush Craft Corporation, 621 Hayward Street, Manchester, New Hampshire 03103 or from *check-off* on page 110.

circuit zaps

Circuit Zaps, copper component patterns and accessories used to produce instant printed-circuit boards, have been developed for the hobbyist. International Rectifier Corporation recently introduced its complete line of Circuit Zaps for custom and prototype production of printed-circuit boards.

Circuit Zaps, which will retail for as low as 9 cents per pattern, enable the hobbyist or design engineer to eliminate the artwork, photography, photoprinting, touchup, etching, stripping and other time consuming and costly steps previously required in prototype and test circuit development and in home electronics projects.

Circuit Zaps are available in four design groups, and each card contains three to twelve of one pattern. Each pattern is precision-etched on a 5-mil glass-epoxy base material and backed with a special pressure-sensitive adhesive.

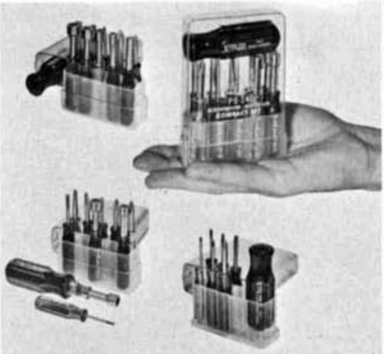
In actual use, the Circuit Zaps are placed on a pre-punched or unpunched circuit board in any desired layout. The Circuit Zaps can be removed and repositioned without damage to the adhesive.

The board is then ready for testing, drilling (for unpunched boards) and positioning of conductor paths. The final step is the mounting of components with standard hand-soldering techniques.

Specific Circuit Zap patterns available include: TO-type patterns with 3, 4, 6, 8, 10 and 12 leads; dual in-line with 14 and 16 leads; resistor/diode types, conductor path and both single-and dual-component pads. Also available from the Semiconductor Division of International Rectifier are: unpunched XXXP laminates, prepunched with 0.1 x 1.0-inch centers and printed-circuit board terminals. Available at your local dealer. For more information use *check-off* on page 110.

More Details? CHECK—OFF Page 110

compact driver sets



Xcelite has just added three new all-purpose screwdriver and nutdriver sets to its family of "compact convertibles." Each set consists of an assortment of color-coded midget tools and a unique "piggyback" torque amplifier handle which enlarges gripping surface, extends reach and increases driving power. The new units bring to nine the number of "compact convertible" sets now available, with various assortments of drivers for slotted, Phillips, Allen, Scrulox, hex and clutch head screws, plus hex nuts.

Featured is a new transparent container with a positive snap-lock. Optically clear for easy set identification, the tough, injection-molded cover stays closed even when tossed into a tool box. The case is also designed to hold tools upright on a bench for easy selection. All nine members of the "compact convertible" family are now housed in these transparent, "show case" cases.

Contents of the three new sets are: PS-6 Screwdriver Set – miniature drivers for No. 00, 0 and 1 Phillips; 3/32", 1/8" and 5/32" slotted screws.

PS-140 Screwdriver and Nutdriver Set – popular assortment includes drivers for No. 0, 1, and 2 Phillips screws; 3/32", 1/8", 3/16" and 1/4" slotted screws; and 1/4", 5/16" and 3/8" hex nuts.

PS-130 Screwdriver and Nutdriver Set – similar to PS-140 with larger assortment of nutdrivers, 3/16", 1/4", 5/16", 11/32" and 3/8" hex sizes; plus drivers for No. 1 and 2 Phillips screws, and 1/8",

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MODEL ACA-1 AUDIO COMPRESSOR features 45 DB compression range ■ Flat 20-20,000 Hz response ■ Extremely low distortion ■ Front panel compression meter and in/out switch ■ Accepts both high and low-impedance mikes ■ Easily installed in mike line ■ 110-volt a.c. or 12-volt d.c. operation ■ Only 5" W x 2¼" H x 4½" D.

MODEL ACP-1
$24.95 KIT
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MODEL ACP-1 COMPRESSOR-PREAMP has 30 DB compression range ■ Flat 20-20,000 Hz response and low distortion ■ Designed for high-impedance mikes ■ Easily installed in mike line ■ 9-volt battery operation ■ Only 4" W x 2½" H x 3½" D.

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Try one of these compressors as an automatic recording-level control. Used by recording studios, schools, and radio-tv stations. Great for p.a. systems, too!

3-CHANNEL WWV RECEIVER
(5, 10, and 15 MHz)
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0.25 microvolt sensitivity ■ Crystal controlled ■ 110-volt a.c. or 12-volt d.c. operation ■ Compact size only 4½" W x 2½" H x 5½" D.

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Now, a Keyer with every feature of the most expensive units without the unnecessary frills. Berkshire Inst. has designed a complete ready to use Keyer on a single printed circuit board. Use your own Key and incorporate it in your station console or power supply. Built-in regulated power supply operates from 6.3 VAC or 7½ VDC and only requires 60 Ma.

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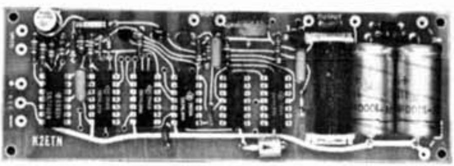
—Dot or Dash Injection.
—Built-in Variable Pitch Side Tone.
—6 to 80 WPM.
—Reed Relay Output for Compatability With any Rig.

$34.95



ALSO AVAILABLE:
K2 12/24 HOUR CLOCK
K2 AGC AMPLIFIER

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3/16" and 1/4" slotted screws. Both this set and PS-140 are expansions of the very popular PS-7.

Complete details are given in the new Catalog 171 available without charge from Xcelite Incorporated, Orchard Park, New York 14127 or by using *check-off* on page 110.

current-controlled resistors

Radiation Devices has introduced two new current-controlled resistor (CCR) pairs. CCR pairs and quads may be used in tuning, switching, attenuating or isolating functions in electronic equipment or systems. They consist of a light-emitting diode optically coupled to a pair or quad of cadmium selenide photo-resistors, selected and adjusted so that their resistance values track over a thousand-to-one range. The control current through the LED required to cover this range varies from a few tens of microamperes at one megohm to 20 milliamperes maximum at one kilohm. Fourteen pin DIP packaging, combined with carefully considered pin assignment, provides convenience in utilization, compatibility with modern circuit-board technique and maximum signal isolation.

The model 500-104 is a resistor pair selling for $18, and the model 500-105 is a resistor quad selling for $24. Complete specifications are available by using *check-off* on page 110 or by writing to Radiation Devices Company, Box 8450, Baltimore, Maryland 21234.

logic tester

Requiring no visual observation, the model 95 logic tester gives a high audio tone to indicate logic 1 state and a low audio tone to indicate a logic 0 state when checking normal 5 V digital logic circuits. The unit sells for $19.95. More information is available from Production Devices, 7857 Raytheon Road, San Diego, California 92111 or by using *check-off* on page 110.

high-current logic driver

High-current loads can now be driven by logic circuitry, using the Motorola MCH2890, dual power driver rather than using more complex discrete circuitry. This new device translates logic voltage levels to high-power outputs. Either DTL or TTL logic levels may be used to control the device, and loads can be either resistive or inductive.

Many applications such as RTTY magnet drivers and tape punches, hammer-drivers, relay drivers, stepping motors and lamp drivers require high current pulses that are digitally controlled. The new device provides this interface in a single package replacing an IC and two Darlington transistor packages.

The MCH 2890 combines a dual 2-input MTTL AND gate similar to the MC3101 and a pair of Darlington power transistors in a hydrid design to provide up to 6 amps at 10% duty cycle and 25 ms pulse width. Continuous output current is 1 ampere maximum. The output Darlington transistors have 120 V minimum breakdown voltage ratings which is desirable for driving inductive loads at high current.

A factor which has hampered IC drivers in the past was package power dissipation. A new 10-pin aluminum package similar to the popular TO-3 was designed for the MCH2890. Besides the power handling capabilities of the TO-3 package, it was also chosen because of its longtime popularity as the standard industrial power package.

For further information contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, P.O. Box 20924, Phoenix, Arizona 85036 or by using *check-off* on page 110.

itu lists available

Formerly available only from Geneva, Switzerland; International Telecommunications Union publications are now avail-

NEW

ILLUMINATED ROCKER SWITCH

American made UL approved 125 Volt 3 A ac Illuminated S.P.D.T. rocker switch. Separate built-in S.P.D.T. switch *operates built-in 115 VAC lamp.* 75¢ ea. 3 for $2.00

CAPACITORS

DIPPED MICA CAPACITORS
ALL 5% 500 VOLTS

12	36	82	240	510
18	39	91	270	560
20	43	100	300	620
22	47	120	330	680
24	56	150	360	
27	62	180	390	
30	68	200	430	
33	75	220	500	

all mmfd

YOUR CHOICE 16 for $1.00 ppd.

RESISTORS

The Following Resistors Are All American Made ½ Watt Carbons With Full Leads, *Completely Standard.* Some 5% Some 10% Some 20%

6.8	1300	24K	1 Meg.
22	1500	33K	1.5 Meg.
27	1800	51K	1.8 Meg.
39	2200	68K	2.2 Meg.
47	2400	75K	2.7 Meg.
82	2700	100K	4.7 Meg.
100	3300	120K	5.6 Meg.
270	4700	180K	9.1 Meg.
330	5600	220K	10 Meg.
390	6800	330K	
470	8200	390K	
620	10K	470K	
750	18K	680K	
1000	22K	910K	

33 for $1.00 ppd.

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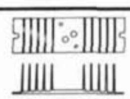
All Units Tested And Guaranteed

500 MILLIWATT UNITS — 3, 3.3, 3.6, 3.9, 4.3, 4.7, 5.6, 6.2, 6.8, 7.5, 8.2, 9.1, 18, 22, 24, 27, Volts.

1 WATT UNITS — 10, 11, 12, 13, 15, 16, 18, 20, 22, 27, 30, 33, 39, 43, 47, 51, 56, 62, 68, 75, 82, 91, 100, 110, 120, 130, 150, 160, 180, Volts. ALL UNITS 10% — 4 for $1.00 ppd. 5% — 3 for $1.00 ppd.



Part #7277151. Black, Aluminum, Anodized, Heat Sink. Size Approximately 4¾" x 1⅜" h. x 1½" w. Predrilled For TO-3 Transistors. Delco Factory New. 75¢ Each 3 for $2.00 ppd.

Rectangular 0 to 1 Mil. Meter. *Panel Size Approximately* 1⅜" x ¾" x 1¼" Deep. Basic Movement 0-1 ma. Easily Adjusted For 0 Center With Instructions supplied. Ideal meter For RTTY Use, Etc. American Made. Price: $1.00 Each ppd.

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CONVERTER

Model 407

$42.95

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144-146 MHz in. 28-30 MHz out or 146-148 MHz with a second crystal available for $5.95 each

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

Can't wait? Then send us a postal money order for $42.95 and we'll rush the 407 out to you. NOTE: The Model 407 is also available in any frequency combination up to 450 MHz (some at higher prices) as listed in our catalog. New York City and State residents add local sales tax.

VANGUARD LABS

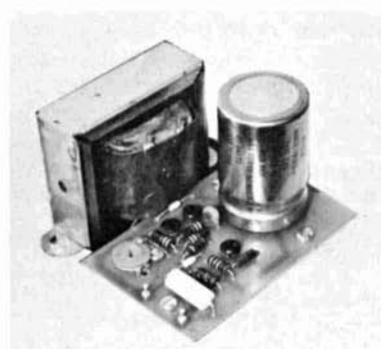
Dept. R, 196-23 Jamaica Ave., Hollis, N.Y. 11423

able in this country from Gilfer Associates. The ITU, world-wide treaty registration center for all radio stations and frequencies, maintains computerized lists of the hundreds of thousands of radio stations. The lists are updated quarterly.

Typical lists and prices are: "List of Ship Stations" ($5.25); "List of Coast Stations" ($6.60); "List of Fixed Stations ($22.00); "List of Broadcasting Stations Operating Below 5950 kHz" ($11.50) and "Alphabetical List of Call Signs" ($7.70). Gilfer will also accept subscriptions to the ITU monthly magazine, "Telecommunications Journal."

A list of ITU publications and prices is available from *check-off* on page 110 or directly from Gilfer Associates, Inc., Box 239, Park Ridge, New Jersey 07656.

ic power supplies



Viking Electronics has introduced a line of low-cost power supplies for logic and linear-system applications. Priced from $17 to $24 in single quantities, the OEM 70 series provides typical outputs of 3.5 to 6 V at 3 A and 8 to 15 V at 1.2 A with regulation of 0.5 to 0.1 percent and ripple of 1 to 2 mV, dependent on models.

Features include electronic current limiting, floating output, stable differential amplifier circuitry, silicon transistors and computer grade capacitors. For more information contact Viking Electronics, Inc., 721 Saint Croix, Hudson, Wisconsin 54016 or use *check-off* on page 110.

More Details? CHECK—OFF Page 110

BROADBAND AMPLIFIERS at a price you can afford

The BBA-1 series offers high gain over the *HF* to *UHF* region with low noise figures. BBA-1PA, a remotely powered unit for mounting near the antenna, covers 30 to 400 MHz with a 20 ± 1dB gain dropping to 17 dB at 500 MHz. Maximum NF is 3.5dB. BBA-1PB has the same specs as *the BBA-1PA but is locally* powered. BBA-1P covers 3 to 500 MHz. with gain between 15 and 30dB. Maximum NF is 3.0dB. For less critical use order BBA-1 with same coverage as the BBA-1P, but reduced UHF gain and higher NF.

BBA-1PA - $50.00; Adapter Kit, KT-1 - $5.00
BBA-1PB, BBA-1P - $45.00; BBA-1 - $30.00

Please include $1.00 for shipping.

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SUNDAY, AUGUST 13, 1972

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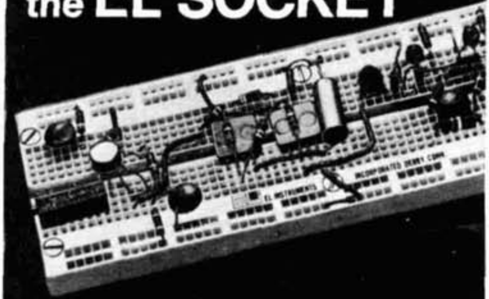

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- ***Nickel/silver plated terminals — very low contact resistance***
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Join the action on 220MHz with the **JANEL** deluxe receiving converters and preamps. Extremely sensitive with freedom from spurious responses.

220CA Converter, 28-32 MHz i-f, **$69.95 ppd**
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Talk-in station on 50.2, 145.71 and 146.94 FM • Tables and Cooking area • Come early — Stay late!

ADULTS — $1.00 IN ADVANCE
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CHILDREN UNDER 12 — FREE

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

It includes numerous technical and construction articles in addition to a complete rundown on the month's events in amateur radio. Surely a most interesting addition to your amateur radio activities.

We can now offer this fine magazine to you along with the other advantages of membership in the RSGB (such as use of their outgoing QSL Bureau) for $9.95 a year.

comtec

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More Details? CHECK—OFF Page 110

the most powerful antennas under the sun!

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80 through 10 meters

hy-gain 18AVT/WB

New, from the inventors of wideband verticals.

Pack some punch! All the omnidirectional performance of Hy-Gain's famous 14AVQ/WB...plus 80 meter capability! Unrivaled performance, rugged extra heavy duty construction, and the price you want...*all in one powerful package!*

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- Hy-Q traps effectively isolate antenna sections for full 1/4 wave resonance on all bands.
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- Very low RF absorption from insulating materials.

The 18AVT/WB is constructed of extra heavy duty, taper swaged, seamless aircraft aluminum with full circumference, corrosion resistant compression clamps at all tubing joints. This antenna is so rigid, so rugged...that its full 25' height may be mounted using only a 12" double grip mast bracket...no guy wires, no extra support...the 18AVT/WB just stands up and dishes it out!

Order No. 386 **$59.95**

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More Details? CHECK—OFF Page 110

(continued from page 4)

amateur gear, plus all the possible mixes of sums and differences, you can appreciate the magnitude of the problem.

Recently, a well known fm'er prevailed upon an airline to test his Motorola HT in one of their aircraft so he could operate during a flight he planned to take. After months of correspondence and personal appearances at airline headquarters, and meetings with communications department people, many of whom are amateurs, the airline agreed to run the necessary tests.

The fm'er tweeked and peaked his trusty HT and checked it thoroughly with a spectrum analyzer; it was clean. On the appointed day the aircraft was removed from line operation and the test began. The test required three hours and four men to complete. The HT caused absolutely no interference, and the fm'er received a letter authorizing the operation of *that* HT on *that* particular trip in only *that* type of aircraft. It is easy to understand why the airlines, who are trying to cut costs, would prefer not to get involved in testing each amateur's fm rig.

Unfortunately for the fm'er who went to all this trouble, the aircraft on which the tests were conducted is to be phased out of operation soon, and he's right back with the rest of us – speechless while aboard an airliner.

Many fm'ers continue to ask the Captain's permission to operate, and he may give it, not realizing the position he is putting himself in. He could have his license suspended or he could be fined. Don't put him in that position, and don't subject yourself and other passengers to a situation which could be hazardous to all on board, and perhaps to someone on the ground.

Remember, you may not cause any interference all across the country, but the ILS glide slope receiver is used only during the last few minutes of flight, and interference to these units may not be noticed until it is too late.

Jim Fisk, W1DTY
editor

SBE

feature after feature after feature

There's a new enjoyment horizon waiting on 420-450MHz, an amateur band 7½ times wider than 144-148MHz! Here's wide open space for the ragchewer — paradise for the antenna experimenter. With half-wave antennas only about a foot in length, "dream beams" — the kind that literally **drip** with gain — become practical realities.

The highly knowledgeable repeater group has, of course, had equipment operating in this band for years. However, SB-450, SBE's exclusive, all-new FM transceiver gives the amateur something he has long needed, equipment with small size, convenience and performance characteristics considered to be indigenous to the "DC" bands. It is certain that repeater and 450MHz band activity will be given a big boost with this fine unit.

Beautifully constructed, book size, the SB-450 has 12 channel capability, delivers 5 watts FM output. It's all solid-state — no warmup — low drain from 12V car battery (operates on 115VAC for base station use with available accessory supply). Two sets of crystals are supplied for repeater and simplex operation.

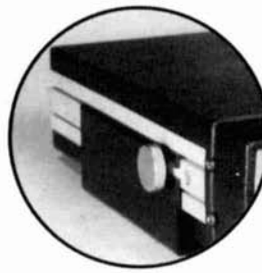
Try UHF! Try SB-450! Enjoy!

SB-450

UHF/FM TRANSCEIVER

All solid state, 1-SCR, 4-IC's, 29 transistors

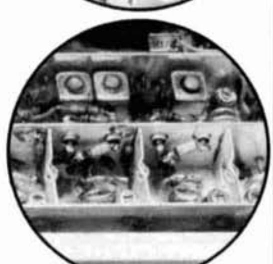
Panel meter shows relative response on receive — power output on transmit



Housing slides in and out of fixed bracket, taper locks with knurledge thumb screws.

Two sets of crystals are supplied. Transmit and receive crystals have individual trimmers for exact frequency centering

Uses exclusive SBE *Super Shape* crystal lattice BP filter



Silver plated brass shielded construction.

SBE

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220 Airport Blvd.,
Watsonville, CA
95076

More Details? CHECK–OFF Page 110

LOOK

The CRICKET is here —
and it has the jump on all the others in its field!

This low cost keyer has more features for your dollar than all others in its price range. Even some in a higher price range! Go ahead — look around — compare the rest! Then buy the BEST — Data Engineering's **new** CRICKET I.

- **Jam-proof spacing**
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- **Built-in key**
- **Relay keys 300 V at 1 amp**
- **Keyed time base. Instant start**
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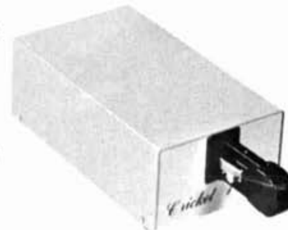

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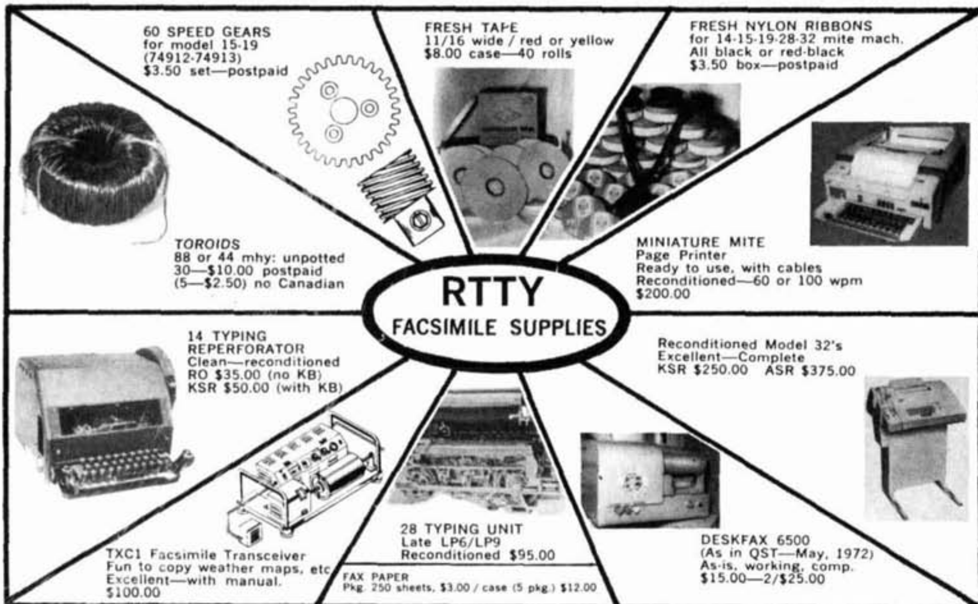

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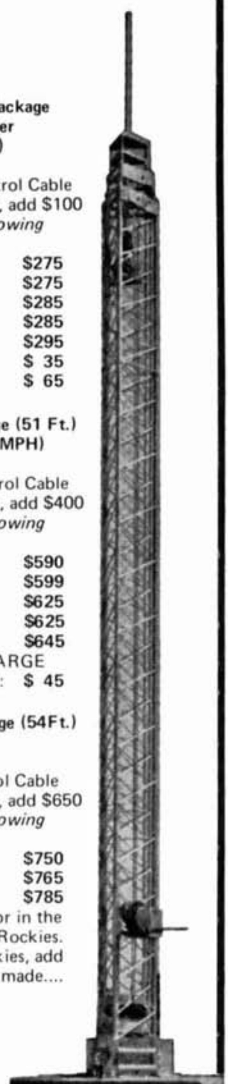

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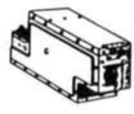

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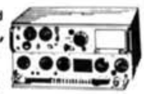

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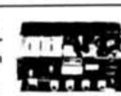

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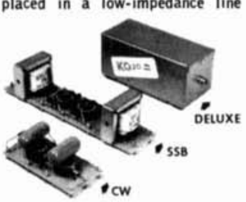

The CW filter has a spot frequency of 780 Hz and a passband of 1100 Hz with a reference level, 40 decibels below the signal level at the design frequency. The peak of the passband is 100 Hz wide at the —3 decibel reference points. The CW filter is specifically designed for low-impedance input and high-impedance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

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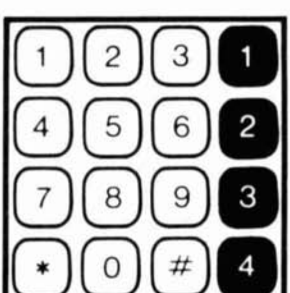

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☐ 500	☐ 5.0K	☐ 20.K	☐ 100.K

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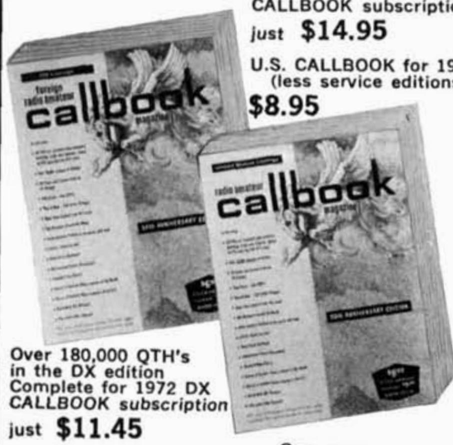

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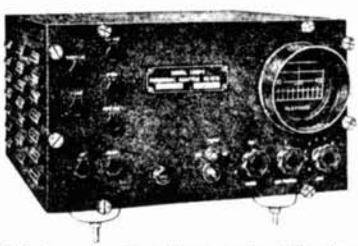

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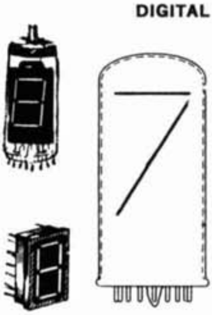

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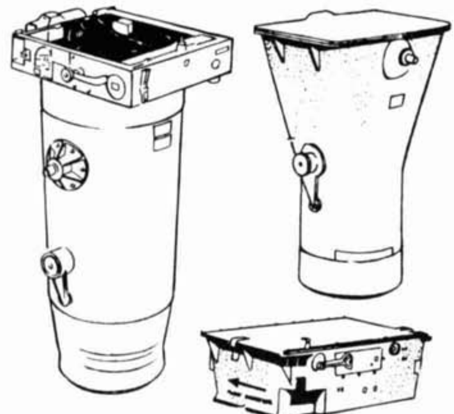

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AUDIO FILTERS: Knock down that background noise. KOJO SSB, AM and CW filters do the job. Write for free brochure and see how serious DX boys hear them. KOJO, Box 7774, 741 E. Highland Ave., Phoenix, Arizona 85011.

FOR SALE: Delco car radios, 12 volts, transistor. Speaker, knobs included. New $14.50 postpaid. W5RSN, 1826 Cypress, Irving, Texas 75061.

THE TEXAS VHF-FM SOCIETY *will hold its annual* Summer Convention August 11, 12, and 13 at the Villa Capri Motor Hotel in Austin, Texas. Technical sessions, manufacturers' displays, door prizes, ladies activities. Call-in will be through the Austin 34-94 repeater. For more inforamtion write Larry Higgins, W5QMU, 2522 Old Hickory Trail, San Antonio, Texas 78230, or Gene Chapline, K5YFL, 2206 La Casa, Austin, Texas 78704.

WANTED — M15 and M32 TELEPRINTERS in any quantity in good condition for use by deaf people . . . will accept donations or for fair prices . . . can be picked up anywhere . . . write Lee Brody, New York-New Jersey Phone-TTY for the Deaf, 15-06 Radburn Rd., Fair Lawn, New Jersey 07410 or call 201-796-5414 evenings.

WANTED: APRIL 1970 HAM RADIO any reasonable price. Tom Morrison, WA3GBU, P. O. Box 13442, Austin, TX 78711.

DISCOUNTS! Standard, Sonar, Clegg, Robyn, Mosley, Cush Craft, Others. Also Marine Gear. Write stating needs. Arena Communications, Dept. H, 1169 N. Military Hwy., Norfolk, Va. 23502.

WARREN, OHIO ARA'S *Family Hamfest. Aug. 20.* Giant fleamart, swimming, picnicking, all free. Displays, mobile checkin. Camping available. Yankee Lake, Rt. 7 near I-80. Details: QSL W8VTD.

POPULAR ELECTRONICS. Complete file. 35 volumes like new. Excellent for library or radio club. Hundreds building projects, etc. $225.00 cashier's check. Postpaid in USA. Dean Brooks, 5141 Lincoln Avenue, Los Angeles, California 90042.

WANTED SWAN 420 VFO in good working condition. Art Orav, 930 Hastings St., Baldwin N. Y. 11510.

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LOWEST PRICES: ON BRAND NEW FULLY TESTED & GUARANTEED IC'S

BEST SERVICE: 10% DISCOUNT ON ALL ITEMS NOT SHIPPED IN 24 HOURS

MOST CONVENIENT: ORDER DESK 1-800-325-2595 (TOLL FREE)

PLEASE NOTE: To qualify for prices in the last three columns, you must order in EXACT Multiples of 10 per item for all IC items on your order.

	ANY QUANTITY PER ITEM(MIX)			MULTIPLES OF 10 PER ITEM(MIX)		
Catalog Number	1-99	100-999	1000 up	100-999	1000-9990	10000 up
7400	.26	.25	.23	.22	.21	.20
7401	.26	.25	.23	.22	.21	.20
7402	.26	.25	.23	.22	.21	.20
7403	.26	.25	.23	.22	.21	.20
7404	.28	.27	.25	.24	.22	.21
7405	.28	.27	.25	.24	.22	.21
7406	.52	.50	.47	.44	.42	.39
7407	.52	.50	.47	.44	.42	.39
7408	.32	.30	.29	.27	.26	.24
7409	.32	.30	.29	.27	.26	.24
7410	.26	.25	.23	.22	.21	.20
7411	.28	.27	.25	.24	.22	.21
7413	.58	.55	.52	.49	.46	.44
7416	.52	.50	.47	.44	.42	.39
7417	.52	.50	.47	.44	.42	.39
7420	.26	.25	.23	.22	.21	.20
7421	.26	.25	.23	.22	.21	.20
7426	.34	.32	.31	.29	.27	.26
7430	.26	.25	.23	.22	.21	.20
7437	.56	.53	.50	.48	.45	.42
7438	.56	.53	.50	.48	.45	.42
7440	.26	.25	.23	.22	.21	.20
7441	1.73	1.64	1.55	1.46	1.37	1.27
7442	1.27	1.21	1.14	1.07	1.01	.94
7443	1.27	1.21	1.14	1.07	1.01	.94
7444	1.27	1.21	1.14	1.07	1.01	.94
7445	1.71	1.62	1.53	1.44	1.35	1.26
7446	1.24	1.17	1.11	1.04	.98	.91
7447	1.16	1.10	1.04	.98	.92	.85
7448	1.44	1.37	1.29	1.22	1.14	1.06
7450	.26	.25	.23	.22	.21	.20
7451	.26	.25	.23	.22	.21	.20
7453	.26	.25	.23	.22	.21	.20
7454	.26	.25	.23	.22	.21	.20
7460	.26	.25	.23	.22	.21	.20
7470	.42	.40	.38	.36	.34	.32
7472	.38	.36	.34	.32	.30	.29
7473	.50	.48	.45	.43	.40	.38
7474	.50	.48	.45	.43	.40	.38
7475	.80	.76	.72	.68	.64	.60
7476	.56	.53	.50	.48	.45	.42
7480	.76	.72	.68	.65	.61	.57
7483	1.63	1.55	1.46	1.38	1.29	1.20
7486	.58	.55	.52	.49	.46	.44
7489	4.25	4.00	3.75	3.50	3.25	3.00
7490	.80	.76	.72	.68	.64	.60
7491	1.43	1.35	1.28	1.20	1.13	1.05
7492	.80	.76	.72	.68	.64	.60
7493	.80	.76	.72	.68	.64	.60
7494	1.18	1.12	1.05	.99	.93	.87
7495	1.18	1.12	1.05	.99	.93	.87
7496	1.18	1.12	1.05	.99	.93	.87
74100	1.52	1.44	1.36	1.28	1.20	1.12
74107	.52	.49	.47	.44	.42	.39
74121	.56	.53	.50	.48	.45	.42

	ANY QUANTITY PER ITEM(MIX)			MULTIPLES OF 10 PER ITEM(MIX)		
Catalog Number	1-99	100-999	1000 up	100-999	1000-9990	10000 up
74122	.70	.67	.63	.60	.56	.53
74123	1.21	1.06	1.00	.94	.89	.83
74141	1.63	1.55	1.46	1.38	1.29	1.20
74145	1.41	1.33	1.26	1.18	1.11	1.04
74150	1.63	1.55	1.46	1.38	1.29	1.20
74151	1.20	1.13	1.07	1.01	.95	.88
74153	1.63	1.55	1.46	1.38	1.29	1.20
74154	2.43	2.30	2.16	2.03	1.89	1.08
74155	1.46	1.39	1.31	1.23	1.16	1.08
74156	1.46	1.39	1.31	1.23	1.16	1.08
74157	1.56	1.48	1.39	1.31	1.23	1.15
74158	1.56	1.48	1.39	1.31	1.23	1.15
74160	1.89	1.79	1.68	1.58	1.47	1.37
74161	1.89	1.79	1.68	1.58	1.47	1.37
74162	1.89	1.79	1.68	1.58	1.47	1.37
74163	1.89	1.79	1.68	1.58	1.47	1.37
74180	1.20	1.13	1.07	1.01	.95	.88
74181	5.20	4.90	4.59	4.26	3.98	3.67
74182	1.20	1.13	1.07	1.01	.95	.88
74192	1.98	1.87	1.76	1.65	1.54	1.43
74193	1.98	1.87	1.76	1.65	1.54	1.43
74198	2.81	2.65	2.50	2.34	2.18	2.03
74199	2.81	2.65	2.50	2.34	2.18	2.03
NE501	2.99	2.82	2.66	2.49	2.32	2.16
NE531	3.81	3.58	3.36	3.14	2.91	2.69
NE533	3.81	3.58	3.36	3.14	2.91	2.69
NE536	7.31	6.88	6.45	6.02	5.59	5.16
NE540	2.16	2.04	1.92	1.80	1.68	1.56
NE550	1.24	1.17	1.11	1.04	.98	.91
NE560	3.57	3.36	3.15	2.94	2.73	2.52
NE561	3.57	3.36	3.15	2.94	2.73	2.52
NE562	3.57	3.36	3.15	2.94	2.73	2.52
NE565	3.57	3.36	3.15	2.94	2.73	2.52
NE566	3.57	3.36	3.15	2.94	2.73	2.52
NE567	3.57	3.36	3.15	2.94	2.73	2.52
N5111	.90	.86	.81	.77	.72	.68
N5556	1.87	1.77	1.66	1.56	1.46	1.35
N5558	.80	.76	.72	.68	.64	.60
N5595	3.40	3.20	3.00	2.80	2.60	2.40
N5596	1.87	1.77	1.66	1.56	1.46	1.35
709	.42	.40	.38	.36	.34	.32
710	.42	.40	.38	.36	.34	.32
711	.44	.42	.40	.37	.35	.33
723	1.00	.95	.90	.85	.80	.75
741	.44	.42	.40	.37	.35	.33
748	.48	.46	.43	.41	.38	.36
1N270	.15	.14	.13	.12	.11	.10
1N751A	.30	.28	.26	.24	.22	.20
1N914	.10	.09	.08	.07	.06	.05
1N4002	.15	.14	.13	.12	.11	.10
1N4154	.15	.14	.13	.12	.11	.10
2N3860	.25	.23	.21	.19	.17	.15

All IC's are supplied in 8-, 14-, 16-, or 24-pin DIP (Dual-in-Line) plastic package, except for NE536, NE537, NE540, and SE540 which come in TO-5 package.
We give FREE data sheets upon request, so ask for those data sheets that you NEED, even for those listed IC's that you are not buying. On orders over $25.00 we'll send you a new 270-page COMPLETE TTL IC data book FREE. Or, you may obtain a new 240-page LINEAR data book instead. Orders over $50.00 will receive both books. Orders over $100.00 will receive a complete LIBRARY of DIGITAL & LINEAR data & application books totaling 1000 pages FREE. PLEASE NOTE: Data books are shipped separate from your order. Please allow two weeks for delivery.

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$4.95 Each

50-99 $4.75
100-999 $4.50
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Large ¼" 7-segment LED readout similar to the popular MAN-1 but with improved brightness. Has left-hand decimal point. Fits in a DIP socket. Expected life: Over 100 Yrs. Regularly $12.95 in single lots. These are BRAND NEW with full data sheet and 4-page MULTIPLEXING Application Note. Needs a 7447 for driver and ONE CURRENT-LIMITING RESISTOR PER SEGMENT. We can supply you with one or ten thousand FROM STOCK. Also available, ±1 OVERFLOW digit at the same prices. Mixing of Regular & Overflow digit allowed.

Package of 8, 47Ω, ¼W limiting R's.....30¢

Incandescent Type of 7-segment display. With right-hand decimal point. Rated 8mA per segment at TTL supply of 5V. Design life of 50,000 hours. Needs a 7447 as a driver. In DIP Package. Each $3.25

MOLEX IC SOCKET PINS: Use these economical pins instead of soldering your IC's to PC boards. Sold in continuous strips in multiples of 100 pins only.

100 for $1.00; 200 for $1.80; 300 for $2.60
400 for $3.40; 500 for $4.20; 600 for $5.00
700 for $5.80; 800 for $6.60; 900 for $7.40
1000 for $8.20. Each additional 1000 $7.50

ALLEN-BRADLEY MIL-GRADE (5-BAND) RESISTORS. Any of the 84 STANDARD 10% values from 2.7Ω to 22MΩ. ¼ or ½ WATT. EACH...............5¢

CERAMIC DISC CAPACITORS. Type 5GA-1000WVDC. 5, 7.5, 10, 12, 15, 20, 22, 25, 27, 30, 33, 39, 50, 56, 68, 75, 82, 100, 120, 150, 180, 200, 220, 250, 270, 300, 330, 360, 390, 470, 500, 560, 680, 750, 820, 1000, 1200, 1500, 1800, 2000, 2200, 2500, 2700, 3000, 3300, 3900, 4700, 5000μμF. EACH.........10¢
0.01μF. Each.....11¢ 0.02μF. EACH......12¢

LOW VOLTAGE DISCS, Type UK.
1.0 μF, 3V.......25¢ 2.2 μF, 3V.......30¢
0.1 μF, 10V......12¢ 0.2 μF, 10V......20¢
0.47μF, 3V.......25¢ 0.01μF, 16V......10¢

ELECTROLYTIC CAPACITORS:
All values are available in both, axial or upright (PC Board) mount. PLEASE INDICATE YOUR CHOICE.
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500 μF, 25V......30¢ 1000 μF, 35V.....50¢

VOLTAGE REGULATORS. Internally-set, overload and short-circuit proof regulators need no external components to set. With data sheet and application notes. TO-3 Package.
LM-335, 5V, 600mA.....................$2.85
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STANCOR P-8180, 25.2 VCT, 1-Amp Transformer. Ideal for use with LM-series. Each....$3.00

Universal Decade Counting Units

1. Easy to read single plane LED or Filament-type Readout with wide angle viewing.
2. BCD Outputs available
3. Readout Lamp Test
4. Selectable Ripple Blanking
5. Selectable Decimal Point
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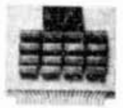
8. Molex Sockets for all IC's & Readout
9. Tin-Plated G-10, 2-oz Copper, Glass Epoxy Board
10. Your choice of 1 to 6 decades on one P.C. board.
11. Grouping of 2 or more readouts on the same card, for custom designed appearance.
12. 5-Volt TTL compatable single supply.
13. First in a series of universal plug-in modules to be introduced soon: for frequency counting, time measurements, event counting, DVM, magnitude comparison, etc.
14. Well documented application note with step-by-step assembly and hook-up instructions.



Price (Per Decade):

Basic Unit consisting of 7490, 7447, Filament-type 7-segment Readout....$8.00

Options:
For LED Readouts instead of Filament-type add $2.00
For 7475 Latch add...$1.25
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For 74192 instead of 7490 add............$1.00
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WRITE OR CIRCLE READER SERVICE CARD FOR OUR CATALOG OF PARTS & SERVICES. IT'S FREE.

More Details? CHECK—OFF Page 110

TECH MANUALS for Govt. surplus gear, only $6.50 each: R-388/URR, R-390/URR, R-390A/URR, R-220/URR, R-274/FRR, TT-63A/FGC, TS-479B/URR, OS-8C/U. Hundreds more. Send 50¢ (coin) for 20-page list. W3IHD, 4905 Roanne Drive, Washington, DC 20021.

2 METER FM, Brand New, Inoue IC-20, 1&10 watts, 12 channels, w/mike, cable, mobile mount $259.50, Bob Brunkow, 15112 S.E. 44th, Bellevue, Wash. 98006. Phone 206-747-8421.

FM HAMFEST — Sunday August 6, Steuben Co. 4-H Fairgrounds near Angola, Indiana. Large flea market, gate prizes — 1st prize an HT220. Picnic grounds, campsites, boating, swimming, food, soft drinks available. Entertainment for XYL. No vendors fee. For more information write FWRA, Box 6022, Fort Wayne, Indiana 46806.

MORSAVERTER READS HAND-SENT MORSE CODE off the air and coverts it to RTTY code. Connects to station receiver and Teletype. No speed adjustments required 5-40 WPM. Supplies CR, LF, LTRS, FIGS automatically. All on one 8x10 inch circuit board. Kits from $185, wired and tested $560. Write for information. Petit Logic Systems, 908 Washington, Wenatchee, Wash. 98801.

WE BUY ELECTRON TUBES, diodes, transistors, integrated circuits, semiconductors. ASTRAL ELECTRONICS, 150 Miller Street, Elizabeth, N. J. 07207. (201) 354-2420.

THE 15th ANNUAL HAMFEST of the Six Meter Club of Chicago Inc. will be held Sunday, August 6, 1972 at the Picnic Grove on U.S. 45, 1 mile north of U.S. 30, 5 miles South of U.S. 6, Frankfort, Illionis. Food and drinks will be available. Swap and Shop section provided. Advance registration $1.50, admission at the gate $2.00. For tickets and further information contact Val Hellwig, K9ZWV, 3420 South 60th Court, Cicero, Ill. 60650. Talk-in frequencies will be on 146.94 MHz FM.

SELL: HEATH HW-32A / AC & DC power supply, EV 664 desk mike, Shure 414 mobile mike, HM-11 SWR bridge. No reasonable offer refused. C. Novak, 110-48 72 Ave., Forest Hills, N. Y. 11375.

STOP — before you buy FM equipment. Check prices of L. M. Communications, 516 Chapman Parkway, Hamburg, New York 14075.

SURPLUS MILITARY RADIOS, Electronics, Radar Parts, tons of material for the ham, free catalogue available. Sabre Industries, 1370 Sargent Avenue, Winnipeg 21, Manitoba, Canada.

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

NU SIGMA ALPHA International Amateur Radio Fraternity. Memberships available. Handsome Certificate, I.D. card, Newsletter, contests, nets, and more. Send for our free illustrated brochure. Box 310, Dept. R, Boston, Mass. 02101.

TOROIDS, iron "E" powder 80-10 meters. .500" 8/$1.00, .940"-4/$1.00, 1.437"-75¢ each or 3/$2.00, 2.310"-$1.50 or 3/$4.00. Please include 50¢ postage, slightly more on larger orders. Fred Barken, WA2BLE, 274 E. Mt. Pleasant Ave., Livingston, N. J. 07039.

TONE BURST . . . Inoue Regency owners, now available, 4 frequency, internally mounted tone burst oscillator . . . $29.50. NHE Communications, 15112 S.E. 44th Bellevue Wash. 98006. Phone 206 747-8421.

P.C. BOARD NEGATIVES made, up to 6" x 6" $2.25 ppd. Same day service. VAF Photo, 4124 Colebrook Rd., Charlotte, N .C. 28215.

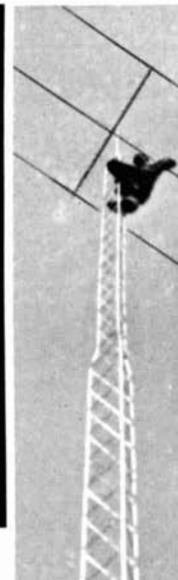
1000 PIV AT 1 AMP new epoxy diodes 10 for $2.50, 1000 PIV at 2½ Amp new epoxy diodes 10 for $4.50, both include disc bypass and bridging resistors. 490MF at 500 volt electrolytic caps 10 for $17.50. All above postpaid USA. East Coast Electronics, 123 St. Boniface Rd., Cheektowaga, New York 14225.

VHF NOISE BLANKER — See Westcom ad in Dec. '70 and Mar. '71 Ham Radio.

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

The "STANDARD"

72' ALUMINUM Tower FOR DX'ing



Strong, Light, No Upkeep

Self-Supporting

Easy to Assemble & Erect

All towers mounted on hinged bases

Complete Telescoping and Fold-Over Series available

*so light you can put it up all by yourself! No climbing, no jin poles, no heart attacks.

And now, with motorized options, you can crank it up or down, or fold it over, from the operating position in the house.

See your local distributor, or write for 12 page brochure giving dozens of combinations of height, weight and wind load.

HEIGHTS MANUFACTURING CO.

NOW In New Larger Facilities
In Almont Heights Industrial Park
Almont, Michigan 48003

CRYSTAL FILTERS
and
DISCRIMINATORS

1 27/64" x 1 3/64" x 3/4"

by

K.V.G.

9.0 MHz FILTERS			
XF9-A	2.5kHz	SSB TX	$23.12
XF9-B	2.4kHz	SSB RX	$32.85
XF9-C	3.75kHz	AM	$35.40
XF9-D	5.0kHz	AM	$35.40
XF9-E	12.0kHz	NBFM	$35.40
XF9-M	0.5kHz	CW	$24.49
XL10-M	0.5kHz	CW	10 pole Gaussian $65.64

9.0 MHz DISCRIMINATORS			
XD9-01	5kHz	RTTY	$17.95
XD9-02	10kHz	NBFM	$17.95
XD9-03	12kHz	NBFM	$17.95

Matching HC25/U crystals XF900 9000.0, XF901 8998.5, XF902 9001.5, XF903 8999.0 (CW) all $2.75 ea.

VHF CONVERTERS UHF

	MM 50	MM 144	MM 220	MM 432	MM 1296
RF Freq. (MHz) †	50-52	144-146	220-222	432-434	
N.F. (typical)	2.5dB	2.8dB	3.4dB	3.8dB	COMING
Nom. Gain	30dB	30dB	30dB	30dB	SOON
	$49.95	$49.95	$54.95	$59.95	

Standard I.F. 28-30MHz †
Power 12v DC
1¼" x 2⅜" x 4⅜" + connectors.
† Other ranges available on request

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ADD A LINEAR* TO MY 2M FM RIG? WHY NOT?

YOU DID IT ON 80-40-20-15-& 10!

And you did it for a good reason!

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A **DYCOMM BOOSTER*** can do for you what it has for over 3,000 amateurs:

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- Is fully guaranteed to perform with an outstanding warranty program.

A model for virtually EVERY 2M rig!

Model Line Up:

10-0	10 W input	100 W output	$195.00
D	10 W input	40 W output	99.95
ES	1 W input	40 W output	99.95

Write for a brochure or call your nearest amateur dealer — you hear!

*A common term for add-on amplifiers.

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More Details? CHECK—OFF Page 110

NEW ELECTRONIC PARTS. Buy-Sell. Free Flyer. Large catalog $1.00 deposit. Bigelow Electronics, Dept. HR, Bluffton, Ohio 45817.

QSLS. Second to none. Same day service. Samples 25¢. Ray, K7HLR, Box 331, Clearfield, Utah 84015.

MINI-MITTER II TRANSCEIVER 40 m. S.S.B. walkie talkie with accessory antenna. Brand new — ready & factory checked — less batteries — $140. Certified check or money order — R. J. Best, 2802 S. Staples, Corpus Christi, Texas 78404.

WORLD QSL — See ad page 109.

WANTED: tubes, transistors, equipment what have you? Bernard Goldstein, W2MNP, Box 257, Canal Station, New York, N. Y. 10013.

COPY MORSE CODE automatically, (Ham Radio November 1971) detailed construction plans $14.95. VMG Electronics, 2138 West Sunnyside, Phoenix, Arizona 85029.

VIKING-NAVIGATOR FOR SALE $50. Will ship. Excellent condx. Pete Burbank, 409 Granite Circle, Lexington, Kentucky 40503.

SIERRA HAMFEST on Saturday, August 5 in California Building in Idlewild Park, Reno, Nev. Write Sierra Hamfest, 1047 Mark Way, Carson City, Nevada 8 701.

THE ZERO-BEATERS AMATEUR RADIO CLUB will hold their annual Hamfest at Washington, Mo. City Park Sunday, August 6, 1972. $600.00 in door prizes, free ham gear auction, free bingo for XYL's free candy scramble for kids.

OMEGA-T NOISE BRIDGE HEADQUARTERS. 1-100 MHz $29.95. 1-300 MHz $39.95 ppd. Miss. add 5% tax. MPQ Industries, P. O. Box 217, Hernando, Miss. 38632.

HOT AS A FIRECRACKER ! That's the upcoming ARRL Hudson Division Convention, Hilton Motor Inn, Tarrytown, NY. October 21-22 are the happy days. So spend your July 4th weekend anticipating exhibits, lectures, 2-Meter FM, RTTY, contests, gabfests, New York sightseeing, fun. It's a blast! Write Dave Popkin, WA2CCF, 303 Tenafly Road, Englewood, N. J. 07631. He'll fire off the info to you.

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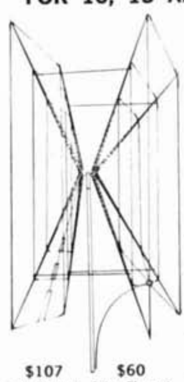

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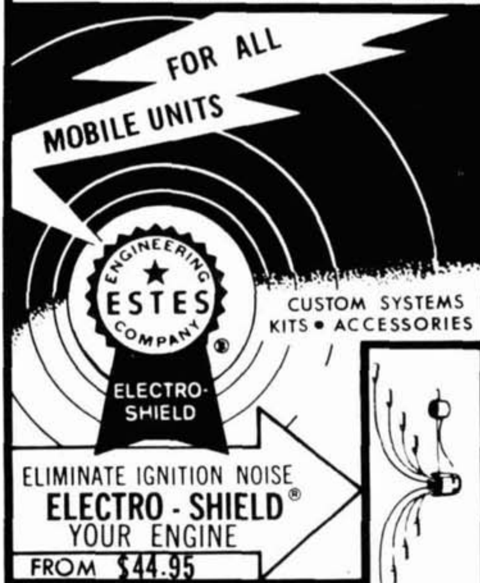

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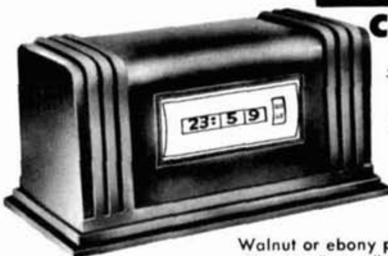

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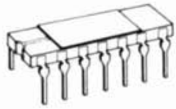

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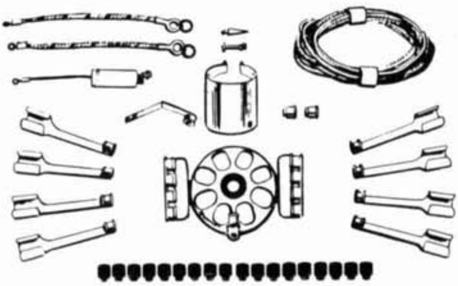

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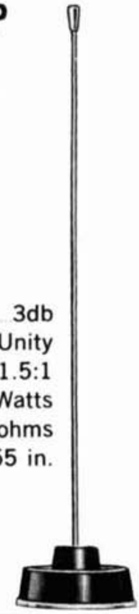

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