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# ham radio

**magazine**

AUGUST 1972



**FREQUENCY  
SYNTHESIZER**  
for the Drake R-4 receiver

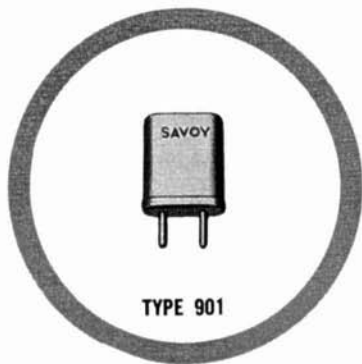
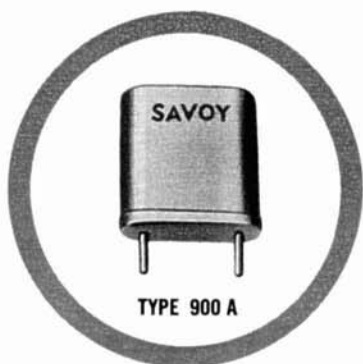
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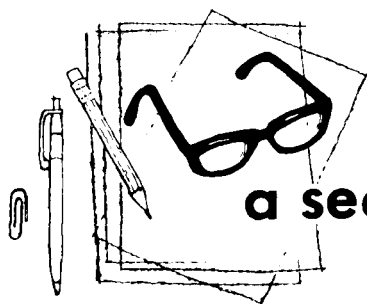
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## a second look

by Jim  
fisk

The newest solid-state device to come down the electronics turnpike offers some interesting circuit possibilities with existing vacuum-tube equipment. Although the new Teledyne junction-fet *Fetron* is not a new device so much as it is a new application of old principles, it offers a simple solid-state plug-in replacement for many vacuum tubes.

At the present time there are only two types of Fetrons available, a triode, the TS12AT7, and a pentode, the TS6AK5. These Fetrons are designed to replace, respectively, the 12AT7 and 6AK5. Both of these tubes are members of large families of similar types, so the two available Fetrons are usable in many different circuits.

The secret to Fetron operation lies in the development of high-voltage junction fets which will operate without breaking down with the high-voltage power supplies found in most vacuum-tube equipment. Most Fetron circuits are built with two fets in a cascode arrangement to obtain high gain; the basic circuit was described earlier in *ham radio*.\*

The Miller-effect capacitance is minimized in the Fetron by using an ultra-low-capacitance, high-gain transistor such as the 2N3823 in the input, with a high-voltage fet such as the 2N4881 at the output. The complete circuit goes into a metal can that has the same pin configuration as the tube it replaces.

Although the current cost of Fetrons is relatively high (\$15.60 for the TS12AT7 and \$12.50 for the TS6AK5), they make good sense as replacements for tubes in communications equipment because their characteristics don't drift with use, making periodic adjustment and alignment unnecessary, and they have

extremely long life. In fact, the estimated life for a Fetron has been calculated at 30 million hours — that is equivalent to running them 24 hours a day, 365 days a year, for over 3400 years!

In addition to higher amplification factors, Fetrons feature noise figures much lower than many equivalent tubes, and much lower heat dissipation; lower heat dissipation because the Fetron requires no filament supply. The lower heat characteristic should be particularly useful in reducing drift in your vfo.

Although most Fetrons are designed for the lower frequency ranges, Fetron pentodes have been operated as high as 500 MHz. Charlie Spitz, W4API, has been experimenting with Fetrons in amateur communications equipment, but he has been hampered somewhat by the limited number of different Fetrons. However, this will be solved in the near future.

Teledyne has already built Fetron equivalents of the 6JC6 and 6EW6, and these can be combined with derivatives of the TS12AT7 and TS6AK5 to provide plug-in replacements for a great variety of tubes.

Next on the list of Fetron replacements will be power pentodes such as the 6AQ5 and 6V6, as well as remote-cutoff pentodes such as the 6BA6. With high-volume production and simplified packaging, Fetrons could well become low-cost replacements for most vacuum tubes. Watch the pages of *ham radio magazine* for the latest developments with this interesting new device.

Jim Fisk, W1DTY  
editor

\*W1DTY, "ham notebook," *ham radio*, February, 1970, page 70.



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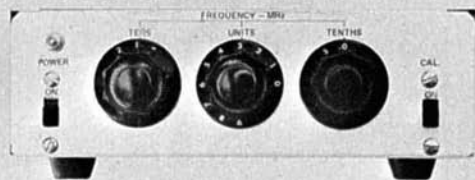
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## frequency synthesizer for the Drake R-4 receiver

This frequency synthesizer  
converts the Drake R-4  
series of  
amateur-band receivers  
to general coverage  
receivers  
that tune  
from 1.5 to 30 MHz

The vast majority of receivers in use in ham shacks today cover the amateur bands only, with the possible addition of one of the WWV frequencies. While this may be entirely adequate for normal operation, there are times when a general-coverage receiver is a very useful adjunct. Being able to receive WWV on only one frequency is often unsatisfactory, since obviously the selection of the optimum receiving frequency depends on the time of day and the distance from Fort Collins.

There are many other uses to which a general-coverage receiver can be put. Tracking down harmonics and other spurious frequencies can be simplified if you have such a receiver. Calibration of some home-built test equipment may require a receiver which covers frequencies other than the ham bands. And it is often desirable to use a receiver as a tunable i-f with vhf converters whose output frequencies are *not* in one of the high-frequency amateur bands.

While reviewing some of the literature which has been published describing the newer integrated circuits, it occurred to me that a frequency synthesizer might be a practical method for converting my Drake R-4 to general coverage without a separate crystal for each 500-kHz range. Frequency synthesis techniques have

Robert S. Stein, W6NBI, 1849 Middleton Avenue, Los Altos, California 94022

been around for some time, involving circuitry which has been complex and difficult. However, recent developments in integrated circuits have simplified the techniques to a point whereby the frequency synthesizer is a practical construction project for the relatively experienced amateur.

Before proceeding with the description of the frequency synthesizer which makes all-band coverage possible without modifying the receiver, a review of the receiver conversion process is in order. Although there are differences in the actual circuits of the Drake R-4, R-4A, and R-4B receivers, the conversion frequencies are identical; therefore, this discussion is applicable to any one of the three receivers.

If we disregard the pre-mixing arrangement used by Drake, the high-frequency oscillator requirements are simply that the oscillator be tuned 11.1 MHz above the low-frequency end of the desired 500-kHz tuning range. Since the high-frequency oscillator is crystal controlled and untuned, this requirement is translated to a crystal of the proper frequency. For example, the 3.5- to 4.0-MHz band requires a 14.6-MHz crystal (3.5 plus 11.1 MHz), which is supplied with the receiver. Drake specifies that the receiver will tune from 1.5 to 30 MHz with the proper crystal, excluding 5 to 6 MHz which bridges the 5645 kHz first intermediate frequency. (The apparent discrepancy between this  $f_i$  and the 11.1 MHz difference between the incoming and oscillator frequencies is taken care of in the pre-mixing scheme, and can be ignored for the purposes of this article.) Since the receiver covers this range in 500-kHz steps, because the vfo has a 500-kHz tuning range, it would require 58 crystals to cover the complete spectrum, starting with a 12.6-MHz crystal for the 1.5- to 2.0-MHz band, a 13.1-MHz crystal for the 2.0- to 2.5-MHz band, and finally ending with a 40.6-MHz crystal for the 29.5- to 30.0-MHz band.

A complete set of crystals would obviously be expensive. Furthermore, there would be the practical difficulty of using all the crystals at one time since the

receiver has provisions for externally installing only ten crystals in addition to the five amateur-band crystals mounted internally.

However, if we can generate frequencies of 12.6, 13.1, 13.6, 14.1, etc. through 40.6 MHz, and substitute them for the crystal-controlled high-frequency

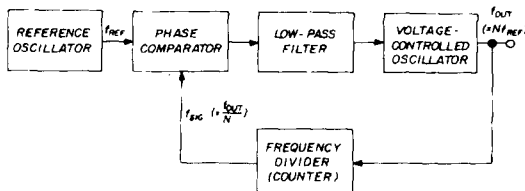


fig. 1. Basic phase-locked frequency synthesizer. The frequency divider is a variable-modulus, or programmable, counter.

oscillator, we can achieve all-band coverage within the specified range of the receiver. Thus, this becomes the requirement for the frequency synthesizer.

### basic phase-locked loop frequency synthesizer

The integrated circuits to which I briefly referred are designed for use in phase-locked loops, one application of which is frequency synthesis. Fig. 1 shows the basic phase-locked frequency synthesizer. A stable reference frequency is applied to one input of a phase comparator. The output of the phase comparator is a dc voltage, which passes through a low-pass filter and controls the frequency of a voltage-controlled oscillator (vco). This oscillator generates the desired frequency, which may be any multiple of the reference frequency. The vco output is also applied to a frequency divider whose function is to divide the vco output frequency to the same frequency as that of the reference oscillator.

Let's assume that the reference oscillator frequency is exactly 100 kHz and that an output frequency of 12,600 kHz is required. If we have a divider or programmable counter which will divide by 126, the signal input to the phase

comparator will also be 100 kHz when the vco output is exactly 12,600 kHz. This is accomplished by the phase comparator producing a d-c output which "tunes" the vco until it is exactly 12,600 kHz. The divided vco frequency is then exactly 100 kHz, the same as the reference frequency. Thereafter, the vco out-

greater detail, references 1 through 3 provide a wealth of information and design data.

A block diagram of the R-4 high-frequency oscillator frequency synthesizer is shown in fig. 2. The loop reference is a 100-kHz crystal-controlled oscillator which uses a Signetics LU380A quad two-

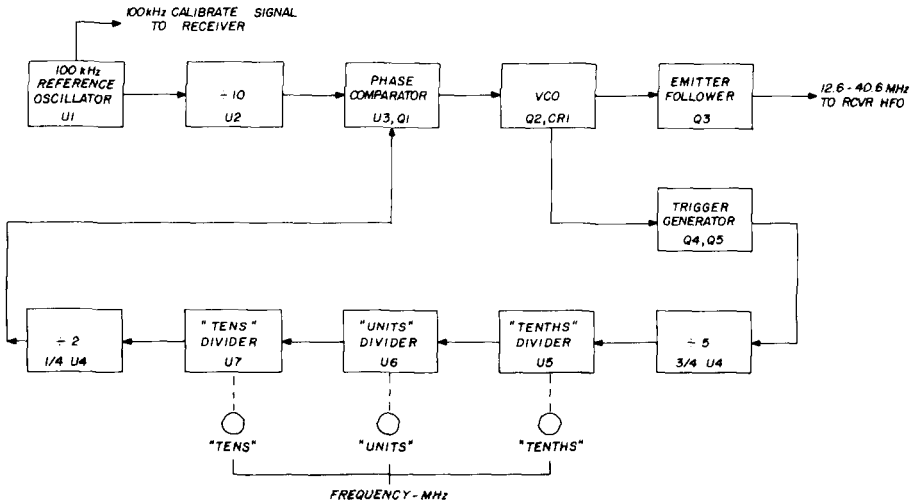


fig. 2. Block diagram of the R-4 hfo frequency synthesizer. The only tuning controls are the front-panel mounted tens, units and tenths rotary switches.

put will stay at 12,600 kHz; any variation from this frequency changes the signal input to the phase comparator, which in turn produces a dc output change and brings the vco back to 12,600 kHz. Thus, the output frequency is locked to the reference frequency, and has essentially the same stability as the reference oscillator.

By using a frequency divider which can be programmed, it is possible to obtain virtually any number of discrete frequencies which are integral multiples of the reference frequency, all of which are phase locked to the reference oscillator. The low-pass filter keeps the reference frequency from appearing at the vco and establishes the lock-up time of the loop. A rigorous analysis of the feedback loop and filter involves extensive mathematics and is beyond the scope of this article. For those of you interested in

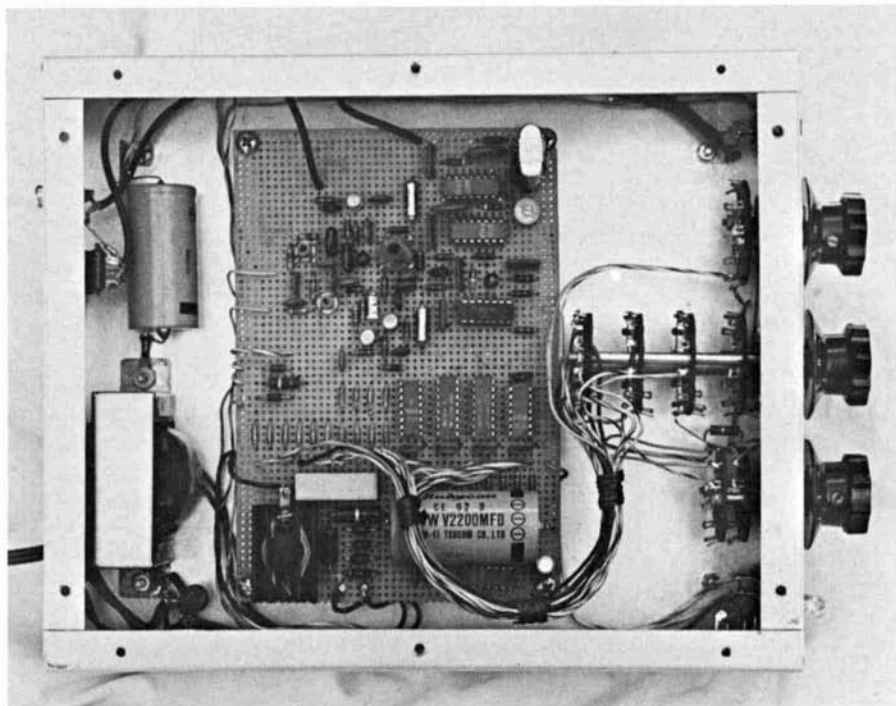
input NOR gate. The oscillator output is divided by ten (for reasons which will be discussed later) and the resultant 10-kHz signal is applied to the reference input of a Motorola MC4044P phase comparator. Because I used the 100-kHz calibrator crystal from the R-4 in the synthesizer, I provided a 100-kHz output jack on the synthesizer in order to introduce a calibration signal back into the receiver.

This is shown in fig. 2, and is covered in greater detail under the circuit description and interconnection portions of this article. It should be noted, however, that this technique will eliminate the 25-kHz calibration markers in the R-4B because the 100-kHz calibration signal is fed back into the receiver antenna input by way of the accessory socket on the receiver. If you wish to retain the 25-kHz markers, use a separate crystal in the synthesizer and delete the calibration circuit.



The low-pass filter, which appeared in fig. 1, is not shown separately in fig. 2 since part of it is integral to the MC4044P with the remainder made up of discrete components. The output of the phase comparator/filter is applied to a varactor diode which controls the frequency of the vco, a Fairchild SE3005 transistor.

reference input to the phase comparator can be explained. If the reference signal were 100 kHz, the divider would need only a divide range of 126 to 406 in order to divide the vco output of 12.6 to 40.6 MHz down to 100 kHz. However, at the upper frequency limit of 40.6 MHz, the propagation delays through three



Interior view showing construction of the synthesizer. The reference oscillator is at the top right of the board, the vco and trigger generator are at the upper center, and the frequency-divider ICs and steering diodes are to the left of the multi-section switch. Power supply components are mounted at the bottom of the board and on the chassis to the left of the board.

The vco output, 12.6 to 40.6 MHz in 500-kHz steps, is routed to the R-4 via a 2N709 emitter follower. The vco output also drives a trigger generator which uses a pair of 2N3478 transistors. The trigger generator is needed to convert the amplitude and waveform of the vco output to that required by the TTL integrated circuits in the frequency divider.

The frequency divider is a variable-modulus counter which can be programmed to divide by any factor between 1260 and 4060 in steps of 50. It is at this point that the reason for using a 10-kHz

synchronous counters arranged in a ripple-through configuration would prevent proper operation of the circuit.

Therefore, the trigger generator output, which is at the vco frequency, is first divided by five in a quinary divider which consists of three of the four flip-flops in a Signetics 8290A high-speed decade counter. The remaining flip-flop is placed at the end of the divider chain, thereby dividing the vco output by a fixed factor of ten. Thus it can be seen that the reference signal must also be divided by ten if the reference and signal inputs to

the phase comparator are to be equal.

Although the trigger generator output could have been divided by ten before it was applied to the programmable counter, placing the binary divider at the end of the chain results in a square-wave output which is readily observable on oscilloscopes of limited frequency response and rise time.

Returning to the signal flow shown in fig. 2, let's assume that the desired vco output is to be 39.6 MHz, for a receiver tuning range of 28.5 to 29.0 MHz. The 39.6-MHz signal from the trigger generator is divided by five to a frequency of 7.92 MHz. To obtain a 10-kHz output from the divider chain, the dividers must be programmed to divide by 396. (Remember that there is a fixed divide-by-two at the end of the chain.) By programming the *tenths* divider to divide by six, the *units* divider to divide by nine, and the *tens* divider to divide by three, the output from the *tens* divider is 7.92 MHz divided by 396, or 20 kHz. This is divided by two in the binary section of the 8290A, resulting in a 10-kHz signal being applied to the phase comparator.

The designations *tenths*, *units* and *tens* used for the dividers refer to the panel markings of the rotary switches which are used to set the dividers. Each of the dividers is a decade counter which can be preset to count by any integer between one and nine. The *tenths* divider, being the first in the chain, determines the least significant figure in the total count. The *units* divider divides the output of the *tenths* divider; therefore its count (multiplied by 10) determines the middle figure in the three-digit count, while the count (multiplied by 100) of the *tens* divider determines the most significant figure in the total count.

In order to obtain divide ratios of 1260 to 4060 in steps of 50, the variable-modulus counter must have a range of 126 to 406 in steps of 5 (since there is a fixed division by ten), i. e., 126, 131, 136, 141 . . . 401, 406. Therefore, the first, or *tenths*, divider is arranged to divide by one or six, as controlled by its

associated switch. The *units* divider can divide by any number from one to nine, and the *tens* divider can be preset to divide by one, two, three or four.

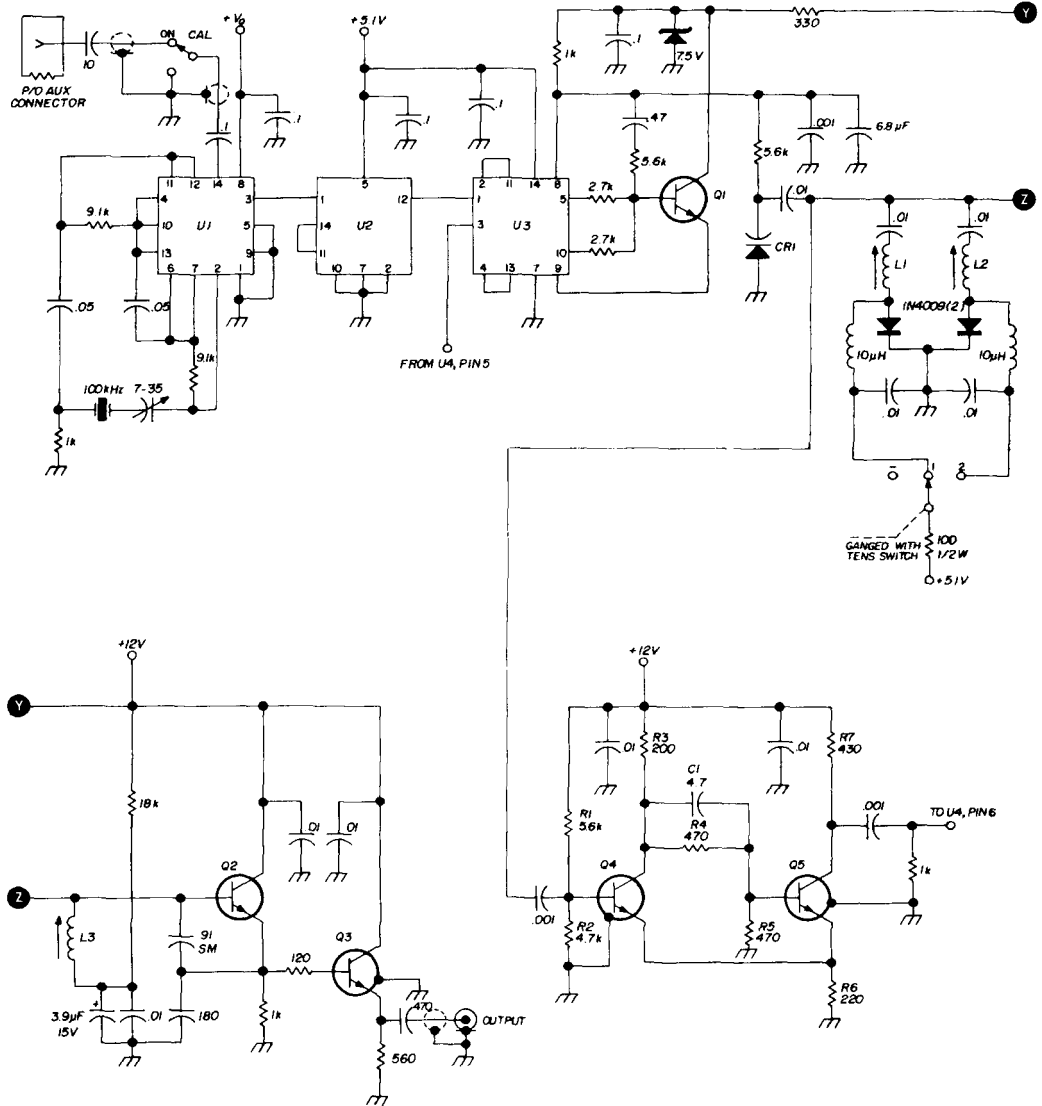
Those of you who have looked at the front-panel knobs with a sharp eye will note that despite the foregoing explanation, the *tenths* switch is calibrated 0.5 and 0.0, while the *tens* switch is calibrated 2, 1 and —. This was done so that the switches would indicate the receiver *tuning* frequency, not the hfo injection frequency. Obtaining this convenience required a considerable increase in switching complexity.

Correctly marking the *tenths* switch involved only applying the proper designations to the switch. When the *tenths* divider divides by one, the receiver tuning range starts at an integral megahertz; therefore that switch position is marked 0.0. When the divider divides by six, the receiver tuning range starts at a half-megahertz; thus that position is designated 0.5.

Explaining the *tens* switch is a bit more complicated. Remembering that the synthesizer output is always 11.1 MHz above the start of the receiver tuning range, we can see that the *units* divider will always divide by a factor of one greater than the *units* digit in the receiver tuning range. For example, if the receiver tuning is to start at 15.5 MHz, the *units* divider must divide by 6. This causes no problem except when the receiver tuning range starts at 9.0, 9.5, 19.0, 19.5, 29.0, or 29.5 MHz. The unit count of 9 means that the *units* divider must divide by 10, which is reserved for the *tens* divider. Therefore, the 9 position of the *units* divider switch actually causes the divider to divide by one and sets the *tens* divider to the next divide ratio.

The net result is that only three, not four, switch positions are needed for the *tens* divider. Even though the divider must divide by four (to obtain a division of 401 and 406 for 29.0 and 29.5 MHz respectively), setting the *units* switch to 9 sets the *tens* divider to the next count and eliminates the fourth switch position. The "—" position corresponds to a zero

100kHz CAL OUT



- |     |  |        |  |
|-----|--|--------|--|
| CR1 | varactor (Motorola MV1401)                                     | Q3     | 2N709, 2N918, or 2N3563  |
| L1  | 8 turns no. 28E, close wound on 5/32" diameter slug-tuned form | Q4, Q5 | 2N3478   |
| L2  | 5 turns no. 28E, close wound on 5/32" diameter slug-tuned form | U1     | quad 2-input NOR gate (Signetics LU380A or SE380A)                               |
| L3  | 8 turns no. 28E, close wound on 7/32" diameter slug-tuned form | U2     | decade counter (Signetics N7490A, Texas Instruments SN7490N or Motorola MC7490P) |
| Q1  | 2N3391 or Motorola MPS6571                                     | U3     | phase-frequency detector (Motorola MC4044P or MC4344P)                           |
| Q2  | Fairchild SE3005, 2N918, 2N3563, or 2N5179                     |        |  |

fig. 3. Schematic of the reference oscillator, phase comparator, vco and trigger generator circuits. Capacitors designated SM are silver micas and those with polarity indicated are electrolytic; all others are ceramic.

when the switches are set to a frequency of 9.5 MHz or lower.

### circuit

Fig. 3 shows the frequency-generating circuits, the phase comparator and the trigger generator. The 100-kHz crystal oscillator uses two gate sections of U1, a quad two-input NOR gate; the remaining two sections are used as buffers. A logic diagram of the circuit is shown in fig. 4. Each of the two gates used in the oscillator is forced into linear operation by connecting a 9.1k resistor between the output and input; a 100-kHz signal is developed by connecting the crystal in series with the feedback loop from the output of the second gate to the input of the first. A 7-35 pF ceramic trimmer in series with the crystal permits adjustment of the crystal frequency. The remaining two gates buffer the output to the succeeding divide-by-ten stage and to the calibrate input of the R-4. The calibrate signal is disabled by grounding the output of the gate by means of the calibrate switch.

Returning to fig. 3, the 100-kHz signal is divided by ten in U2, and the resultant 10-kHz signal applied to pin 1 of phase comparator U3. This device consists of two separate functional parts, a phase detector and an amplifier, which must be connected externally. The outputs from the phase detector section appear at pins 5 and 10 and are applied to the loop filter consisting of two 2.7k resistors, a 5.6k resistor and a 0.47-μF capacitor. Transistor Q1 isolates the RC filter from the amplifier section of U3, the input of which is pin 9. The amplifier output from pin 8 is applied as the control voltage to varactor CR1. It was found necessary to add the 6.8 and 0.001 μF capacitors to reduce the amplitude of the 10-kHz reference signal which reached CR1. The supply voltage for the amplifier section of U3 is regulated at 7.5 volts by means of a zener diode and 330-ohm dropping resistor from the 12-volt supply.

The voltage-controlled oscillator consists of Q2, a high-frequency npn transistor, in a Colpitts circuit with varactor

CR1 across the tank circuit. CR1 is a Motorola MV1401 and has a ratio of maximum-to-minimum capacitance of approximately ten, as compared to usual ratios of two to four for conventional varactors. (It also happens to be the most expensive single component in the entire synthesizer.) Despite the large capacitance ratio, the oscillator cannot cover the entire range of 12.6 to 40.6 MHz without switching. This is accomplished by diode switching, using a switch section ganged with the *tens* divider switch.

In the "—" position of the *tens* switch, the two 1N4009 diodes do not conduct, so coils L1 and L2 are each effectively in series with a 10-μH choke to ground. The high value of this inductance has only stray effect on the circuit; thus the oscillator frequency is determined by coil L3 and the tank-circuit capacitance. When the *tens* switch is set to position 1 or 2, one of the diodes is biased into forward conduction and

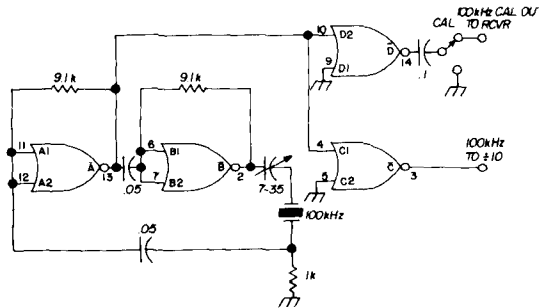
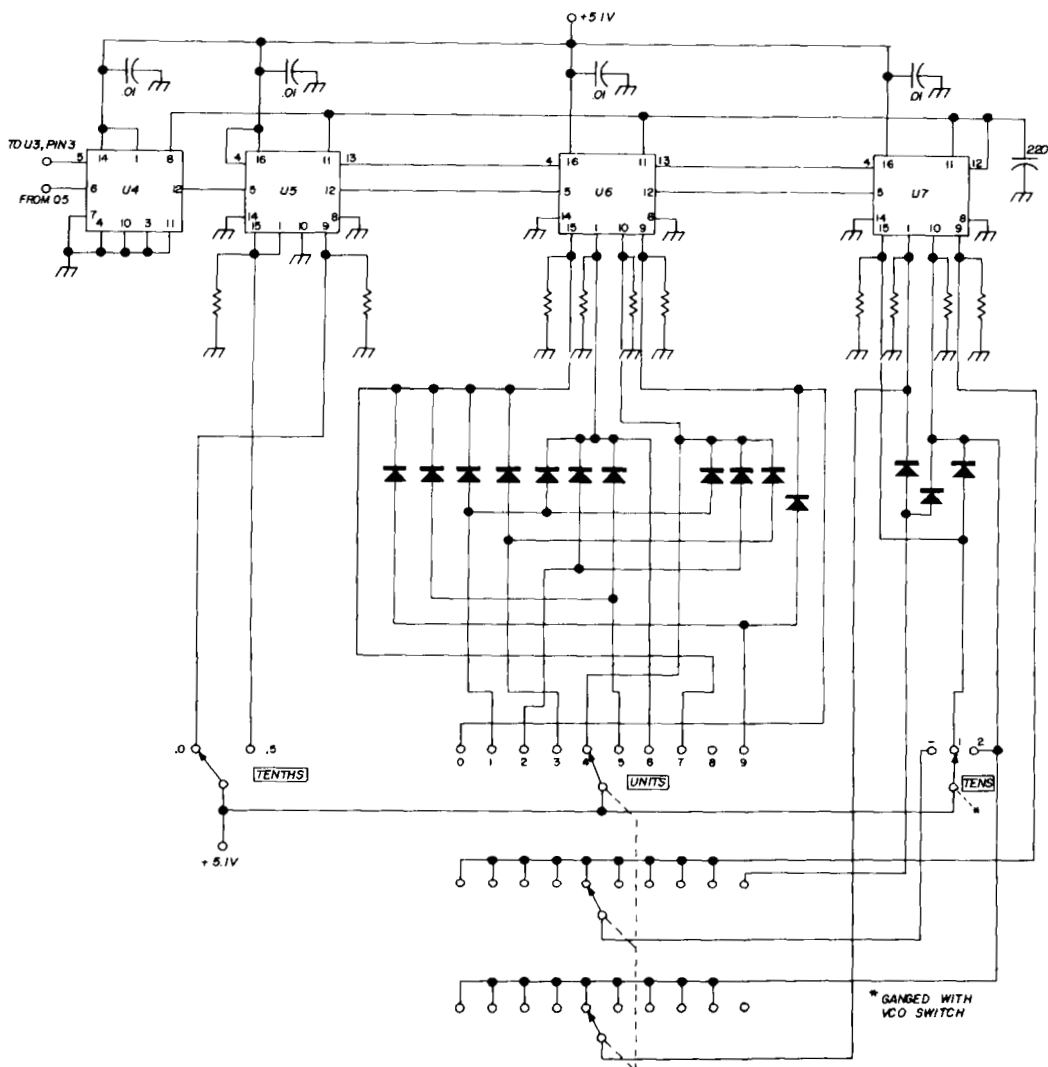


fig. 4. Logic diagram of the reference oscillator circuit.

brings the low end of the associated coil close to rf ground, shunting L3 and thereby lowering the tank-circuit inductance. The 100-ohm resistor in series with the switch arm limits diode current to a safe value.

The oscillator output is taken from the emitter of Q2 and coupled to the base of emitter follower Q3, which isolates the vco from the loading effects of the receiver and connecting cable.

Oscillator output is also taken from



U4 high-speed decade counter (Signetics 8290A)

U5,U6, U7 synchronous up/down counter (Signetics N74192A or Texas Instruments SN74192N)

fig. 5. Schematic of the frequency divider. All diodes are general-purpose germanium, all resistors are 910 or 1000 ohms, and all capacitors are ceramic. Switches are non-shorting rotary types.

the base of Q2 and coupled to the base of Q4, which with Q5, forms a Schmitt trigger. This circuit, when driven by the approximate sinusoidal output of the vco, generates a rectangular wave of several volts peak-to-peak amplitude with fast rise and decay times, which are required

to toggle the TTL logic of the frequency divider. Since the Schmitt trigger is not common in amateur equipment, a brief explanation of its operation is in order.

In the absence of an input signal to the base of Q4, that transistor is normally cut off by the combination of base bias at the

junction of R1 and R2 plus the reverse bias at its emitter resulting from the collector current of Q5 flowing through R6. The resultant high positive voltage at the collector of Q4 forward biases Q5 to saturation. When an incoming signal of sufficient amplitude is applied to the base of Q4, the positive excursion drives the

voltage at the collector of Q5 drops. Thus a rectangular output pulse is generated from a sinusoidal input, having rise and decay times sufficiently short to trigger the TTL decade counter which follows the Schmitt trigger.

As shown in fig. 5, the output of Q5 is applied to pin 6 of U4, which is the clock

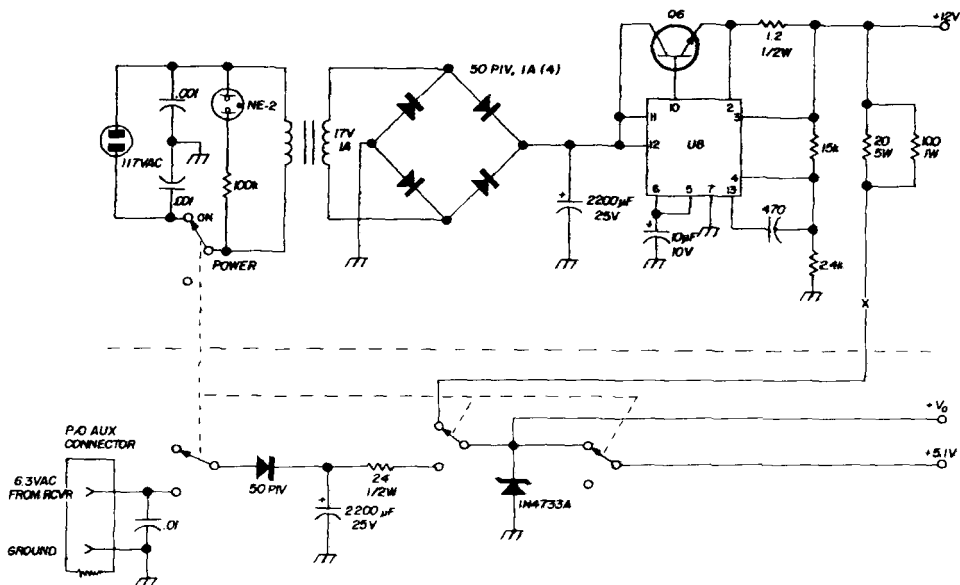


fig. 6. Schematic of the power supply. Power transistor Q6 is a 2N3054 or equal. U8 is a Signetics NE550A. See text for details of the power transformer and the meaning of point X.

base into conduction and causes collector current to flow. The drop in collector voltage is coupled via R4 and C1 to the base of Q5, reducing the base current and the collector current of Q5. The decrease in current lowers the voltage drop across R6, which results in increased collector current in Q4. This regenerative action causes Q4 to saturate quickly and cuts off Q5, producing an extremely fast rise time as the collector voltage of Q5 reaches the supply voltage level.

When the input signal becomes negative, the circuit operation reverses, cutting off Q4 and driving Q5 back into saturation. The regenerative action likewise results in a short decay time as the

input for the quinary divider section of that decade divider. The output from pin 12 is applied to pin 5 of U5; U5, U6 and U7 are programmable decade counters set up as a ripple-through or variable-modulus counter. The count or divide ratio is determined by setting pins 1, 9, 10 and 13 of each IC high or low. This is accomplished by the rotary switches described previously; steering diodes are used to reduce the number of switch sections which would otherwise be required. The two extra sections of the units switch are needed to advance the count of the tens divider when the units switch is in position 9.

The output of the variable-modulus



counter is applied to pin 8 of U4, the clock input of the binary section of that device. This flip-flop generates a 10-kHz square wave from the 20-kHz input from the counters; the square wave is the signal input which is applied to pin 3 of U3 for phase comparison with the 10-kHz reference signal.

## power supply

The power requirements of the frequency synthesizer are +12 volts at 65 mA and +5 volts (nominal) at 360 mA. The 12-volt supply must be well filtered to prevent frequency modulation of the vco. The schematic appears in fig. 6.

A full-wave bridge rectifier is supplied from a 17-volt, 1-ampere transformer. The positive output from the rectifier is fed to the input (pin 12) of voltage regulator U8. Since the current drain exceeds the current limitation of the regulator, power transistor Q6 is incorporated as a series-pass device. The output of the regulator is set to 12 volts by means of the 15k and 2.4k resistors connected to pins 3 and 4 of U8. The 5.1-volt supply for the logic elements is developed across a 1N4733A zener diode. Current limiting and overload or short-circuit protection is provided by monitoring the voltage drop across the 1.2-ohm resistor in series with the emitter of Q6.

As previously noted, the 100-kHz calibrator crystal from the R-4 was removed from the receiver and used as the reference oscillator, with provisions incorporated to reintroduce the 100-kHz signal back into the receiver. In order to enable the reference oscillator in the synthesizer for this calibrate signal when the synthesizer is not in use, an auxiliary power supply is included which keeps the reference oscillator energized whenever the receiver is on, whether or not the synthesizer is being used.

A 4-pole, double-throw slide switch is used as a power switch. In the on position, 117 volts ac is applied to the transformer, and all circuits in the synthesizer are energized. When the power

switch is in the off position, a separate half-wave rectifier and filter are connected through a cable to the 6.3-volt pin of the accessory power receptacle on the rear of the R-4. When the receiver is on, this voltage is rectified and filtered, and applied to the same zener diode used in the main power supply of the synthesizer. However, only the reference oscillator circuit is powered from the auxiliary power supply.

If a separate crystal is used in the synthesizer and the calibrator crystal is left in the receiver, the auxiliary power supply and its switching circuit can be eliminated. This is accomplished by deleting all of the circuit below the dashed line in fig. 6 and connecting the 1N4773A zener diode between point X and ground. The 5.1-volt supply is obtained from point X. Pin 8 of U1 (see fig. 3) should be connected directly to the 5.1-volt supply.

## construction

Construction of the frequency synthesizer is shown in the various photographs. An 8 x 10 x 2½ inch aluminum chassis is inverted and used as a base. Four rubber bumper feet mounted on the bottom prevent scratches from the screws and nuts which fasten the parts to the chassis.

The three rotary switches, the power and calibrate slide switches and the NE-2 pilot lamp are mounted on the front apron. On the rear apron are the coax output connector and a three-pin miniature connector used for the 6.3-volt and 100-kHz calibrate signal connections between the synthesizer and the receiver. The power transformer and the filter capacitor for the auxiliary power supply are mounted directly on the chassis. All other parts are mounted on a piece of perforated Vector board, copper-clad on one side. The board measures 4½ x 6½ inches overall and the hole spacing is 0.1 inch. This spacing is important, since it matches the pin spacing on DIP integrated circuits. The use of copper-clad board provides a ground plane for the rf and high-speed logic circuits.

The ungrounded leads of the various

components are soldered to insulated pads cut into the board by means of a Vector P-138C pad-cutting tool, and then connected by conventional wiring. The usual precautions of short, direct leads should be observed; although the highest radio frequency involved is only 40.6 MHz, vhf techniques should be followed because of the high-speed logic pulses. The integrated circuits may be soldered directly to the board, or sockets or Molex pins may be used. My personal preference is the latter, since they are inexpensive and eliminate the very difficult job of unsoldering an IC should it be necessary.

The power supply series-pass transistor is mounted in a Thermalloy 6166-B heat sink, which is then mounted on the perf board. Although the heat sink is quite small, it is adequate for the power dissipated in the transistor and holds the case temperature to about 60° C, well within the temperature-power dissipation curve limits.

All connecting leads to and from the board should be soldered in place on the board before connecting the other ends. Leave plenty of wire on each lead to make the connections to the parts which are not on the board so that if any troubleshooting must be done it will not be necessary to unsolder the wires. Wiring for the output and calibrate signals is made with RG-174/U coax and should be kept reasonably short. All other wiring carries dc only, making lead length un-critical. The board is mounted on four metal standoff posts, which ground the copper side to the chassis.

The power transformer should supply 16 to 17 volts ac at approximately 0.5 ampere. Any higher voltage only results in greater heat dissipation in the series-pass transistor. The simplest way to obtain a transformer is to modify one of the inexpensive 24-volt, 1-ampere transformers which are readily available. First connect the primary of the unmodified transformer and measure the secondary voltage. Then disconnect the transformer and carefully cut through the outer paper insulation to expose the outer winding layer of the secondary. Remove ten turns

from this winding and remeasure the secondary voltage. The difference between the new lower voltage and the original secondary voltage, divided by ten, is the volts per turn of the secondary. From this figure calculate the number of turns required for a voltage equal to the difference between the original secondary voltage and the desired 17 volts, and remove this number of turns from the secondary. After checking the new secondary voltage, wrap the transformer windings with insulating tape.

There is a short-cut to this procedure if you can see and count the number of layers comprising the secondary. This is generally possible by closely examining and probing the open ends of the windings. Assume that you have determined that the secondary has eight layers, which means that each layer develops approximately 3 volts (24 divided by 8). Removing two and a half layers should result in a secondary voltage of 16.5 volts, which is just about right.

A cover for the chassis was made from aluminum cane-metal stock, which provides adequate ventilation for dissipating the heat generated. The numbered positions on the rotary switch knobs were made by using white number transfers on the black skirts of the knobs. Several heavy coats of Krylon fixative were sprayed on to keep the numbers from peeling off.

### alignment and test

After all wiring and connections have been checked and rechecked, the synthesizer is ready for the few adjustments necessary to set the vco on frequency. The only test equipment absolutely necessary is a voltmeter, although a grid-dip oscillator and oscilloscope can be helpful.

Apply power to the synthesizer and check the supply voltages to make sure that they are within five percent of the nominal values. Then make sure the reference oscillator is working by bringing a lead from the receiver antenna jack close to the crystal. Harmonics of the 100-kHz oscillator should be heard in the

receiver. The oscillator may also be checked with a scope; 100-kHz square waves should be observed at pin 3 of U1, and 10-kHz square waves should be present at pin 12 of U2. If the oscillator is not working, adjust the trimmer capaci-

is reached. (If the voltage does not drop between switch positions 1 and 0, it is nothing to be concerned about. The vco may not tune this low, but the switch position corresponds to a receiver tuning range starting at 500 kHz, which is



The frequency synthesizer atop the Drake R-4 receiver, which has the new general-coverage preselector scale on its front panel. Operating the equipment in this physical configuration is not recommended because the synthesizer interferes with convection flow from the R-4.

tor; it is not necessary for the crystal to be oscillating at exactly 100 kHz at this time.

Connect a voltmeter between pin 8 of U3 and ground. Set the rotary switches for a frequency of 9.5 MHz and move the tuning slug of L3 into the coil until a maximum positive voltage is obtained. This should be 7.5 volts, the same as the zener-regulated voltage applied to U3 and Q1. Slowly tune the slug out of L3 until the meter reading drops to 7.1 volts, or about 0.4 volt less than maximum.

Rotate the *units* switch toward zero, noting that the voltmeter reading drops with each switch position until position 1

outside the receiver preselector tuning range.) If you have a grid dipper available, couple it to L3. It should indicate 12.6 MHz in position 1 of the *units* switch, and 20.6 MHz in position 9.

Next, set the rotary switches for a frequency of 19.5 MHz, and adjust the slug in L1 for a voltmeter reading of approximately 6.5 volts. *Do not touch the slug in L3.* Again turn the *units* switch toward zero and note that the voltage drops with each switch position, including position 0 this time. Turn the *tenths* switch to its 0.0 position and make sure that the voltage also decreases at this step. A grid dipper coupled to L1 should

indicate 21.1 and 30.6 MHz for the minimum and maximum *units* and *tenths* switch settings.

Finally, set the rotary switches for a frequency of 29.5 MHz and repeat the preceding adjustment for L2. The frequency range of the vco with the *tens* switch in position 2 is 31.1 to 40.6 MHz.

As a further check, a scope can be connected to pin 3 of U3. If the scope has a triggered sweep, the signal at pin 12 of U2 should be connected to the external trigger input of the scope. A 10-kHz square wave should be displayed on the scope for each and every position of the frequency selection switches on the synthesizer. This completes the entire alignment procedure, except for setting the crystal to exactly 100-kHz after the synthesizer has been connected to the receiver.

## operation

The vco output from the synthesizer is fed into one of the auxiliary crystal sockets on the rear of the receiver, so that no modification of the receiver is required. The auxiliary crystal sockets are arranged in two rows of five each, one above the other. The socket holes in these two rows which are adjacent to each other (the bottom holes in the top row of sockets and the top holes in the bottom row of sockets) will be referred to as the *inside* holes.

If the synthesizer is to be used with a Drake R-4, build a cable as shown in fig. 7A, and insert the crystal-base end into one of the auxiliary crystal sockets so that the pin connected to the coax shield is in the *inside* hole of the crystal socket. The 150-ohm resistor keeps the receiver hfo from oscillating on its own.

If the synthesizer is to be used with the Drake R-4A or R-4B, use the cable shown in fig. 7B. The pin on the center conductor goes to the *inside* hole of one of the auxiliary crystal sockets, while the coax shield is grounded to the receiver chassis. The ground stud on the receiver is a convenient terminal for the shield.

Fig. 7C shows the cable to be used

when the auxiliary power supply and 100-kHz calibrate circuit have been included in the synthesizer. The lead lengths are not critical, but there is little point in having the cable appreciably longer than the output cable.

After the receiver and synthesizer have been interconnected, turn on both and

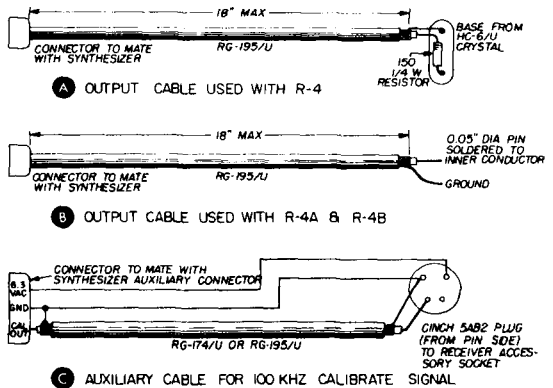


fig. 7. Interconnecting cables.

set the synthesizer switches to one of the WWV frequencies suitable for good reception. Set the receiver *xtals* switch to the position which corresponds to the crystal position now occupied by the synthesizer output cable. Refer to the preselector tuning chart in the receiver manual and set the receiver *band* switch and *preselector* tuning to the settings specified for the frequency selected. WWV should be heard when the receiver vfo is tuned to zero. Then allow the synthesizer and receiver to warm up for about a half-hour.

If you have incorporated the calibrate function in the synthesizer, turn on the calibrate switch and carefully adjust the 100-kHz crystal trimmer capacitor in the synthesizer to zero-beat the calibrate signal with WWV. It may be necessary to retune the receiver slightly when doing this because changing the frequency of the crystal changes the synthesizer output frequency which is now the receiver hfo frequency.





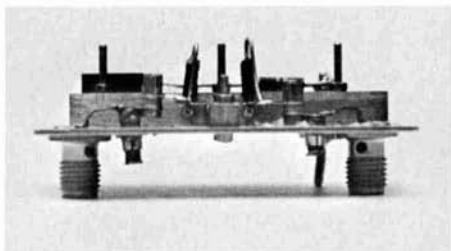
## solid-state 2304-MHz preamplifier

High-performance  
preamplifier for  
2304 MHz  
provides 8 dB gain  
and 5.7 dB  
noise figure

Dolph Vilardi, WA2VTR, 14 Oakwood Terrace, Spring Valley, New York

Receivers and converters for the uhf range have traditionally used mixer-type front ends because of the unavailability of devices which could provide useful amplification and sufficiently low noise figure. The noise figures provided by uhf mixers are limited not only by the devices available, but to a very large extent by the circuit in which they are used. For simplicity, amateurs usually use single-ended diode circuits.

Until recently, if uhf signal amplification was essential, parametric amplifiers were about the only way to obtain a



Side view shows vertical mounting of resistors. If 1/8-watt resistors are available, these components can be smaller. The partition mounting screws are longer than necessary. The miniature potentiometer is mounted with a drop of epoxy. It is adjusted through a hole in the side of the Pomona box.



substantial improvement in performance. However, several years ago transistors became available which had useful gain and reasonable noise figures on 1296 MHz. The cost of these early uhf transistors was high, but they eventually

amateurs, and it had to offer improved performance over the straight diode mixer front end. With this in mind, I first tried to modify the pi-network circuit used in the 1296-MHz preamplifier. However, all attempts to develop a 2304-MHz

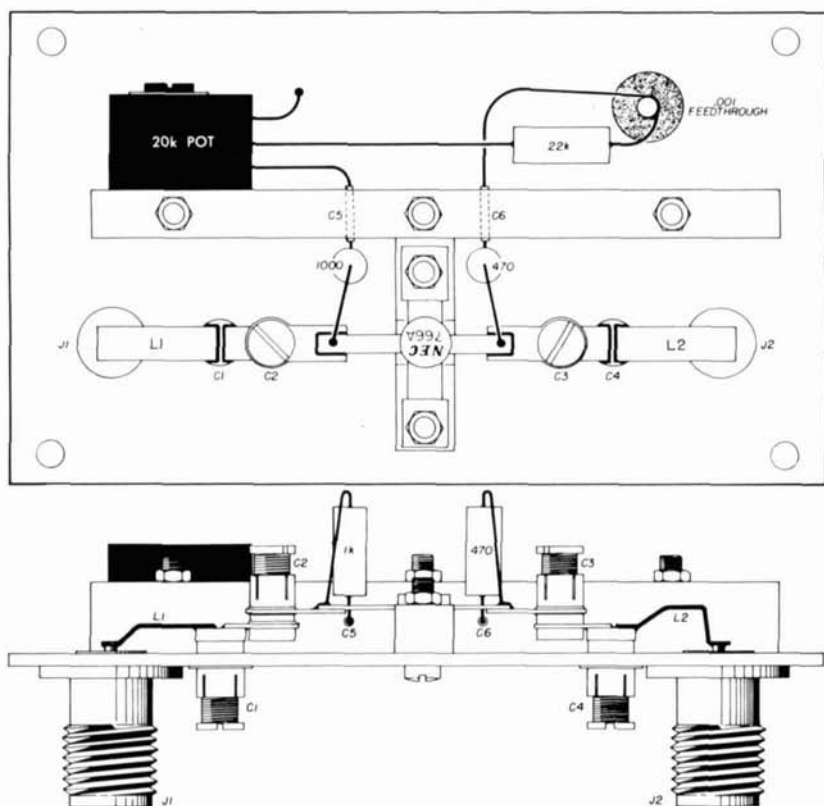


fig. 1. Layout of the 2304-MHz preamp. This drawing is twice actual size. Feed-through capacitors C5 and C6 consist of Teflon-covered wire, fit snugly into hole in 1/8" chassis partition.

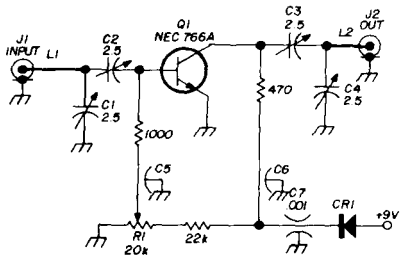
found their way into many high-performance 1296-MHz converters. Now the price of these devices is substantially lower, and the day of the parametric amplifier on 1296 and 2304 MHz, with its attendant isolators and circulators, is gone.

The success of the Nippon Electronics 766A as an 1296-MHz rf amplifier,<sup>1</sup> and its published performance specifications for 2000 MHz made it an interesting candidate for 2304 MHz. The proposed 2304-MHz pre-amplifier had to be easy to build, so it could be duplicated by other

preamplifier using this concept were unsuccessful, primarily because of problems with self oscillation.

About this time, my attention was called to a capacitor being marketed by the Johanson Corporation which was designed specifically for uhf impedance-matching purposes. This capacitor, the Johanson model 4991, is a miniature dual-variable capacitor with two 0.8- to 10-pF sections.

After conversations with Mr. Harvey Bruning of the Johanson Corporation



- C1,C2 0.5-2.5 pF piston trimmers
- C3,C4 Johanson Gigatrimms JMC 6927
- C5,C6 gimmick feedthrough capacitors (see fig. 2)
- C7 0.001  $\mu$ F feedthrough
- R1 miniature 20,000-ohm potentiometer
- L1 brass strip, 0.001" thick, 3/32" wide, 5/16" long
- L2 brass strip, 0.001" thick, 3/32" wide, 3/8" long, formed as shown in fig. 2
- J1,J2 miniature coaxial connectors (type SMA)

fig. 2. Schematic diagram of the 2304-MHz preamplifier. Transistor Q1 is a Nippon Electronics type V766A.

regarding a T-network arrangement using the new dual capacitors, he suggested a new device marketed by their company called the *Gigatrim*. These are very small concentric air-type capacitors which lend themselves to the circuit I wanted to use for the 2304-MHz preamplifier.

Mr. Bruning sent along some *Gigatrimms* for experimental use, and with a layout suggested by him, the preamplifier seemed to work quite well. With the *Gigatrimms* I was able to keep the overall size small enough so lead inductance did not become a limiting factor. Tuning the preamplifier is quite easy, and when built as described in this article, it shows absolutely no signs of instability.

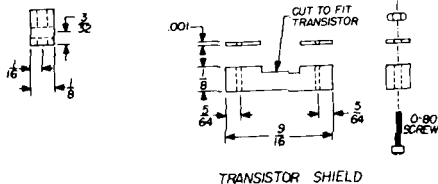
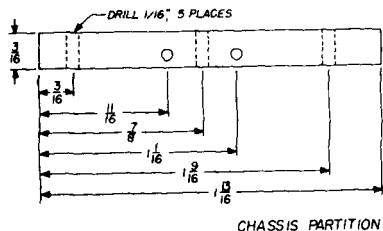


fig. 3. Full-size layout drawings of chassis partition and transistor shield. Material is brass.

Construction of the preamplifier should be obvious from the drawings and photographs, and should present no difficulties to the experienced uhf worker. Although Johanson states that the solder used in their products will withstand relatively high temperatures, I used TIX high-conductivity, low-temperature solder which melts at 250° and flows easily.\* When using this type of solder, which contains indium, you must be careful because the liquid solder flows readily, and can get into the threads of the capacitor where it doesn't belong. According to Mr. Bruning, regular 60/40 solder should work well.

When you are building the 2304-MHz preamplifier, it is important that you follow directions closely, particularly in regard to the dimensions of the holes in the center partition which make up the feedthrough capacitors (C5 and C6). Close fit is essential for these capacitors to work as required.

Almost any miniature potentiometer may be used for R1. The *Cubic* potentiometer shown in the photographs is very compact and is easily mounted with a drop of epoxy. For the fixed resistors, use the smallest size that is available, 1/8 watt if possible.

### tuning

The preamplifier is initially tuned up by injecting a 2304-MHz signal through a suitable attenuator and filter, and tuning all capacitors and adjusting the bias control (R1) for maximum gain. This technique is adequate because the noise figure

\*TIX low-temperature solder is available from the Brookstone Company, 5 Brookstone Building, Peterborough, New Hampshire 03458.

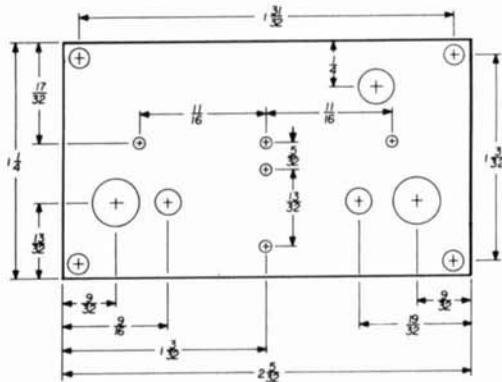


fig. 4. Full-size drill template for the chassis.

of the converter with which it will be used is probably higher than the gain of this single-stage preamplifier.

A simple, low-level signal source suitable for tuning up the 2304-MHz preamplifier is shown in fig. 5. This unit is quite similar to the design used with the 1296-MHz preamplifier, except that a 96-MHz overtone crystal is used in the oscillator. Also, it's desirable to feed the output of the multiplier through another resonant trough which serves as a signal filter. This is shown in fig. 5.

Final tuning should be accomplished with the preamplifier in the system, with the antenna connected. The low-level 2304-MHz signal is then fed in through the antenna.

I obtained a noise figure of 5.7 dB with this circuit, but there is little doubt that a selected transistor can improve this by at least 1 dB. Gain of 7 to 8 dB is

about all that can be expected from this transistor at 2304 MHz.

However, Nippon Electronics supplied a type V578A. This device offers a tremendous improvement at 2304 MHz in the same preamplifier. Noise figure measured at 3.8 dB with the V578A, as opposed to 5.7 dB with the V766A — all with no changes except tweaking. At 1296 MHz, the noise figure was 2.3 dB for the V578A vs 2.8 dB for the V766A.

Soon after the preamplifier was built, I wrote to Fairchild and asked if they would be interested in supplying an MT4578 transistor which, from the published specifications, looked like it

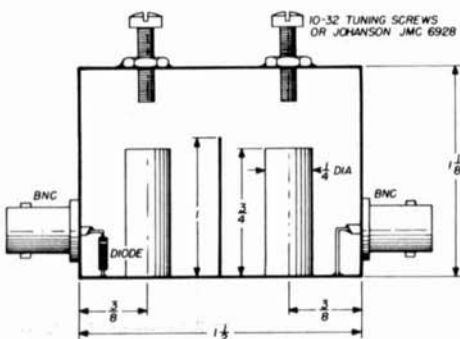


fig. 5. Weak-signal source for tuning up the 2304-MHz preamplifier. Small varactor diode or 1N914 is useful as multiplier, but try anything! Although 10-32 screws can be used for tuning screws, the Johanson JMC 6928 capacitive tuning screw is designed specifically for this purpose, and is virtually noise free. Cavity is 1" deep.

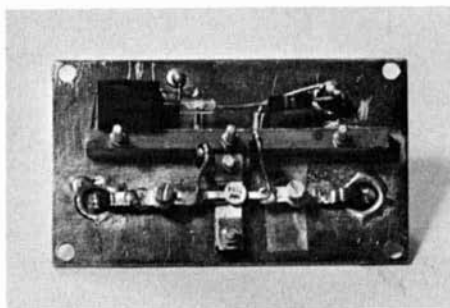
would give improved performance. After waiting some time, I have not heard anything from Fairchild, so I haven't made any experiments with the MT4578.

I wish to express my appreciation to the Johanson Corporation and Nippon Electronics for their cooperation, and particularly to Mr. Harvey Bruning of Johanson whose suggestions were quite helpful.

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1. Dolph Vilardi, WA2VTR, "Low-Noise Transistor 1296-MHz Preamplifier," *ham radio*, June, 1971, page 50.
2. W. Stanton, K2JNG, D. Moser, WA2LTM, D. Vilardi, WA2VTR, "Solid-State 2304-MHz Converter," *ham radio*, March, 1972, page 16.

ham radio



This photo shows the thickness of the partition which is important because it determines the capacitance of the feed-through capacitors, C5 and C6.

# inexpensive audio filters

Surprisingly effective  
audio filters  
are possible  
using non-critical  
readily available  
components

Bill Wildenhin, W8YFB, 41230 Butternut Ridge, Elyria, Ohio 44038

Ed Noll's excellent summary of direct conversion receivers in the November, 1971 *ham radio*, correctly points out the common failing of inadequate audio selectivity.\* In the course of developing some simple, inexpensive equipment for novices, I have found it possible to obtain really excellent selectivity without resorting to rigorous filter design and without careful selection and matching of parts. A sufficient number of these filters, and variations of them, have been made to insure reproducibility.

## basic circuit

Fig. 1 shows the basic circuit — just three constant-K pi sections in cascade. For those who are not familiar with filters, this filter is a low pass filter — it

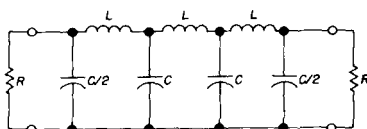
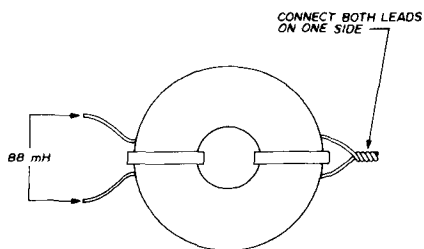


fig. 1. The basic circuit of a three section, constant-K pi-section low-pass filter. Some typical practical values for filters like this are given in table 1.

passes all frequencies from zero to some desired cutoff frequency. The impedance into and out of the filter is fixed by the chosen values of inductances and capacitances, and in this case the impedance in and out is the same value. **Table 1** was made up to allow choice of easily available capacitors. In all cases the inductors are surplus 88-mH toroidal units. **Fig. 2** shows how the leads are connected for use. As supplied, all four leads are usually brought out individually.

The curves in **figs. 5** and **6** are indica-



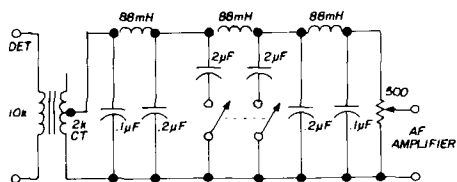
**fig. 2.** Proper way to connect leads to surplus 88-mH toroids for use in simple audio filters.

tive of the results to be expected with off the shelf components. Some filters were built with used tubular paper capacitors pulled from scrap TV sets. Others used new components. If you purchase new components, the Cornell-Dubilier 100 V Mylar capacitors in their inexpensive WMF series are compact and reliable. *In no case should you use electrolytics.*

**Fig. 3** shows a variation that has been successfully duplicated by several fellows.

**table 1.** Component values for low pass filters with different cutoff frequencies. Component values have been rounded off, but are accurate enough to produce excellent results in the circuit of **fig. 1**.

cutoff freq. Hz	R ohms	C μF	C/2 μF
550	150	4.0	2.0
800	200	2.0	1.0
1000	300	1.0	0.5
2300	650	0.2	0.1
2750	750	0.15	0.075

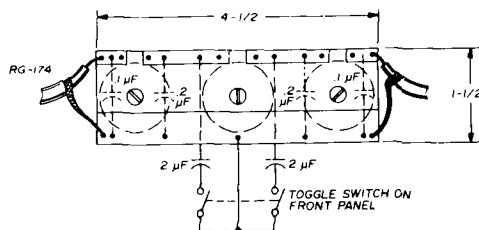
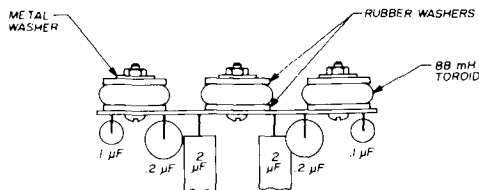


**fig. 3.** Audio filter whose cutoff frequency can be switched between two values.

This filter allows a higher cutoff frequency for phone reception and a low frequency for CW work. A double-pole single-throw slide switch is adequate for this application. The interstage transformer is a low cost import from Radio Shack. The detectors used either a Motorola MC1550 or RCA CA3028 integrated circuit. The 500-ohm volume control terminated the output of the filter and feeds a solid-state amplifier described in *ham radio*, March 1970.<sup>2</sup>

### construction

**Fig. 4** shows a compact suggested layout. For those who have no etching facilities, use Instant etch. No chemicals, no mess. Draw the layout on the copper, and score the copper with a sharp hobby knife. Carefully lift a corner of the



**fig. 4.** Suggested compact layout for the audio filter of **fig. 3**.

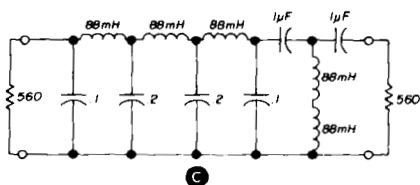
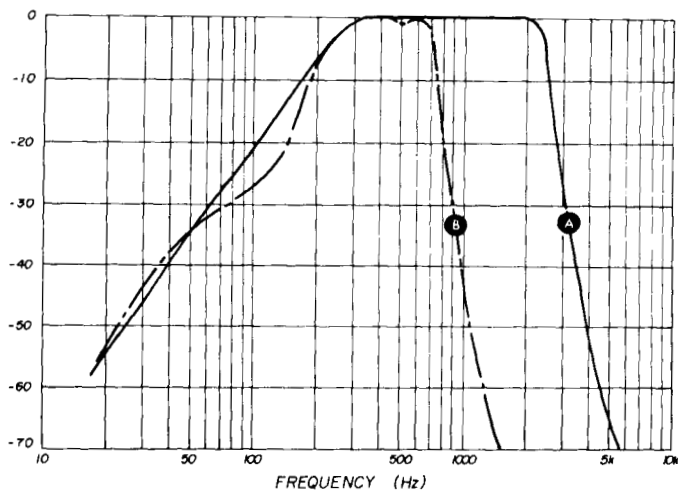


fig. 5. Performance curve of the low-pass filter shown in 5C. Curve A is of a filter similar to 5C, but with all 0.1-mF capacitors increased to 1.0 mF, and all 0.2-mF capacitors increased to 2.0 mF.

undesired copper with a knife edge, and then peel it off with a pair of long-nose pliers. If you wind the copper up as if opening a sardine can the work goes easier. You can also use thin, unplated board material and substitute pieces of

number 22 tinned solid wire for the various busses. When mounting the toroids, it is best to use flexible washers. These can be cut from inner tube rubber, Teflon or soft vinyl plastic. The mounting screw should be covered to prevent

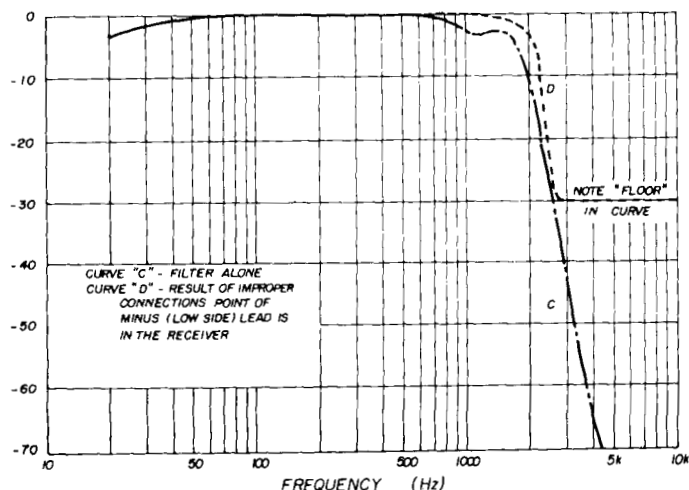


fig. 6. Example of a filter using off-the-shelf components. Also shown is the filter performance degradation due to improper grounding.



damage to the windings. A piece of insulating tubing or a wrap of plastic tape is sufficient. The 2-mF capacitors should connect to the switch with not much more than their lead lengths. The ground return from the switch should *not* go to the nearest chassis ground, but should return to the board, as illustrated. The interstage transformer and 500-ohm pot can be located at some distance from the board. In such a situation, it is best to use a shielded single-conductor cable with an outer jacket to prevent the shield from touching ground at any point except those selected. Miniature coax such as RG-174/U is excellent for this purpose.

### ground loops

The foregoing precautions will help to prevent ground loops which can form paths around the filter. As a matter of fact, if your filter fails to do its job, it might be well to investigate this possibility before condemning the filter. Fig. 6 is an excellent example of what can happen. I plotted curve C using a good lab setup. I installed the filter in an experimental receiver and it performed horribly. Curve D showed that so much signal was being fed around the filter that it was very ineffective. The culprit in this case was the low-side lead from the power supply. The point where it connected into the receiver was such that it formed a beautiful ground loop.

Fig. 7 illustrates a common trouble. The resistance marked R in that figure represents power supply impedance. For these receivers an adequate power supply can be the capacity multiplier job described by W6GXN in the February, 1970 *ham radio*.<sup>3</sup> A poorly regulated supply with an inadequate output filter capacitor may have an impedance measured in ohms. If connected to the receiver with a few long number 22 wires, the lead resistance will contribute significantly to power supply impedance. From fig. 7 you can see that this power supply impedance is a common series element for both the detector and an audio amplifier. A small portion of the audio signal from the detector is fed around the filter by this

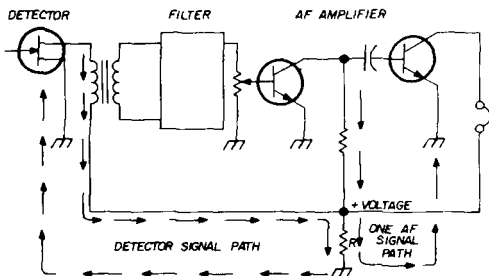


fig. 7. Example of audio filter performance degradation as a signal bypasses the filter by flowing through the common impedance of the power supply.

route, thus degrading filter performance. How much can be tolerated? If a weak signal reaching the first audio stage is 0.1 mV (not an impossible situation) filter degradation will begin to be felt with an unfiltered signal 80 dB below this

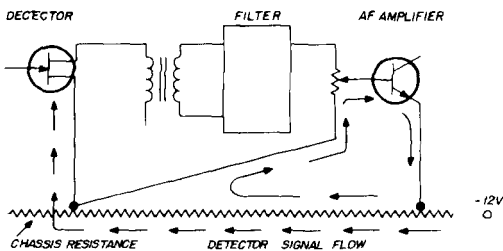


fig. 8. Filter performance is degraded by unexpected chassis resistances providing a signal path around a filter.

level — or 0.01  $\mu\text{V}$ . In all probability, you won't notice any appreciable degradation at this level, but it is starting. A one microvolt sneak signal will produce a fair amount of degradation.

Fig. 8 shows another common mis-

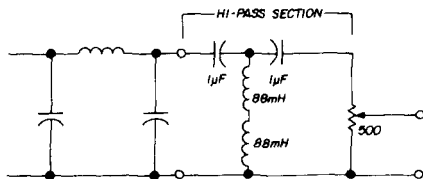
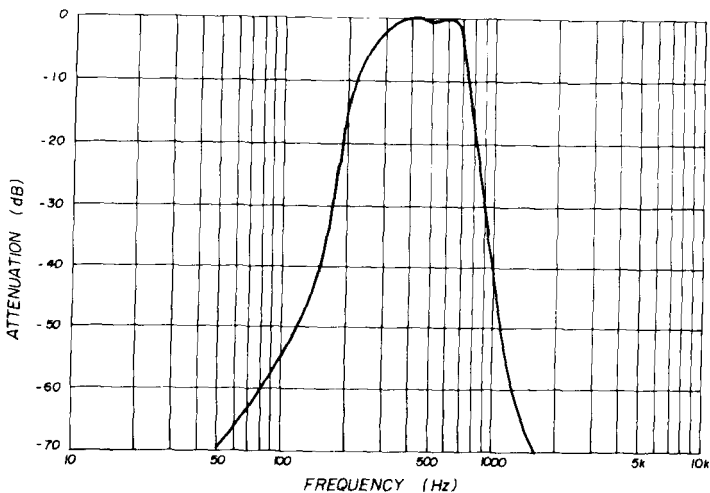
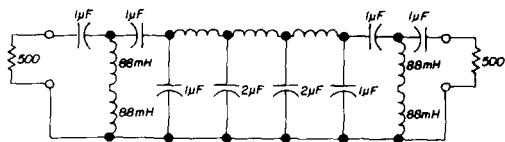


fig. 9. High-pass filter section that can be built from 88-mH toroids.



(A)



(B)

fig. 10. Extrapolated curve to approximate performance to be expected from a high-pass section at each end of a three-section low-pass filter. A typical filter of this type is shown in 10B.

take. The chassis here is represented as a resistance, which in actuality, it is. Particularly with aluminum, pressure connections (a solder lug bolted to a chassis) often show resistance due to the formation of oxides. Where dissimilar metals contact aluminum, oxidation is often accelerated, and aluminum oxide is an excellent insulator. Fig. 8 shows that a part of the chassis resistance is common to both the detector and first audio stage. In this case, it could be that the volume control was located close to the detector to provide a symmetrical panel arrangement. Perhaps a person sees a ground point at the detector and thinks, "a ground is a ground," and fastens the volume-control ground wire there. As the illustration shows, the ground was merely a tap on a resistor.

Don't throw away that aluminum chassis, though! It will still do its job well if you follow a few simple rules. Try to make a straight-line layout of the stages. Avoid a U-shaped layout if possible, because that often brings high- and low-level

ground points close together and leads to instability. In this case it can ruin filter performance. In a tiny 80- to 10-meter direct conversion receiver I built, it was necessary to locate the filter at one remote point and the volume control at another in order to cram the thing together. In a case like this, at audio frequency, it is possible to carry the ground through by using RG-174/U miniature coax. Both the hot and the ground leads were carried to the detector. No ground was allowed below decks on the filter board. The output went by coax to the volume control. Again, the ground side of the control was not physically grounded there. The coax led from the control back upstairs to the audio board. That is where the braid of the coax was allowed to go to ground. The result was a receiver with remarkable selectivity on CW. No filter degradation has been noticed. The last suggestion on grounding is to group the ground points associated with each stage at one point or close together. Don't include grounds for another stage at this

same point, if possible.

### high-pass filter

The last item is **fig. 9** which shows a high-pass filter section that can be built from the same 88-mH toroids. This section passes all frequencies *higher* than some selected cutoff frequency and can be used with the foregoing filters to suppress frequencies below about 250 Hz. **Fig. 10** is not a measured curve, but is extrapolated from other measured curves to give a rough idea of what might be expected from a high-pass section at each end of a three section low-pass filter with the capacitor values indicated. A fairly substantial degree of mismatch has been deliberately introduced in the experimental filters here without much loss in usefulness.

If you want to try out a filter on an existing receiver having a headphone jack fed from a 4-ohm speaker line you could get a fallacious picture of expected performance. In such a case, since you would have audio power to spare, I would suggest connecting an appropriate terminating resistor *in series* with the hot side of the filter input to the hot side of the receiver phone jack. Similarly, if you are using high impedance phones, I would connect a resistor across the output of the filter to properly terminate it there, and then connect the headphones across that resistor so the parallel combination of the phones and resistor terminate the filter in a value roughly similar to that to be used in the final application. This sort of check will give a fair approximation of what to expect if it is built into something, and can serve as a method of determining if a bad feedthrough situation occurs in the finished project.

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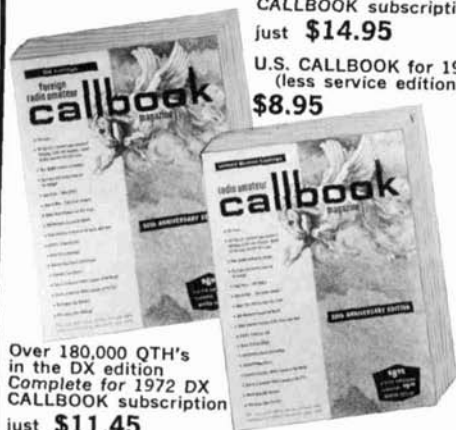
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# n-way power dividers and 3-dB hybrids

This article describes  
n-way power dividers  
and 3-dB hybrids,  
and shows you  
how to design them  
for hf and vhf

Richard S. Taylor, W1DAX, 15 Frederick Street, Belmont, Massachusetts 02178

Although amateurs who operate on the microwave frequencies are familiar with 3-dB hybrids and n-way power dividers, these devices are strangers to amateurs who operate on the lower bands. However, 3-dB hybrids have recently become available for use on vhf and uhf.

A 3-dB hybrid splits the input power equally between two outputs and provides more than 20-dB isolation between the output ports. An n-way divider splits the power n-ways (up to about 10 ways practically, I imagine) with similar isolation between output ports. The outputs of the n-way are in phase, while the outputs of the 3-dB hybrid are 90° out of phase.

Such devices are very useful in paralleling transistor rf amplifiers, for example, or in ssb systems where 90° phase shifts are necessary between two rf signals. By using lumped constant equivalents to the microwave transmission-line circuits, it is possible to design these devices for any amateur hf or vhf band. Circuit bandwidths are typically 10%. One circuit will, therefore, be useful over the full amateur band for which it was designed.

## n-way divider

This is the most straightforward of the two devices (fig. 1). Power coming into the input port is divided equally between

the  $n$  output ports. Power coming in an output port is dissipated in the resistors connected to the common node rather than being reflected into the other outputs. This is the isolation mechanism that makes the circuit so attractive.

The design is accomplished with lumped-constant quarter-wave transmission lines (fig. 2).  $L$  and  $C$  values are determined by the characteristic impedance of the line ( $Z_0$ ) which, in turn, is determined by the number of output ports. That is,

$$Z_0 = \sqrt{NR_0}$$

where  $R_0$  = characteristic impedance of input and output lines and  $N$  = number of output ports.

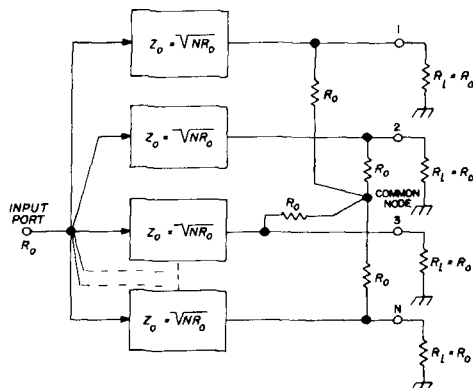


fig. 1. N-way power divider and combiner.

Fig. 3 shows a three-way divider for the 40-meter band (50-ohm transmission line). In this case,

$$Z_0 = \sqrt{3 \times 50} = 86.5 \text{ ohms}$$

Therefore, in fig. 2,  $L = j86.5$  and  $C = -j86.5$  ohms. At 7.2 MHz,  $L = 1.6 \mu\text{H}$  and  $C = 250 \text{ pF}$ . Use a reactance slide rule to determine these values (such as the ARRL R/L/C Calculator) or use the two reactance formulas:

$$L = 2\pi f X_L$$

$$C = \frac{1}{2\pi f X_C}$$

The closer you get to the calculated values, the more accurate the power division and the better the isolation. To obtain 30-dB isolation between output ports, you will have to be within 1% of the calculated values. Five percent com-

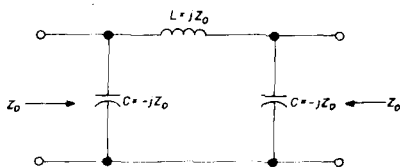


fig. 2. Lumped-constant quarterwave transmission line with a characteristic impedance,  $Z_0$ .

ponent tolerances are probably the highest that can be tolerated without completely ruining output port isolation. Matching components between the paths is more important than obtaining precise values, so if you are unsure of your component tolerances, choose identical, if slightly inaccurate, values and sacrifice input impedance matching for improved isolation.

Turned around, an identical divider would be used to combine amplifier outputs. All the same isolation advantages apply. Total insertion loss through each divider (or combiner) is about 0.3 dB. Advantages of the  $n$ -way divider include equal power division from a straightforward circuit and good isolation between parallel amplifiers. Isolation at the signal frequency means fewer sneak paths to cause oscillation. Also, you can lose one paralleled amplifier without losing all output power.

## hybrid

The 3-dB hybrid is more complex than the  $n$ -way power divider, but it is also more interesting and versatile. It may be used as a divide-by-two power divider in lieu of an  $n$ -way system and will provide similar isolation between the output ports. More interestingly, however, it can be used as a wideband phase shifter in systems requiring quadrature signals over a wide fre-

quency range. In a hybrid, the two output signals maintain a  $90^\circ$  phase relationship over a range of frequencies much wider than the useful 10% bandwidth which is determined from power match-

port 3 divides between ports 1 and 2 with no output at port 4, and so on.

If equal power is applied to ports 1 and 2, then equal power appears at ports 3 and 4 with  $90^\circ$  delay through the

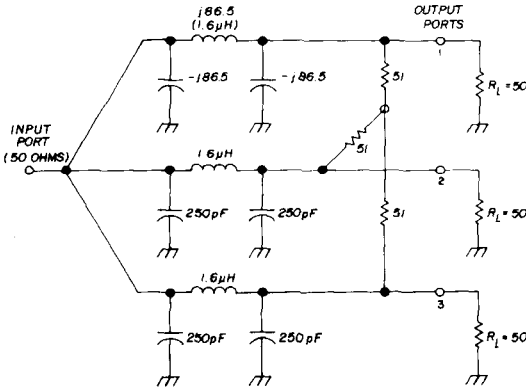


fig. 3. Three-way divider for operation at 72 MHz with 50-ohm transmission lines.

ing considerations. Further properties will become apparent as we discuss the hybrid circuit.

Fig. 4 defines the basic hybrid properties. Power applied to input port 1 divides equally between ports 3 and 4. No power appears at port 2 (which must be terminated by a resistor equal to the characteristic line impedance). The signal at port 3 is delayed by  $135^\circ$ , the signal at port 4, by  $45^\circ$ , resulting in a  $90^\circ$  phase difference between the outputs.

This four-port network is completely symmetrical. Therefore, power applied to port 2 divides between ports 3 and 4 equally, with  $135^\circ$  delay to port 3 and  $45^\circ$  delay to port 4. Power applied to

circuit. However, if the two equal input signals are  $90^\circ$  out of phase, then all the input power appears at either port 3 or 4 depending on whether the input to port 2 leads or lags the input to port 1.

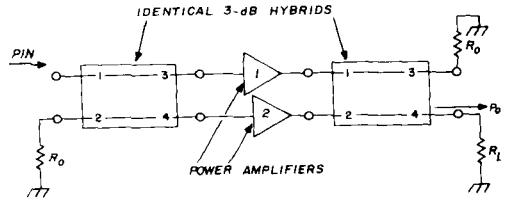


fig. 5. Using 3-dB hybrids as power dividers and combiners.

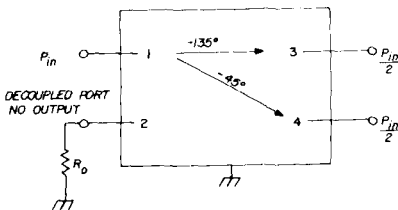


fig. 4. Basic circuit of the 3-dB hybrid.

For power division and combining, a pair of hybrids is necessary (one at the input and one at the output) to maintain the proper phase relationships (fig. 5). Approximately the same signal frequency isolation between power amplifiers is obtained from the hybrid as from an n-way divider. All the power reflected back from amplifier 1 (due, say, to a mismatch) appears at ports 1 and 2 and

not at port 4 where it could feed amplifier 2. Similarly, on the output side none of amplifier 1's output appears at port 2 of the output hybrid, all of it drives the load. The insertion loss through each

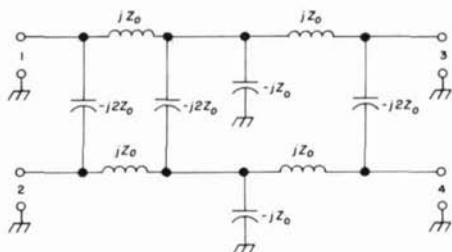


fig. 6. Basic circuit for a 3-dB hybrid for use with transmission lines with a characteristic impedance,  $Z_0$ .

hybrid is 0.3 dB or about the same as the n-way divider.

### hybrid design

Hybrid design proceeds much the same as for the n-way divider. Fig. 6 shows the circuit and gives component values in terms of the characteristic impedance of the transmission lines with which it is used.

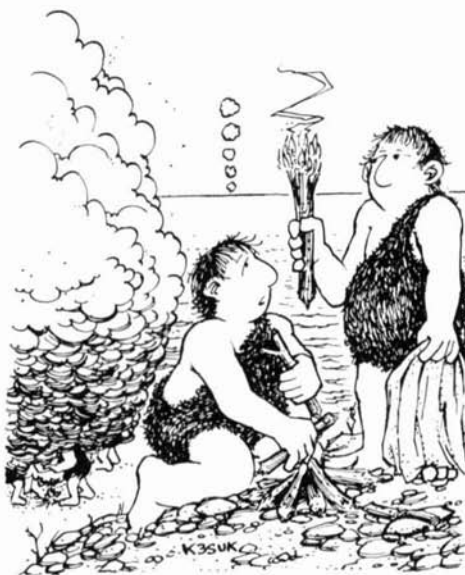
As a design example, a 3 dB hybrid for 50-ohm transmission lines will be developed for a center frequency of 14.25 MHz. I used this hybrid in a direct-conversion receiver so a good amount of test data is already available.<sup>1</sup>

From fig. 6 you can see that all reactances will be either  $Z_0$  or  $2Z_0$  (50 and 100 ohms in this case). These values are included in the circuit drawing in fig. 7. The component tolerances seem to be somewhat less critical here than in the n-way divider but you must still strive to obtain values as close as possible to those calculated. Component matching plays an important part in determining the ultimate isolation possible, so try to match components even if precise values are not available.

The hybrid for the ssb receiver was built with 10% inductors and 5% capaci-

tors. The true center frequency was measured to be 14.45 MHz. At 14.25 MHz, the two output amplitudes were within 1 dB. Amplitude variation between the outputs around the true center frequency was 0.5 dB over  $\pm 0.3$  MHz, which was acceptable. Over 10.3 MHz, the change in the  $90^\circ$  output phase relationship was not measurable. To provide for precise phase adjustment in the receiver,  $C_x$  (fig. 7) was replaced by 150 pF in parallel with the 100 pF trimmer. Amplitude compensation between the outputs was achieved elsewhere in the receiver. Power dividing and combining applications are not as critical as phase shifting and the trimmer might be omitted.

An interesting variation to the 3-dB coupler allows its use in swr measurements. By changing component values, it



"You can't get through a pileup like that . . . yuh gotta better chance out here on the edge."

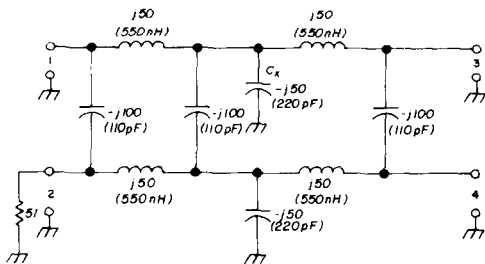


fig. 7. A 3-dB hybrid designed for 14.25 MHz operation with 50-ohm transmission lines.

is possible to change the coupling from 3 dB to 20 dB. With  $P_{in}$  applied to port 1 of this device, 99% is transferred through to port 3, and 1% to port 4. Since ports 3 and 4 are decoupled, the power at port 4 is only 1% of the forward power. Reflected power applied to port 3 is treated in the same fashion. Port 2 output is 1% of the reflected power only. These characteristics describe a directional coupler.

Fig. 8 shows the circuit values required for a 20-dB hybrid, and fig. 9 shows the directional coupler application. Actual component values are determined from the circuit values as for the 3-dB hybrid.

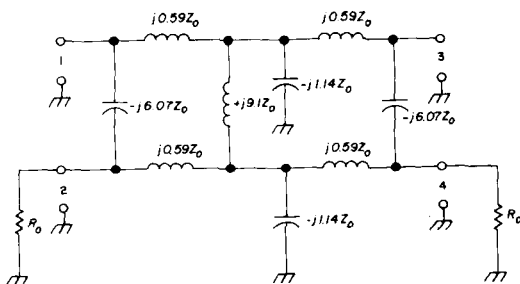


fig. 8. Reactance values for a 20-dB hybrid. These reactances can be converted into component values for any frequency of interest.

### comments

Construct all of these devices symmetrically to minimize stray impedance differences between the paths. In fact, it is best to arrange them like the circuit diagrams. When paralleling power ampli-

fiers, it is well to remember that these dividers provide isolation at the signal frequency only, and additional low-frequency isolation may be required in the power and signal leads to prevent destructive low-frequency oscillation.

Since all ports of these dividers are at transmission-line impedances, it is necessary to include matching networks in each amplifier. This is, in fact, an advantage since each individual amplifier can be tuned up in a 50- or 75-ohm system before you connect them in parallel. When they are paralleled, it will be necessary to retune each of them slightly

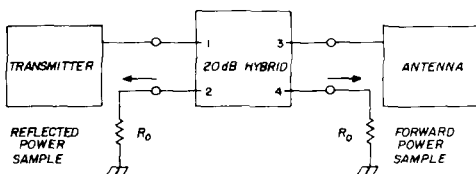


fig. 9. Using the 20-dB hybrid as a directional coupler for measuring SWR.

to match phase shifts throughout the system. Choose one path as a reference and adjust the others to match it.

I have used n-way couplers in vhf solid-state transmitters to achieve output powers of 750 watts at 75 MHz and 650 watts at 225 MHz, the latter in a varactor tripler. Three-dB hybrids have been used in similar transmitter designs to achieve 1,000 watts at 100 MHz and also as quadrature phasing elements in ssb systems.

A computer program was the source of the hybrid component values. It was written by E. A. Johnston, K1YFY, and Bill Higgins of the staff of the MIT Center for Space Research in Cambridge, Massachusetts.

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## phase-shift RTTY monitor scope

By complete analysis  
of received signals,  
this unit  
can be the answer  
to RTTY tuning  
in heavy interference

Elmer E. Mooring, W3CIX, 9318 Millbrook Road, Ellicott City, Maryland 21043

A tuning indicator is almost a necessity to properly tune an RTTY signal. A cross-pattern scope<sup>1</sup> or the ST-5 and ST-6 type tuning meter<sup>2,3</sup> is fine for normal signals; however, with heavy interference, a phase-shift type scope<sup>4,5</sup> is preferable to help sort out the proper mark-space frequencies. The cross-pattern scope or tuning meter uses heavily-filtered discriminator signals from the demodulator for indication, thus filtering out much information about the received signals.

This phase shift monitor scope uses signals directly from the receiver output (or TU input) and displays a rotating line on the face of the CRT. The angle of rotation is a measure of frequency, while the length of trace indicates amplitude; thus, most of the CRT face area contains useful spectrum information. At a glance, you can determine approximate frequency shift of either the received or transmitted frequency, can tune the signal rapidly to correspond to the TU filters and can determine frequency separation of interfering signals. With a little practice, the scope pattern will aid in making the proper adjustment to produce usable copy from marginal signals.

### theory

The heart of the phase-shift indicator is the simple RLC network shown in fig.

1. The series LC circuit is resonant near the mid frequency of interest. Near resonance, the impedance of the LC circuit is minimum; thus,  $E_T$  approaches zero while  $E_L$  is large and exhibits rapidly changing phase about the resonant frequency. Recalling oscilloscope display theory, two sine waves applied to the deflection plates of a CRT produce a pattern dependent upon the phase and amplitude relationship between these sine waves. With the two sine waves either in phase or  $180^\circ$  out of phase, a straight line appears whose angle on the CRT face depends upon the relative amplitudes of the two signals. With  $90^\circ$  phase shift between the two deflection plates, an ellipse appears with an eccentricity dependent upon the relative amplitudes between the two signals.

Referring to fig. 1, the LC circuit is resonant near the center of the frequency range of interest (2550 Hz). For a high-Q circuit, the phase angle varies from near  $180^\circ$  to  $0^\circ$  over a narrow frequency range, which yields straight lines on the oscilloscope tube. Around resonance, an ellipse appears since the signals are around  $90^\circ$  out of phase; however, the width of the ellipse is collapsed to zero since the amplitude of the series voltage ( $E_T$ ) decreases to zero as the phase approaches  $90^\circ$ . Thus a straight line will appear on the CRT for all frequencies of interest and will rotate about the center of the tube face as the frequency varies.

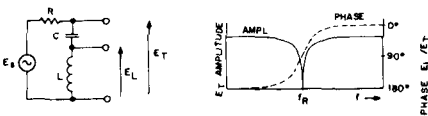


fig. 1. The basic RLC phase-shift network.

Typical patterns for a complete phase-shift monitor scope receiving RTTY signals appear in fig. 2. Most people prefer a vertical line at mid-frequency and clockwise rotation with increasing frequency. The CRT can be rotated in the mount to yield a horizontal or vertical pattern, and connections to one pair of

deflection plates can be reversed to yield the desired direction of rotation.

### circuit description

The complete schematic appears in fig. 3. The design is straightforward. The

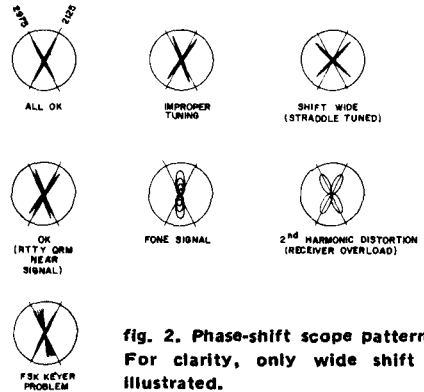


fig. 2. Phase-shift scope patterns. For clarity, only wide shift is illustrated.

input stage (Q1) yields a gain of about 20 (approximately the ratio  $R4/R5$ ).  $R2$  and  $R3$  are bias resistors.  $Q2$  is an isolation emitter follower which reduces  $Q1$  collector loading and offers the current drive necessary for the LC phase-shift network.

The series compensating network ( $C2$ ,  $L1$  and  $R8$ ) could be eliminated; however, it offers two advantages.  $C2$ - $L1$  resonate at approximately 2 kHz to yield an increasing impedance over the 2 to 3 kHz indication range which compensates for the increasing voltage across the phase-shift inductor ( $L2$ ) with increasing frequency. This maintains the same scope trace length at 2125 Hz and 2975 Hz. Another advantage of  $C2$ - $L1$  is that signals far above and below the 2-3 kHz range do not produce a trace on the CRT, thus limiting the pattern to the desired range of frequencies.

$R9$ ,  $C3$  and  $L2$  comprise the phase-shift network described in fig. 1.  $C3/L2$  resonate at 2550 Hz.  $Q3$  serves as an isolating emitter follower to offer a high impedance across  $L2$  and maintain high Q for the network.  $Q4$  and  $Q5$  are identical amplifier stages with a gain of approximately 20 (ratio of collector load to

emitter resistor). Resistors R14/R16 and R18/R19 furnish bias for the transistors.

### construction

Parts layout is not critical. In the several prototype units built, I tried various construction techniques, all with equal success (point-to-point wiring, Vector board with push-in terminals, and

tion of the steel. Of the several prototype units I built, no trace hum was experienced with the exception of an early prototype in which the power transformer touched the CRT socket. This hum was eliminated with a cut-down CRT neck shield (Heathkit 206-180).

You can use a standard CRT bezel to dress up the front panel and for mounting

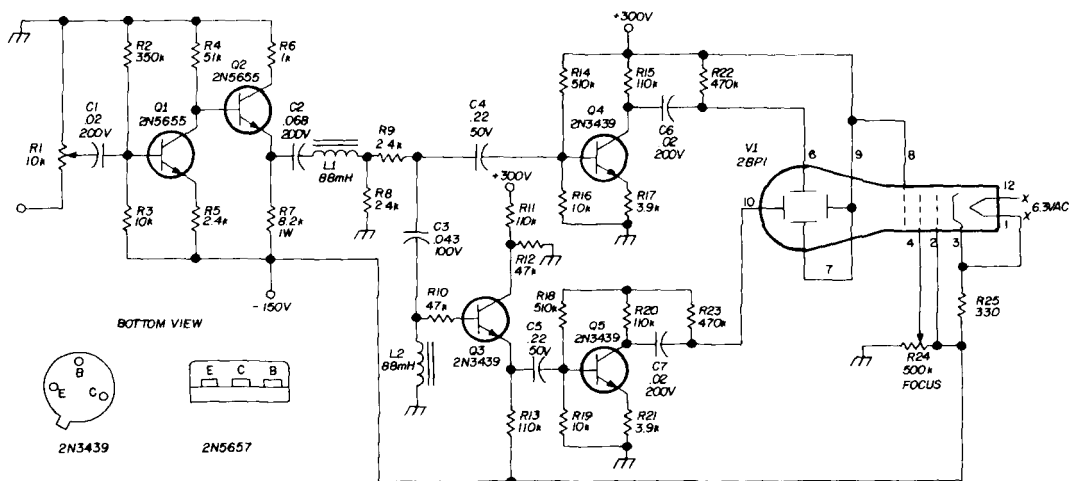


fig. 3. Schematic of the transistorized RTTY monitor scope. Q1 and Q2 may be the same as Q3, Q4 and Q5; there are many substitutes for these transistors. In choosing V1, try to get a CRT with P1 phosphor.

Vector board with Circuit-Stik). Preferably, the power transformer should be mounted to the rear of (and away from) the CRT, and the input amplifier (Q1) should be towards the front of the unit. The only shielded wire is from the input jack to the gain control (R1).

For the two inch CRTs, a 3 x 5 x 9½-inch aluminum chassis is adequate to house the CRT and circuit board. Use an aluminum cover for the hooded top, or alternately, build the unit with the chassis open side down and use a flat aluminum bottom plate. If the power supply is to be built-in, select a chassis size to prevent crowding the power-supply components around the CRT. Otherwise CRT shielding may be required to prevent trace hum. Do not use a steel chassis or bottom plate, since the centering would tend to shift with magnetiza-

tion of the CRT. The unit shown uses an old meter case with the movement and glass removed and the opening filed round. A rubber O-ring slipped into the modified meter case provides a shock mount for the CRT face. The tube neck is supported with an aluminum clamp and spacers. Save space and money by omitting a special tube socket for the CRT. Push pins from a standard old tube socket, solder them directly to the leads and push spaghetti over the pins for insulation (octal socket pins fit the 2BP1, larger pins are required for the 3AP1).

Circuit components are not critical, so feel free to use the junk box; however, C2 and C3 should be high quality (e.g. Sprague Orange Drop) to yield high Q and to minimize calibration drift. Of course, it is an unusual junk box that contains 400 V breakdown transistors. C6

and C7 should have low dc leakage so as to not affect CRT centering. Although not necessary, the gain and Q can be optimized by selecting high gain ( $h_{FE}$ ) transistors for Q2 and Q3. The circuit in fig. 4 can be used to select the two highest gain transistors from a batch for those persons using the same type throughout. The lower the measured

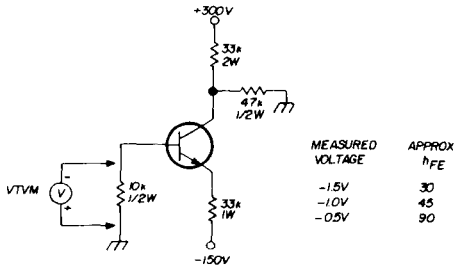


fig. 4. Test circuit for selecting high-gain transistors for Q2 and Q3. The lower the negative base voltage, the higher the Q.

(negative) voltage at the transistor base, the higher the gain. You can use a 20-kilohm-per-volt (or greater) vom in place of the vtvm for selection; however, the  $h_{FE}$  values listed in fig. 4 will be inaccurate (lower voltage reading for equivalent  $h_{FE}$ ). For Q2 and Q3, just choose the units with the lowest measured base voltage.

Minimal power is required and voltages are not critical, thus power for the unit can be taken from other station equipment. The 300-volt load 6 mA and the -150 volt load draws 12 to 14 mA (including CRT). For a built in supply, I used the smallest transformer on hand, a 250 V ct unit and the circuit shown in fig. 5.

Almost any small CRT can be used, such as the 2BP1, 2AP1, 3BP1, 3AP1 or 3RP1. This power supply and circuit has adequate output swing to drive any of them, since the CRT deflection factor is greater with reduced accelerating potential (usually 1000 to 1500 volts). All are sufficiently bright for a well lighted room.

Be certain to check pin connections

and heater requirements for your tube (3AP1 requires 2.5 volts). Don't worry about the spot that finally burns on the center of the tube face during undeflected periods; no useful information is contained in the center of the tube anyway. Centering controls are dispensed with for simplicity. Most CRTs are fairly well centered as is; however, a small magnet can be glued to the CRT neck for centering if desired. A small chip from a dime store magnet is adequate, just slide it around the tube neck until the display is centered on the tube face, and a spot of epoxy or Q-dope will hold it. Flexible black-board magnets are ideal for this purpose.

The phase-shift circuit can be used with a standard test oscilloscope for RTTY monitoring; however, connect the output directly to the deflection plates and not through the scope amplifiers. The additional amplifier phase shift would distort the straight-line pattern.

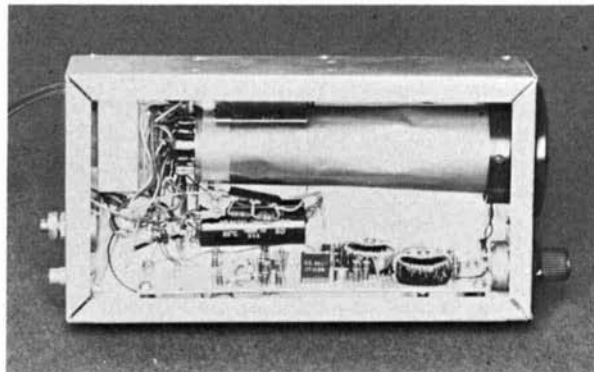
### initial adjustment

At first turn-on, defocus the spot until after the centering magnet is glued on. With power applied and the input sensitivity control (R1) turned down, the following dc voltages should exist:

- Q1 collector = approximately one half of the negative supply voltage
- Q2 emitter = within 1 volt of Q1 collector
- Q3 emitter = -0.5 to -1.0 volts
- Q4 collector = approximately one-half
- Q5 collector = the positive supply voltage

If the readings are off more than 20%, adjust the operating points by changing the value of R2, R14 and R18. Increasing R2 reduces the negative voltage at Q1 collector and Q2 emitter. Increasing R14 increases Q4 collector voltage, increasing R18 increases Q5 collector voltage.

Apply a 2 to 3 kHz sine wave at the input. A straight line should appear on the scope; adjust the amplitude for three-quarter screen deflection. Focus the spot



Interior view of an early prototype using point to point wiring on Vectorboard. Note the power transformer crammed against the CRT socket and filters near the neck of the CRT — both necessitated the neck shielding shown to eliminate a slight trace hum. Input connector and focus pot are visible at the rear, sensitivity pot and power switch are on the front panel.

with R24. As the frequency is varied, the line should rotate about the center. If you prefer rotation in the other direction, reverse the deflection plate connections (pins 6 and 7) at the CRT (or the equivalent pins for your CRT). The line should rotate smoothly from 2 to 3 kHz, with reduced angular rotation and shrinking in size below 1 kHz and above 4 kHz.

Calibration marks are placed on the face of the tube by stretching fine black threads across the tube face (behind the bezel) and adjusting these over the trace at desired frequencies (2125, 2295, 2975 Hz). A spot of Q-dope out of sight at the tube edge will hold the threads in place. If available, a calibrated oscillator or an oscillator and counter may be used for calibration.

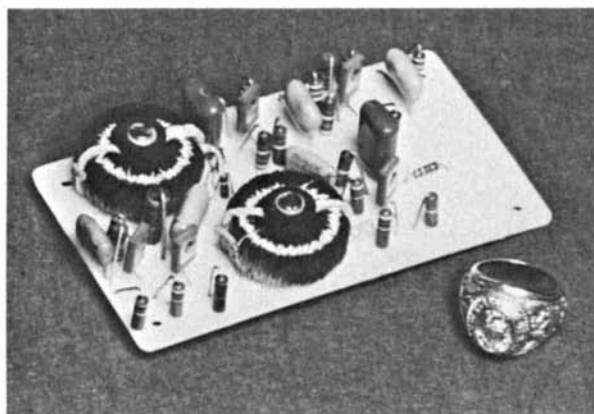
For those using the ST-5 or ST-6 demodulator, the calibration threads can be easily placed right on the filter fre-

quencies. Just connect the monitor scope to the receiver output, turn on the calibrator and bfo and tune for a tone. Tune the bfo for maximum on the TU tuning meter (or maximum measured discriminator test point voltage), and repeat for mark and space on the narrow and wide shift positions of the TU. Adjust the calibration threads over the scope pattern when you obtain the peak discriminator output.

#### parts

The junkbox should supply most resistors and capacitors, but few will have CRTs and 400-volt breakdown transistors.

The power transformer can be a Stancor PA-8416, \$4.66 from Lafayette or the Thordarson 22R39, \$4.37 from Newark. These transformers have a 6.3-volt heater supply which is suitable



The complete phase-shift circuit without power supply. This unit was built using Circuit-Stik and the transistors used were 2N5657 rather than 2N3439 shown on the schematic.

for the 2BP1 and other CRTs. For the 3AP1, you will need a 2.5-volt heater supply. Van's (W2DLT) 302 Passaic Avenue, Stirling, New Jersey 07980, offers unused 3AP1 CRTs in original boxes for \$4.50, first-class postpaid anywhere in the United States. Van also has 88-mH toroids at five for \$2.00. 2N3439 transistors are available for \$1.19 each from General Radio Supply Company, 600 Penn Street, Camden, New Jersey 08102. M. Weinschenker, K3DPJ, Box 353 Irwin, Pennsylvania 15642 offers .22  $\mu$ F 100-volt mylar capacitors at 16 for \$1, and .02  $\mu$ F 200-volt dipped capacitors at 20 for \$1, postpaid.

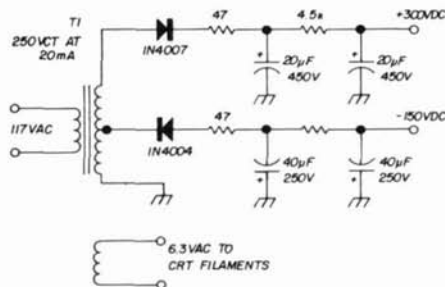


fig. 5. Suitable power supply for the monitor scope.

I wish to thank Bruce Meyer, W0HZR, for his original vacuum-tube phase-shift scope design which served as the foundation for this design. I also want to thank Don, WA6PIR, for urging me to write this article to share the design with other RTTY enthusiasts.

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# crystal oscillator

## frequency adjustment

The frequency of crystal oscillators is often adjusted with a trimmer capacitor — this article explains how much adjustment can be expected

Because of its inherent stability in oscillator circuits as a frequency-control component, the quartz crystal finds much use in circuits where frequency tolerance and stability are important to proper circuit operation. Present transmitter and receiver design trends make good use of this fine frequency-control element.

Much has been said about the necessity for operating the crystal in a circuit suitable to produce the required nominal operating frequency. Although the crystal has an extremely high Q factor and good stability with temperature changes, the circuit in which it is used can, and does, have some measure of control of the final

frequency of the overall oscillator circuit. This is because of the ultimate load capacitance across the crystal terminals in the circuit it is plugged into. For this reason, crystal manufacturers specify this operating load capacitance to ensure that the crystal will control an oscillator within their manufacturing tolerances and specifications.

Of course, good use can be made of the small effect load capacitance has on the operating frequency. You can make slight frequency adjustments by changing the load capacitance with a small trimmer capacitor somewhere in the oscillator circuit. Receivers and transmitters can be set on frequency in this manner. Another use of this frequency change is in RTTY applications where the oscillator frequency is changed for frequency-shift keying.

While these frequency adjustments are often used, not much is usually known in advance as to what magnitude of frequency change can be expected for a given change in load capacitance. All too little has been published in this area, and manufacturers often do not specify this sort of data for their crystals. There are, however, some rather simple relations that apply to most crystals common to amateur use that can be used to get at least a better idea of the frequency changes involved. This is good information to have available, as it allows the selection of a crystal which will produce the desired nominal frequency as well as the necessary adjustment range for final frequency calibration. Also, it allows a lesser tolerance crystal to be selected; variations are compensated for by trim-

Calvin Sondgeroth, W9ZTK, 800 Fifth Avenue, Mendota, Illinois 61342



ming the frequency. A lower tolerance crystal is usually much less costly than one with tighter specifications.

### crystal equivalent circuit

In the course of developing crystals and associated oscillator circuits, early researchers came up with an equivalent circuit for the crystal. This familiar circuit and its impedance vs frequency curve is shown in fig. 1. The circuit consists of an inductance, capacitance and resistance in series with the entire combination shunted by another capacitance,  $C_o$ .  $L_m$  in fig. 1 is the equivalent inductance of the crystal,  $C_m$  is the equivalent capacitance, and  $R_m$  is the effective series resistance.  $C_o$  is a static capacitor formed by the crystal electrodes and the stray capacitance of the holder leads and terminals.

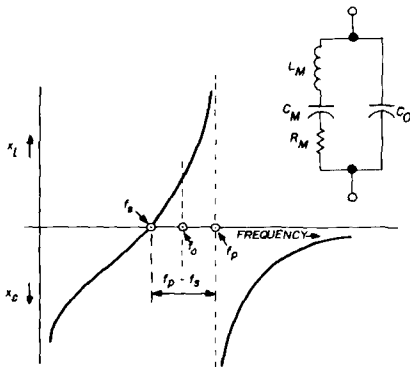


fig. 1. Crystal equivalent circuit and impedance vs frequency curve.

The thing that makes the crystal so stable in an oscillator circuit is that the motional inductance,  $L_m$ , is measured in terms of hundreds of millihenries and has an equivalent  $Q$  of 20,000 to 200,000. The coil in an ordinary tuned circuit might have an inductance measured in microhenries with a  $Q$  of only 100 or less at the same frequency.

The motional capacitance,  $C_m$ , has an extremely small equivalent capacitance in order to be resonant with  $L_m$  at the crystal frequency. These parameters are

termed *motional* inductance and capacitance because they are equivalent values associated with the actual molecular vibrations in the quartz.

Inspection of the circuit in fig. 1 indicates that the crystal has two points of resonance, one of which is the frequency of  $L_m$  and  $C_m$  in series. (This is an approximation, close enough for amateur purposes.) At this series-resonant frequency, indicated as  $f_s$  on the impedance curve, the crystal looks like a small equivalent resistance,  $R_m$ , which is determined by the mounting and preparation of the crystal in its holder.

As the frequency is increased above  $f_s$ , the overall series arm of the circuit takes on the character of an inductance; at some slightly higher frequency the overall circuit, consisting of the equivalent inductive reactance of the series arm and the electrode holder capacitance,  $C_o$ , in parallel, becomes resonant. At this point, the crystal looks like a parallel tuned circuit with a very high  $Q$  and resistive impedance. This frequency is called the parallel or anti-resonant frequency and is indicated by  $f_p$  in fig. 1.

A few oscillator circuits use the series-resonant frequency for operation, but they are usually confined to applications at extremely high or low frequencies. Most of the oscillator circuits found in amateur equipment, Pierce, Colpitts, etc., use the crystal in a parallel-resonant fashion. Since there is some load capacitance offered by the oscillator circuit, the actual operating frequency,  $f_o$ , comes out at a point that is between the two natural crystal frequencies,  $f_s$  and  $f_p$ .

The following discussion deals with relations that can be used to determine what this operating point will be in terms of approximate crystal parameters and added oscillator load capacitance.

### capacitance ratio

An important crystal parameter that is not often mentioned or specified by the manufacturer is one called the *ratio of capacitances*. This is simply the ratio of the crystal holder static capacitance,  $C_o$ ,

to the crystal motional capacitance,  $C_m$ . Since the motional capacitance is very, very small in terms of "real" capacitors, its value for most high-frequency AT-cut crystals does not change much over a fairly wide frequency range. At least its change is very small with respect to the rather large (in comparison) value of  $C_o$ .

The holder capacitance is usually 5 pF or so, while the motional capacitance is often measured in thousandths of a picofarad. Thus, the ratio of capacitances is relatively constant over a wide range of frequencies, and is primarily a function of the type of crystal cut. It would be well to repeat that this discussion applies only to fundamental-mode AT-cut crystals. This is the type most common in high-frequency amateur equipment.

The capacitance ratio, as it is often called, turns out to have a value between two and three hundred. For experimental purposes, you might consider it to be 250. There is a very simple relationship between this capacitance ratio and the natural bandwidth of the crystal. This bandwidth is defined as the frequency spread between the series and parallel resonant frequencies. It is equal to  $f_p - f_s$ , and can be expressed as:

$$f_p - f_s = \frac{f_p}{2rc} \quad (1)$$

If you take the capacitance ratio,  $r_c$ , to be 250, **equation 1** shows that to find the natural bandwidth for a crystal you divide the crystal frequency by  $2r_c$  or 500. It should be noted that this relationship indicates that, for any crystal, the relative bandwidth in terms of percentage of the nominal frequency is the same. For a capacitance ratio of 250, you can expect a bandwidth of the order of 2000 parts per million for any frequency. As you already know, the higher you go in frequency, the more you can trim your oscillators. **Equation 1** gives an indication of why this is true.

As an example, you might have an eighty-meter, fundamental AT-cut crystal at 3.600 MHz. Using **equation 1** you find that its natural bandwidth comes out to be 3,600,000 divided by 500 or about

7200 Hz. This gives a rough idea of the frequency adjustment possible with this crystal, although, as you shall see, the actual amount of adjustment becomes quite a bit less as oscillator load capacitance is added to the crystal.

### effects of load capacitance

In an oscillator circuit of practical design, the crystal becomes loaded with additional capacitance which appears in shunt with the holder capacitance,  $C_o$ . If this additional capacitance is designated  $C_x$ , the following expression gives the amount of frequency spread between the crystal's series resonant frequency,  $f_s$ , and the new parallel resonant frequency of the oscillator,  $f_o$ .

$$f_o - f_s = \frac{f_o}{2r_c} \cdot \frac{1}{1 + \frac{C_x}{C_o}} \quad (2)$$

From **equation 2** you can see that the amount the oscillator frequency is above the series resonant frequency is some fraction of the total crystal bandwidth given by **equation 1**. The fractional multiplier is a function of the ratio of the added load capacitance to the holder capacitance,  $C_o$ . This relationship indicates why the actual amount of frequency adjustment when load capacitance has been added is considerably less than the overall bandwidth of the crystal. To build a practical oscillator circuit, it is necessary to add considerable load capacitance,  $C_x$ , and the manufacturer's specifications often require on the order of 32 pF for nominal frequency tolerance to apply.

Since the holder capacitance is about 4 or 5 pF, the difference between the series resonant frequency and the operating frequency of the oscillator is somewhere around 1/6 to 1/8 the total crystal bandwidth. For the 3.6-MHz example above, the oscillator might be operating as little as 1000 Hz above the series-resonant frequency. Further investigation of **equation 2** indicates that even though you might add considerable capacitance above 32 pF, the actual frequency change will be quite small.

To use equation 2 it is convenient to set up the load capacitance values as integral multiples of  $C_0$ . For HC6/U crystals, the holder capacitance is usually quite close to the 5 pF figure mentioned. This is due to the fact that the electrodes are electro-deposited on the quartz blank and are very close to the same size for any frequency. Likewise, thickness dimensions do not vary too greatly.

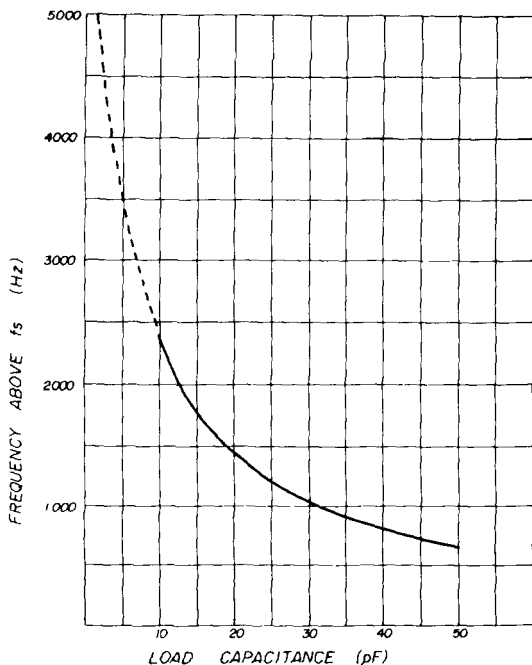


fig. 2. Effect of load capacitance upon the operating frequency of a 3.6-MHz AT-cut crystal. For this graph,  $C_0 = 5$  pF, capacitance ratio = 250,  $f_0 = 3.6$  MHz.

For FT-243 pressure-mounted crystals, the static holder capacitance can be estimated by calculating it from the blank dimensions since the crystal is easily disassembled. The dielectric constant for quartz is about 4.3 and these crystals usually have somewhat higher holder capacitance — on the order of 8 or 9 pF. Thus, it can be seen that they are more difficult to shift in frequency than their wire-mounted plated-blank counterparts. (Their shunt capacitance, and thus, capacitance ratio, is higher.)

## practical values

A graph plotted for equation 2 is shown in fig. 2. The frequency changes on the vertical axis are for the 3.6-MHz crystal used as an example above, and values of load capacitance are plotted along the horizontal axis. The portion of the curve below 10 pF is dotted. This area is normally unusable in practical oscillator circuits since a minimum load must be presented to the crystal to establish the circuit parameters suitable for oscillation. This is why the oscillator frequently quits when you try to turn the trimmer too far out in an attempt to get it on frequency.

There is, of course, some limit to the maximum capacitance that is suitable, but by the time this value is reached, the effects of capacitance change on frequency are practically nil. In fact, by the time you get the circuit oscillating with 10 pF across the crystal, you are down in frequency from  $f_p$  by roughly 2/3 the crystal's natural bandwidth. Further increases in load produce less and less change in frequency.

This is one of the reasons why 32 pF has become somewhat of a standard for crystal calibration. This amount of load minimizes changes in frequency that might occur with unwanted capacitance changes vs temperature, shock and the like. Nevertheless, a 32-pF load allows a nominal range of adjustment for final frequency calibration.

It should be pointed out that fig. 2 is plotted for a holder capacitance of 5 pF and a capacitance ratio of 250. Actual values will be somewhat different for different crystals, but these are good figures for estimating how much frequency adjustment can be expected at a given nominal operating frequency.

Going back to the 3.6-MHz crystal as an example, you can see from fig. 2 that if it is going to be on nominal frequency at say, 30 pF load capacitance, you might expect an adjustment range of about 1680 Hz. However, in going from 30 pF to 50 pF, you would get a change of about 360 Hz, while going the other way from 30 pF down to 10 pF, would

produce a change of some 1320 Hz. This indicates quite clearly what amateurs have frequently taken advantage of — it's easier to move a crystal up in frequency with the trimmer than down.

For some circuits, it might be well to obtain crystals which are a bit low (specified at 32 pF) and depend on the trimmer to move them up on frequency. The equations and fig. 2 can be used to get a fair idea of where the nominal frequency should be specified.

Bear in mind however, that the data presented here is based on a couple of approximations; i. e., assumed values for the capacitance ratio and holder capacitance. The latter can usually be calculated or measured to within rather close limits. And for "worst case" calculations, a value of 300 for the capacitance ratio might be a better value to use. This would ensure that adequate frequency adjustment range would be available for a given crystal.

### actual measurements

As a rough check on some of the relations presented here, I set up the oscillator circuit shown in fig. 3. The equivalent load capacitance offered to the crystal is about 20 to 35 pF as the 30-pF trimmer is adjusted through its range. To simulate the 3.6 MHz example calculations above, three HC6/U crystals in the 80-meter band (3.5 to 3.6 MHz) were tested. For a capacitance change of 20 to 35 pF, the measured frequency changes were 360, 455 and 510 Hz, indicating a fairly wide spread among different crystals at the same approximate frequency. Equation 2 shows that slightly more (about 550 Hz) adjustment should be possible, indicating that a value of 300 for the capacitance ratio would probably be a better choice. In addition, it would be well to try to obtain a value for the holder capacitance — either from the manufacturer or by actual measurement. These crystals measured very close to 5 pF  $C_0$ .

### conclusions

Because of its large value of motional

inductance and extremely high Q factor, the crystal is a good component to use as a frequency-control element in oscillators. It is interesting to note that the impedance vs frequency curve in fig. 1 is drawn out of scale. Below the series

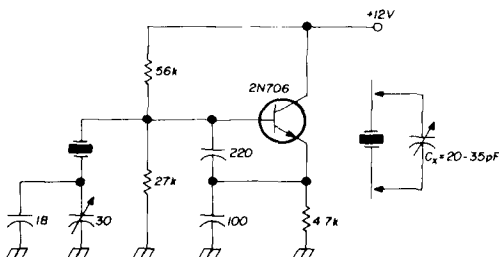


fig. 3. Oscillator test circuit.

resonant frequency of the crystal the capacitive reactance (of the holder capacitance) is much smaller than the reactance values between series resonance and the parallel-resonant frequency.

Likewise, the frequency scale is expanded a great deal around the crystal frequency to indicate the crystal's bandwidth. If the reactance axis were drawn to scale, the inductive reactance approaching parallel resonance would go off the paper. It is this rapid change in reactance with minute frequency changes that gives the crystal its ability as an excellent frequency controller.

Although the crystal is extremely stable, small changes in load capacitance in the oscillator circuits can provide for slight adjustment of the oscillator frequency. Equations 1 and 2 show the amount of adjustment that can be obtained once some crystal parameters have been either measured or estimated.

While these calculations are not exact, they give good approximations for the amounts involved for a given frequency of operation. The accuracy, of course, is limited by the inability to establish true values for these parameters. With exact values, the relations in equations 1 and 2 are exact.

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## direct-reading capacitance meter

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and compact packaging  
are featured  
in this  
useful addition  
to your test bench

R. W. Johnson, W6MUR, 7511 (9) Clay Avenue, Huntington Beach, California

A direct-reading capacitance meter is a particularly useful test instrument. With it you can check capacitors whose markings have worn off or whose tolerance is unknown. Many ceramic capacitors, for example, are GMV (guaranteed minimum value), and their actual value may be as much as ten times the value marked. A direct-reading capacitance meter is also more convenient to use than a bridge when determining capacitance changes with temperature.

The instrument described can measure capacitance in five decades, from 100 pF through 10  $\mu$ F, to an accuracy determined mainly by the calibration standard used and the accuracy of the meter movement. Five percent accuracy is easily attainable. The parts cost is under \$20. The instrument will operate from any unregulated power supply reasonably free of ripple and capable of delivering 28 volts at 50 mA.

### block diagram

A reference sawtooth oscillator (fig. 1) determines the over-all period of a pulse whose width is linearly proportional to the unknown capacitance,  $C_X$ . A sawtooth oscillator was chosen because of

This circuit is reprinted from *Electronics*, Sept. 13, 1971; copyright McGraw-Hill, Inc. 1971.

the simplicity of switching ranges using only one RC time constant. Through an emitter follower (to avoid loading the reference capacitor,  $C_R$ ), the sawtooth operates a Schmitt trigger, which changes state at a fixed and repeatable voltage level part way up on the sawtooth as the reference capacitor charges. A shaper-driver operates a discharge transistor to discharge the unknown capacitor,  $C_X$ ,

or the other of the two inputs is positive and no gate output occurs. As soon as the second Schmitt trigger,  $ST_2$ , fires part way up on the charging waveform for  $C_X$ , the NAND gate output again returns to zero.

Thus the time between leading edges of the NAND gate output is the same as the period of the reference oscillator, and the width of the NAND gate output is

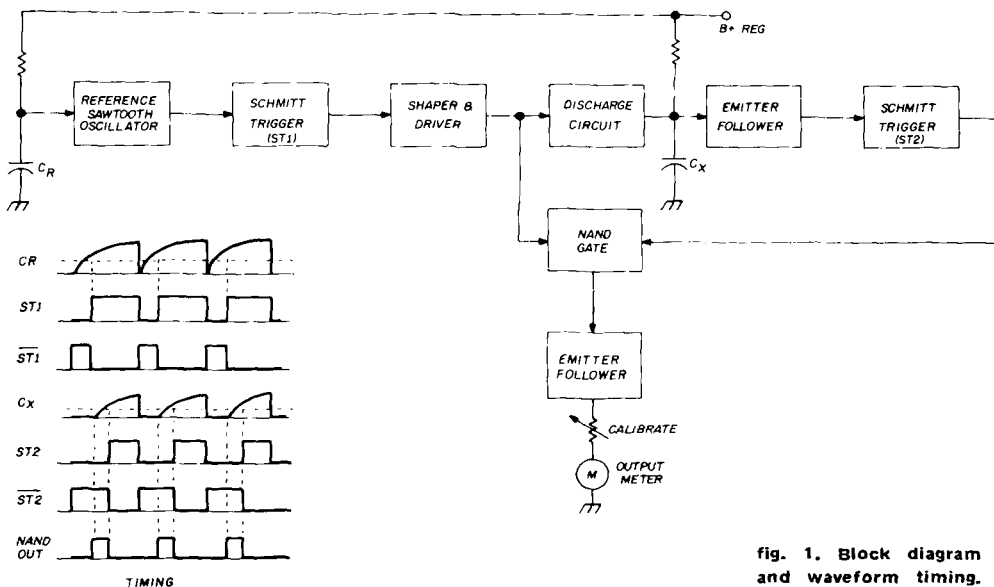


fig. 1. Block diagram and waveform timing.

and also furnishes one input of a NAND gate. The charging voltage across  $C_X$  is applied through an emitter follower to another Schmitt trigger, whose output is the second input of the NAND gate.

$C_R$  is the reference capacitor of the sawtooth oscillator. When the first Schmitt trigger,  $ST_1$ , resets at the end of the reference period, the unknown capacitor is discharged and clamped by the discharge circuit at almost zero volts until  $ST_1$  triggers again part way up on the reference sawtooth (see timing diagram). At this time the unknown capacitor,  $C_X$ , charges through a precision resistor of value switched by a range switch. During this  $C_X$  charging period (only) an output from the NAND gate occurs since both of its inputs are negative. At other times one

linearly proportional to capacitance  $C_X$ . If this pulse is applied to an average-reading meter, the meter will read a current linearly proportional to the unknown capacitor,  $C_X$ . No special scale is required on the meter.

### design considerations

The main reason for using the NAND gate over other possibilities was to make the meter read zero when  $C_X$  is disconnected; the  $ST_2$  circuit is triggered for this condition.

Fig. 2 is the schematic for the capacitance meter. While this circuit seems to have a lot of transistors, it should be remembered that the RCA CA3046 IC array is used, which has five transistors in a single 14-pin dual in-line package. So

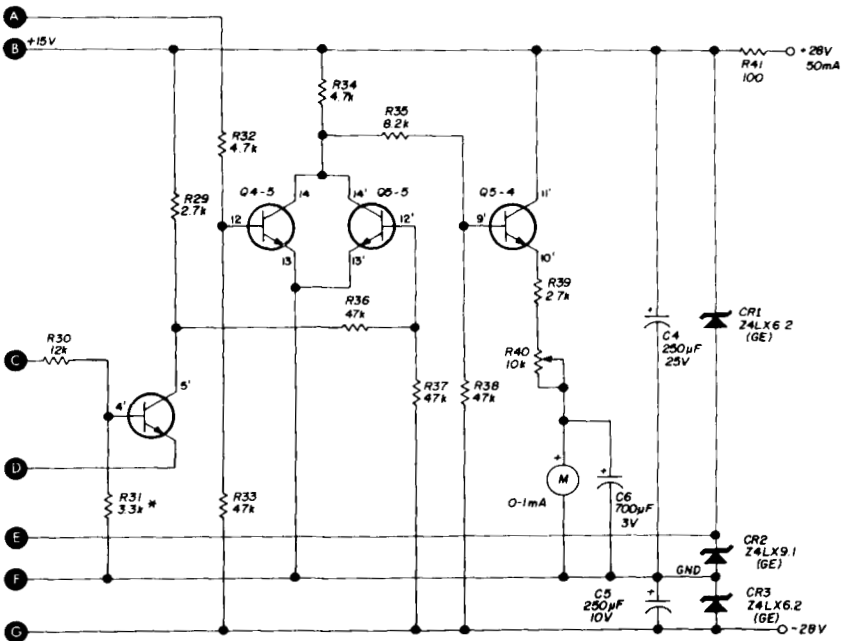
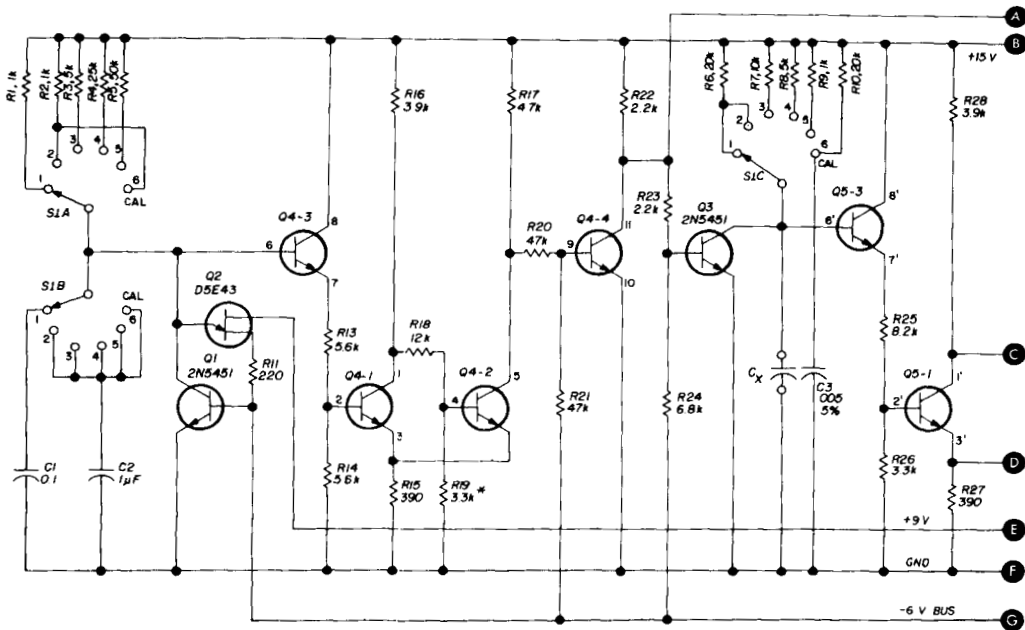


fig. 2. Circuit schematic. RCA type CA3046 IC arrays are used for Q4, Q5, resulting in compact packaging. Resistors R1 through R10 are 1% tolerance. R26 is 3.3k nominal (may be 2.7k or 3.9k). Sometimes it is necessary to change R19 to 12k to make the first Schmitt trigger work. Also, due to variations in the components used, changing R7 to 9720 ohms and R8 to 4875 ohms will probably improve accuracy on ranges 3 and 4. R19 and R31 may need to be increased because of CA3046 gain.



ten of the transistors are simply two CA3046s, labeled Q4 and Q5 in fig. 2.

One of the more troublesome problems in developing this capacitance meter was getting a good unijunction sawtooth oscillator without having to resort to expensive UJTs such as the 2N494. In the ordinary UJT sawtooth oscillator circuit, the range of resistance values that can be used is limited. The charging resistor value can't be too low or the UJT will sustain conduction at the valley current after it has fired once and won't oscillate. The charging-resistor value can't be too high or leakage will cause calibration nonlinearity and, in the extreme case, oscillations will cease because the voltage across the capacitor won't build up to the UJT peak point. In the present case it's desirable to use fairly large values for the charging capacitor to eliminate other problems, and at the same time fairly high frequencies are required; thus a low value of charging resistor is necessary, and the ordinary circuit simply will not operate.

This problem was solved by adding an inexpensive high-current transistor to help discharge the reference capacitor. In fig. 2 the UJT is Q2, and Q1 has been added to furnish a more rapid and complete discharge of C1 or C2. The firing

point of Q2 determines the period of the oscillation for a given RC time constant and fixed supply voltage. As soon as Q2 triggers it furnishes base drive to Q1, which then discharges C1 or C2 with a peak discharge current of around 400 mA. Either capacitor is discharged in a few microseconds to a voltage below the valley voltage of Q2, so Q2 is forced to

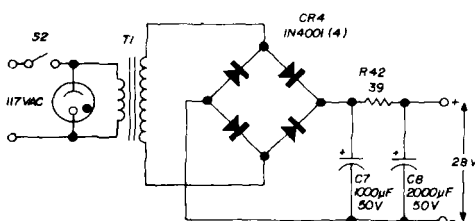


fig. 3. The capacitance meter power supply.

cease conducting despite the low value of charging resistance.

The effect of leakage is also reduced in the oscillator circuit by returning Base 2 of Q2 to a regulated voltage less than the charging voltage. This lowers the UJT peak point; and while the output amplitude is reduced and the waveform has a dc level above ground, the sawtooth waveform is much more linear. Hence the effects of leakage shunting C1 or C2 are greatly reduced.

### circuit analysis

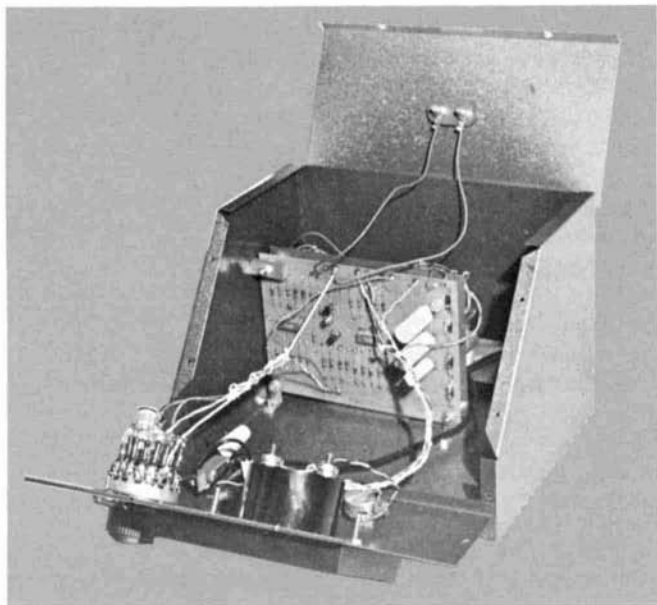
The sawtooth generator output appears across the emitter load of emitter follower Q4-3. Voltage divider R13,R14 sets the firing point in the reference sawtooth waveform at which Schmitt trigger Q4-1, Q4-2 changes state. When the voltage at pin 2 of Q4 (junction between R13 and R14) is low Q4-2 conducts, Q4-4 is cut off, and Q3, the discharge transistor for  $C_X$ , conducts. The voltage across  $C_X$  is thus clamped during this time to the  $V_{ce(sat)}$  of Q3. Part way up on the reference sawtooth, the voltage at pin 2 of Q4 rises to the trigger point, whereupon Q4-2 swings to cutoff, Q4-4 conducts, and Q3 is cut off.  $C_X$  begins to charge at this instant since it is no longer clamped by Q3.

- C1 0.1  $\mu$ F 50 V Mylar (CD WMF05P1)
- C2 1.0  $\mu$ F 50 V Mylar (CD WMF05W1)
- C3 0.005  $\mu$ F 5% Polystyrene (CRL CPR-5000)
- C4 250  $\mu$ F 25 V (Mallory MTV250DN25)
- C5 250  $\mu$ F 10 V (Mallory MTV250CP10)
- C6 700  $\mu$ F 3V (Mallory MTV700DJ3)
- C7 1000  $\mu$ F 50 V
- C8 2000  $\mu$ F 50 V
- CR1 GE Z4LX6.2 1-watt zener 6.2 V
- CR2 GE Z4LX9.1 1-watt zener 9.1 V
- CR3 GE Z4LX6.2 1-watt zener 6.2 V
- CR4 (4) Motorola 1N4001
- M1 0-1 mA meter (Simpson Model 524)
- Q1 TI 2N5451 (800 mA)
- Q2 GE D5E43 unijunction
- Q3 TI 2N5451 (800 mA)
- Q4 RCA CA3046 5-transistor IC array
- Q5 RCA CA3046 5-transistor IC array
- S1 Range switch, 2 pole 6-position (Mallory 3236J)
- S2 Toggle switch
- T1 24 V filament transformer

As  $C_X$  charges toward the regulated power supply voltage through a resistance determined by the range switch, a point is reached where the voltage of  $C_X$  is high enough to trigger the second Schmitt trigger, Q5-1, Q5-2. At this instant Q5-2 cuts off, the voltage applied to the NAND

control, which is set so that the meter reads a value corresponding to a known capacitance applied to the  $C_X$  terminals. The calibration circuit is R10, C3, which may be switched in at position 6 of the range selector. C3 was chosen to give a mid-scale reading on the 0-0.01  $\mu F$  range

Inside the direct-reading capacitance meter. The printed-circuit board, with or without semiconductors mounted and soldered in place, is available from the author. Choose value of R42 for 28-volt output.



gate again becomes positive, and Q5-5 saturates. Since Q5-5 and Q4-5 share the same collector resistor, R34, it is apparent that both these transistors must be simultaneously cut off (base negative) for any voltage to appear at the common collector junction with R34. So only during the time when  $C_X$  is charging and Schmitt trigger Q5-1, Q5-2 has *not* triggered, is this condition met. Hence the positive pulse that appears at the common collectors of Q5-5 and Q4-5 has a width exactly proportional to the value of  $C_X$ . This voltage, the gate output, energizes the meter through emitter follower Q5-4. Capacitor C6 provides additional averaging and was chosen for the lowest frequency range (highest capacitance of  $C_X$ ) to prevent meter flutter.

### calibration

Potentiometer R40 is the calibration

(range 2). The unit must be calibrated with no capacitance connected across the  $C_X$  terminals. C3 is a 5-percent capacitor; but if a 1-percent capacitor is used, or one whose value has been determined by an accurate bridge, the over-all accuracy of the meter will be improved.

### range switching

The range switch uses resistors rather than capacitors, except for the lowest range, because 1-percent resistors are less expensive than 1-percent capacitors. C1 and C2 are the only capacitors switched in the ranging circuit. Their exact value isn't important (e.g., their tolerance can be 10 or 20 percent; but C1, C2 must be reasonably stable with temperature and time and have low leakage). Their ratio should be 10:1. Since C1, C2 will not, in general, have this ratio, it's necessary to trim R1 with an additional series or shunt

resistor so that a known 0.001- $\mu$ F capacitor, say, when set to read 0.1 mA on M1 when on range 2 will read 1.0 mA when on range 1. In the constructed unit, R1 was trimmed by shunting a 12k  $\frac{1}{2}$  watt 10 percent resistor across it.

## leaky capacitors

These components will show a higher apparent capacitance than their true value. However, the leakage has to be pretty bad for this to occur because the maximum charging resistance in the  $C_X$  circuit was kept deliberately low — a maximum of 20k on the lowest capacitance range to 1k on the highest range. If the capacitor to be measured shows less than about 10 times the charging resistance for the range in use, when measured with a good ohmmeter, some error will occur in the capacitance measurement. This much leakage would be unacceptable in most circuits anyway, so the capacitor should be discarded.

The five ranges are (in  $\mu$ F):

Range 1	0-0.001
Range 2	0-0.01
Range 3	0-0.1
Range 4	0-1.0
Range 5	0-10.0

In Range 1 the lowest value of capacitance that can be accurately measured is about 100 pF, or about 10 percent of full scale on the meter. Stray capacitance in the leads to the selector switch, etc., will cause the meter to indicate slightly on Range 1 with no capacitance connected to the  $C_X$  terminals. Also on this range, the reference frequency sawtooth is fairly high; and even though the discharge time of C1 is only a few microseconds, it is not an altogether negligible percentage of the sawtooth period. This, together with other effects such as finite rise and fall times in the waveforms, limits the minimum capacitance that can be accurately measured.

## power supply

The power supply schematic shown is

\*Those interested may obtain the printed circuit with or without all semiconductors installed. Write the author for details.

used in the unit pictured, but it can be any unregulated 28-volt supply delivering 50 mA with reasonably low ripple. Zener diodes CR1-CR3, inclusive, compensate for line voltage variations to which the instrument would otherwise be sensitive.

## construction

The direct-reading capacitor meter is mounted in a Bud C1585HG sloping panel cabinet. The unknown-capacitance terminals are a pair of binding posts on top. The main circuit is on a 4 $\frac{1}{4}$  x 6 inch, G-10 epoxy etched circuit, with two-sided etching and plate-through holes.\* The power supply is mounted on a piece of printed circuit material the same size as the main circuit, and bolted to it by spacers. The assembly is then secured inside the Bud case by right-angle brackets.

## conclusion

This instrument has been useful in checking unknown capacitors from the junk box and from surplus equipment. A surplus audio filter tuned to 1050 Hz, for example, contained many unmarked capacitors that had to be measured to decrease the filter frequency to around 400 Hz for more comfortable cw listening. Dried-up electrolytics, supposedly 10 or 8  $\mu$ F, were found to be less than 1  $\mu$ F when measured. Some ceramics read as much as ten times their GMV value. It was also interesting to study the temperature drift of various capacitor types by watching the meter move as a soldering iron was held near one lead of the capacitor or as cold chemical spray was applied.

Finally, if you need a more precise measurement than is possible with the Simpson 524 (i.e., when matching two capacitors for an equal value or a given ratio), an oscilloscope can be used to measure the duration of the NAND gate pulse appearing across R39 in the circuit. If you have a burning desire for a digital readout, the NAND gate pulse width can be read with a time-interval meter and Nixie or LED displays.

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# oscilloscope voltage calibrator

This simple  
voltage calibrator  
increases the utility  
of your scope  
with calibrated  
voltage-reference points

As anybody who regularly uses one will agree, the oscilloscope is one of the most versatile electronic instruments available and in many situations is indispensable. A scope may be a relatively simple general-purpose device or a greatly refined and complicated, high-precision, laboratory instrument. Its cost may run from considerably less than one hundred dollars to many hundreds of dollars. Its most important use, of course, lies in the study of alternating-current waveshapes; whether such waveshapes are regular in form, as in sine, sawtooth, square, trapezoid or pulse-train shapes; or irregular, as in speech, clipping, video, integrated sync pulses, transients or other distorted-type forms.

In addition to actual waveshape observation, a very useful application of the scope is the measurement of the voltages of waveshapes for which averaging-type meters, such as multimeters and vtvm's, are unsuitable. This usage requires that the scope be calibrated as a voltmeter; this article describes an accessory which will make less-expensive scopes more useful in this respect.

There are many thousands of general-purpose scopes in use. Most of them do not have a voltage calibrator as an integral part of the instrument. If a calibrator is included it consists of a single calibration

F. Everett Emerson, W6PBC, 1709 Notre Dame Avenue, Belmont, California 94002

voltage of doubtful accuracy. By the use of the simple accessory described here, the utility and voltage measuring accuracy of these scopes may be greatly enhanced. An accurate square wave is ideal for calibration usage. This accessory develops such a waveform.

### circuit

The schematic, fig. 1, is straight forward. Parts are easily obtainable and no trick circuitry is involved. It works beautifully and can be assembled in one evening's time. Parts placement is in no way critical. A shielded box, such as a small LMB box, not only makes a nice package, it will also prevent noise pickup from reaching the scope. The switched outputs give square-wave voltages of 1, 2, 3, 4 and 5 volts peak-to-peak.

The rectifier diode CR1 is specified at 1 amp so that it will handle the charging current for the electrolytic capacitor. The zener diode assures a constant and stable dc reference voltage for fluctuating line voltages. Nearly any silicon diodes may be used for CR2, CR3 and CR4 but all three should be the same type. The circuit current is normally less than one mA.

Resistors R1, R2 and R3 may be 10% values (silver band). Resistors R5, R6,

R7, R8 and R9 do not have to be precision resistors. Ordinary, good quality 1/2- or 1/4-watt carbon composition resistors, 5% tolerance (gold band), will do very nicely for the voltage divider. The potentiometer, R4, should preferably have a linear taper although an audio taper is usable.

Because a stable dc reference is used, this potentiometer may be mounted inside the box since once it is set it should require no further attention. The switch should be a non-shorting type. Suit your personal preference regarding the output jack, but since the lead from the calibrator to the scope should be shielded, a BNC or good phono type should be used. The use of shielded lead (preferably coax) will prevent hum modulation of the calibrator's square-wave output.

### calibration

The test point TP is used when adjusting the calibrator output voltages. Use a good vtvm on its 3- or 5-volt dc range (positive to chassis, negative to TP). Adjust R4 to give -2.65 volts dc, as accurately as you can, at the test point. Your calibrator will then be calibrated for a 5-volt peak-to-peak square wave. Switch positions will provide one-volt steps; thus, peak-to-peak outputs will be 1.0, 2.0, 3.0, 4.0 and 5.0 volts.

On dc-coupled scopes there will be a perfectly flat top and bottom to the wave. On ac-coupled scopes you will notice a very slight slant to this line; on such scopes a reduction of the horizontal gain will reduce its effect on the visual accuracy with which you can read the calibrated lines on the scope.

With the usual X1, X10 and X100 step attenuator positions on most scopes, your scope gain control will permit you to set a wide range of accurately known voltage values for any scope division you may wish; such as 1.0 volt, 0.1 volt or 0.01 volt, or perhaps 4.0 volts, 0.4 volt or 0.04 volt per centimeter or per inch or per any other division marking you may desire.

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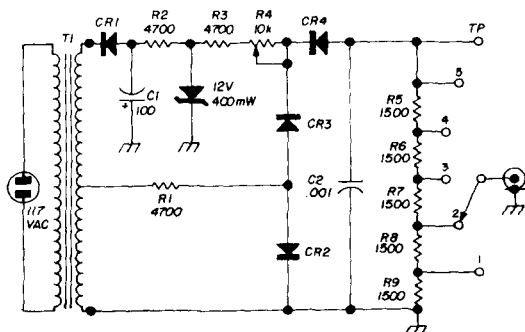


fig. 1. Oscilloscope voltage calibrator provides outputs of 1, 2, 3, 4 and 5 volts peak-to-peak. Resistors R5, R6, R7, R8 and R9 should have 5% tolerance. Diode CR1 is 100 PIV, 1 A. Diodes CR2, CR3, and CR4 are silicon diodes with 1-mA rating (1N4005, 1N914, etc.).

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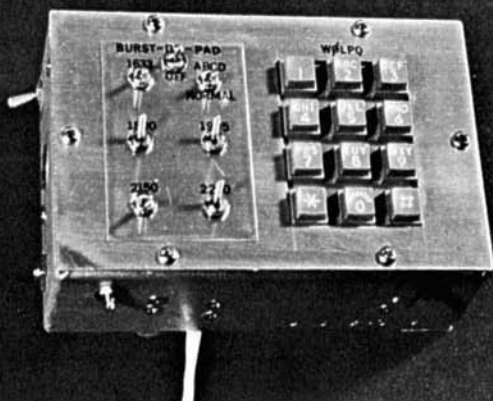
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## mobile operation with the touch-tone pad

Making a clean  
and dependable  
mobile fm  
installation  
with a touch-tone pad

William P. Lambing, WØLPO, 1480 - 25th Street, Marion, Iowa 52302

Numerous articles have been written concerning the utilization of Touch-Tone for control functions on vhf and uhf repeater systems, even going so far as to include partial control of an existing hf system via the uhf radio link. However, articles that deal with actual construction and operation of these Touch-Tone pads are few.<sup>1,2</sup> This article hopes to fill the void, and help some enterprising amateurs build operational Touch-Tone systems.

The Touch-Tone pad enclosure shown in fig. 1 has been in use for over a year with great success and much satisfaction. My unit is being used with the Regency HR-2. Both the HR-2 and the HR-2A are quite popular in this area, and to date, no one has experienced any difficulty with this circuit. Over the last year there have been some basic design changes, most of which are incorporated in this article.

### the pad

The Touch-Tone pad I used is the Western Electric Model 35A3. I understand the Model 35Y3 is basically the same, only a newer model. The schematic diagram shows color coding of each wire on the 35A3 and to what point each wire



goes. Note that the *red* lead is not used, and is taped or clipped off. The *green* lead is the power and audio take-off point. There are two leads left that are rather unusual. They are the *white* and the *white and blue* leads. Both leads terminate in the pad at a switch contact which is shunted by a 5.1k resistance. When *any* button on the pad is depressed, the switch contacts open giving this 5.1k dc resistance. Both leads are free from any other contact or connection within the pad, leaving them free for control of an external circuit, which, in our case, will be push-to-talk keying of the vhf transceiver.

## operation

*The keying circuit:* The two leads (*white* and *white and blue*) from the pad are fed to a transistor switch. One lead (*white*) is grounded, and the other (*white and blue*) is tied to the base of Q1 (2N2222A, 2N718A or other npn). With the pad in an idle condition, the base of Q1 is grounded and is cut off. When any button on the pad is depressed, the base is returned to ground through the 5.1k resistance in the pad. This resistance plus the external 2.2k resistor form a voltage divider, forward biasing the transistor switch, Q1.

*The relay:* The relay is a subminiature dpdt type with a holding current of about 30 mA. Both sets of contacts are used, one for push-to-talk keying of the vhf transceiver, and the other to toggle the microphone audio of the transceiver between the microphone and the output of the Touch-Tone pad. Breaking the microphone connection insures that no background audio can reach the transceiver microphone input causing distortion of the Touch-Tone tones.

Across the relay coil is a series combination of R1 and C1 and parallel diode CR1. The capacitor C1 sets a delay (relay dropout) time of about 1.5 seconds, allowing the user time to complete a dial sequence without the transceiver dropping out of key. This feature prevents excessive wear of the transceiver relay contacts. It also allows completion without

the chance of missing any tones, in case you are a fast dialer.

*Dc power source:* If you plan to utilize this unit with an existing base station supply, then no problems should be encountered. But, if you are going mobile, the combination of zener diode CR2 and the 33-ohm resistor should be included. This unit likes a stiff dc source.

If your particular installation is noisy, it may be wise to include more than 1500 mF of filtering across the zener. I have just purchased a very noisy car. I had to add an additional 1500 mF of filtering, which has apparently cleaned up the system. Installation of a coaxial bypass capacitor at the firewall helps the filtering. All cables should be shielded, even from the battery to the firewall bypass capacitor. You can even add an inductor in series with the coaxial bypass, with as much as 2000 mF to ground for that extra measure of protection. It has been found that it is hard to have too much filtering in the automotive system.

*Pad impedance and levels:* The Touch-Tone pad is a low-impedance unit. Variable resistor R2 should be not more than 1k. The specified value of 500 ohms is preferable, otherwise you will have to be almost at the bottom of the pot to keep the tones from distorting. The electrolytic capacitors (1 mF and 2 mF) feeding the audio output line are there to preserve the low-frequency characteristics of the Touch-Tone pad. The 100k resistor, R3, is included to isolate the microphone input from the pad, and to present a high impedance for the microphone input of the transceiver.

Note that there is also included a 1M resistor from the output of the tone burst generator. This value may require some experimentation by the user to obtain the required deviation of the transceiver. If your tone burst generator is one of the types with a very high output level, this resistor may be a value of several megohms. If your transceiver is equipped with *any* kind of clipping, then pot R2 should be set for a level that is below clipping. If it is not, then the receiver Touch-Tone decoder will not recognize

any of the tones, as they will not be true sine waves. These values (1M) work well with the Regency HR-2. If your radio is a different brand, you may need to juggle them to achieve the desired results. But,

the box. However, phono connectors would make it easier to connect and disconnect cables if you ever wanted to work on or modify your pad.

Cutting holes for the pad is a matter of

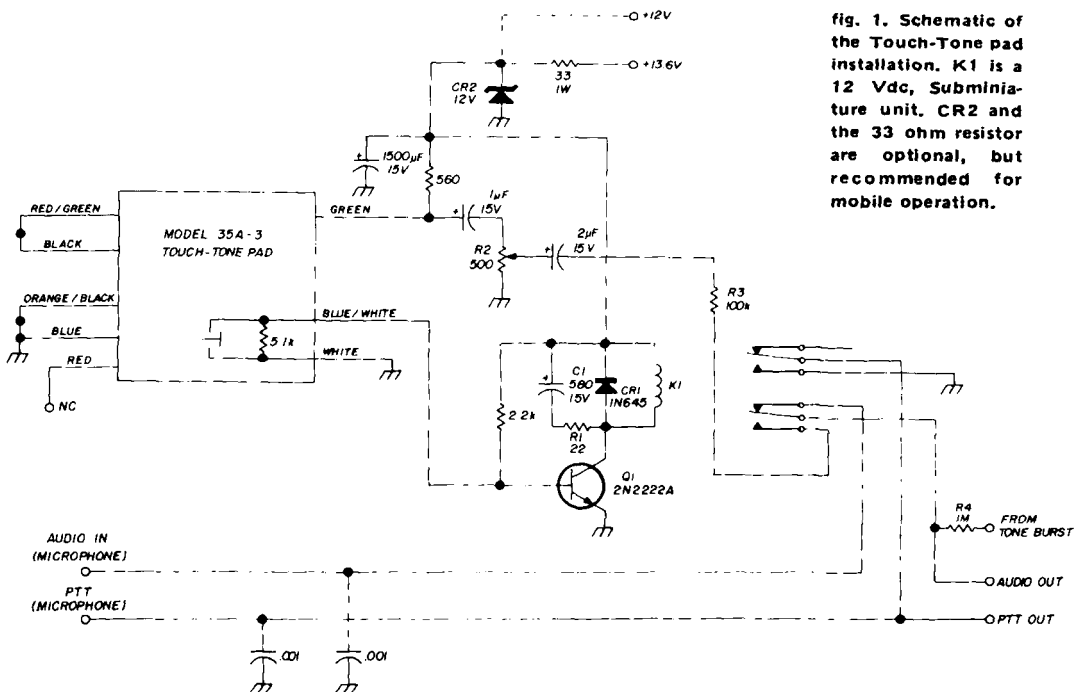


fig. 1. Schematic of the Touch-Tone pad installation. K1 is a 12 Vdc, Subminiature unit. CR2 and the 33 ohm resistor are optional, but recommended for mobile operation.

in no way should these resistors be less than 47k.

### construction

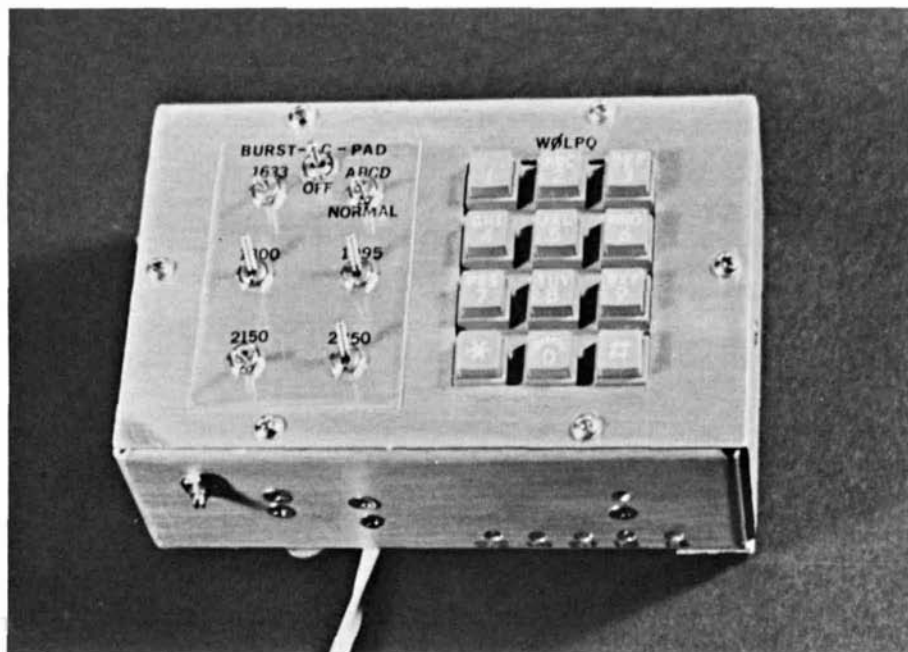
My chassis box is 4 x 6 x 2 inches. The pad is mounted on the right side of the box, and is held in place by two 6-32 screws. One screw mounts on the right hand wall of the box, and the other mounts via an angle bracket secured to the bottom of the box. The tone burst generator is a small pc board (1½ x 2½ inches) mounted on the left wall of the box. I have used a matching microphone jack, which accepts the HR-2 microphone. Shielded Teflon cable runs to the HR-2. In my case, the Teflon cables are soldered directly into the enclosure, and routed out via a small hole in the back of

personal choice. You can either make one large hole to show all of the buttons, or get more ambitious and cut individual holes for each button. If your choice is the latter, then using an existing Touch-Tone instrument (assuming you have Touch-Tone in your home) for exact placement of the holes, will leave you with a very neat enclosure. A nibbling tool is very handy for this, but do it from the inside of the box, or your end result will be a disaster area for a front panel.

In the photograph of my unit, I have used individual toggle switches for the tone burst frequency selection. A rotary switch could have been used, but a suitable subminiature type was not available. The center toggle switch which is labeled BURST-DC-PAD, selects dc power

for either the tone burst generator or Touch-Tone pad. My feeling is that I will never require power to both at the same time, so why not toggle the power between them? Also, why have a beep on

neighborhood ham population, take a little patience, and do it right the first time. You may need to add some more filtering, but really, isn't it worth it? If you are plagued by stray rf floating



The pad and associated circuitry are mounted with a toneburst generator in one box.

the air if you are not using a repeater access channel? A clean radio installation goes a long way toward making friends, and in this age of radiomania, who needs enemies? We've got enough now, why add to the collection?

### trouble

Murphy and his laws always get into the act. Again, I caution the reader about the +12 V supply line. Why? Well, re-read the last two sentences of the preceding paragraph. If you could hear some of the mobile installations around here, you would not ask why. Alternator whine abounds inside that engine compartment. Even with the installation of resistor plugs, radio ignition wiring, and the rest of today's modern innovations, it is still there. But, to keep peace with the

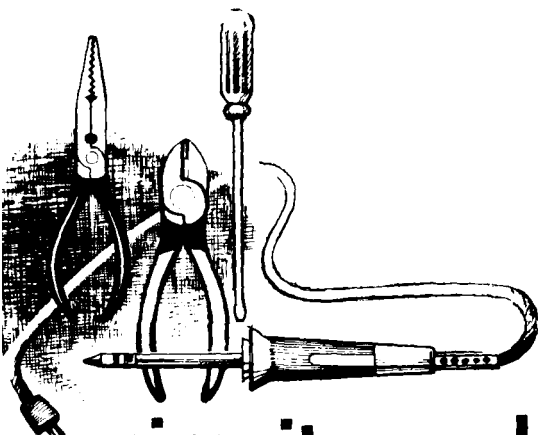
around your mobile, then add a couple of 0.001-mF capacitors across the microphone audio and push-to-talk lines inside the pad enclosure.

Special thanks are due to Tom Yocom, WØZHT, whose digital mind dreamed up the repeater with which this pad is used. Also, thanks are due to Jerry Buck, KØYBM, who designed the tone burst generator used in this system and who also did the photograph for this article.

### references

1. John C. Tirrell, W1DRP, "Push-to-Talk for a Styleline Telephone," *ham radio*, December, 1971, page 18.
2. R.B. Shreve, W8GRG, "Integrated-Circuit Sequential Switching for Touch-Tone Repeater Control," *ham radio*, June, 1972, page 22.

ham radio



# circuits and techniques

ed noll, W3FQJ

## digital IC oscillators and dividers

Modern counters and calibrators feature crystal-controlled multivibrators using NAND and NOR gates plus digital IC divider chains. A popular low cost unit is the 7490 decade divider. In its in-line case it houses separate 5-to-1 and 2-to-1 counters, fig. 1. The 5-to-1 counter is actually an inhibited group of three individual 2-to-1 units as described in detail in last month's column. The 5-to-1 and 2-to-1 combinations have separate input and outputs. The output of one is wired externally to the input of the other to obtain an overall count of 10-to-1. This wiring can be done in two ways; the output of the 2-to-1 fed to the input of the 5-to-1, or, depending upon the desired intermediate count, the output of the 5-to-1 applied to the input of the 2-to-1. The device has great versatility in terms of the selection of preferred counts.

Let's consider these possibilities in terms of calibration frequencies, assuming a 100-kHz drive signal and two decade dividers. Note in fig. 2A that calibration points would be made available at 100, 20, 10, 2 and 1 kHz. If you take the first decade counter and connect the 2 to the 5 rather than the 5 to the 2, a different combination is set up consisting of 100-, 50-, 25- and 2.5-kHz points.

Additional combinations are shown in figs. 2C and 2D. There are other possibilities, too, that can be established using interwiring combinations as shown in fig. 3.

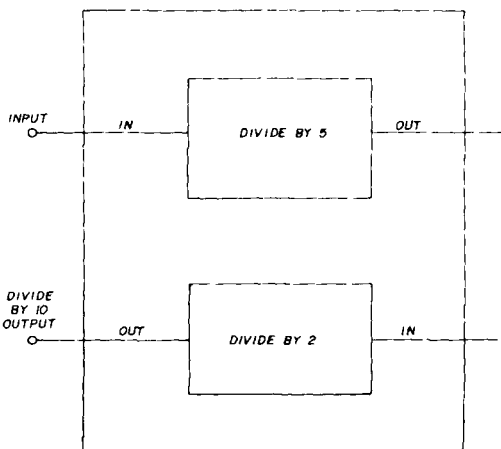


fig. 1. Basic plan of the 7490 decade divider.

All of the above is quite understandable. However, to the uninitiated, the wiring of an integrated circuit appears to be a very complicated thing. Actually it is simple and the major complication is usually the printed-circuit board. However, this can be avoided by using straight wiring techniques as suggested in the first experimental procedures in the June column. If you use binding posts and jumpers it is also possible to change the count sequences between the combinations shown in figs. 2 and 3.

The pin-out wiring diagrams for the 7490 are given in figs. 4 and 5. Note how very simple it is. There are a number of terminals to which no connection is made and another group which are all tied to common. Of course, there are supply voltage as well as input and output connections to be made. The diagrams of fig. 4 are for using the 2-to-1 and 5-to-1 counters separately. Both the connections of fig. 5 provide the 10-to-1 count.

However, in the first example, the first count is 5-to-1; the second, 2-to-1. The second example is the converse, using the initial 2-to-1 count and then the 5-to-1.

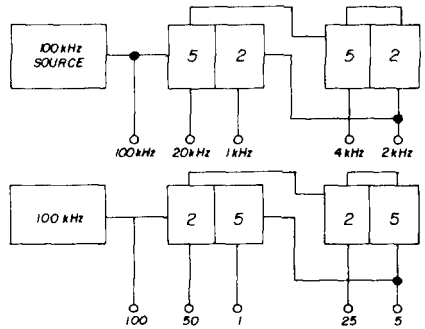


fig. 3. Two additional count possibilities using two decade dividers.

### digital IC oscillators

Digital ICs of suitable design can also be used as high-frequency crystal-controlled square-wave generators. The 7400 NAND gate used initially in this series can be operated as a high-frequency oscillator. Two of the four gates are wired as a multivibrator while a third one is used as a buffer output. Doug Blakeslee, W1KLLK, has used this common IC successfully with the circuit of fig. 6A.<sup>1</sup> Its output is followed by two 7490 decade dividers.

Ted Bensinger, W5PCX, uses the 7400 in the 3-MHz IC oscillator arrangement of fig. 6B.<sup>2</sup> Two of the NAND gates again serve as the multivibrator while the two other sections are pressed into service as buffer and calibrate outputs. Two decade dividers provide the countdown to 30 kHz. W1KLLK operates his circuit at 3 MHz to get the same 30-kHz output. However, he employs a high-frequency 74H00 NAND gate. Theoretically this IC should provide steeper sides and higher harmonic output levels.

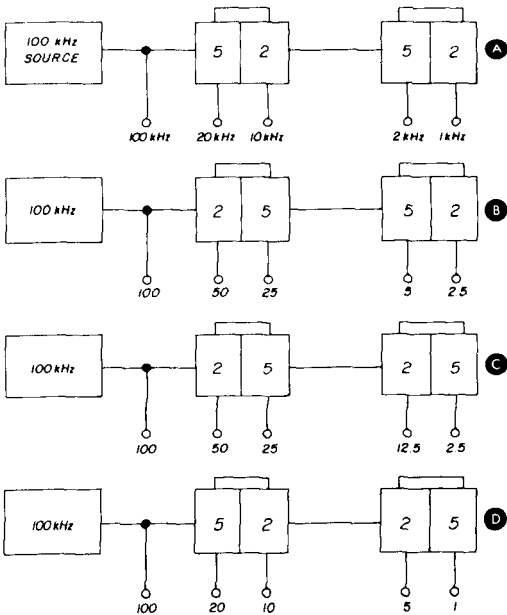


fig. 2. Some of the count possibilities using two decade dividers.

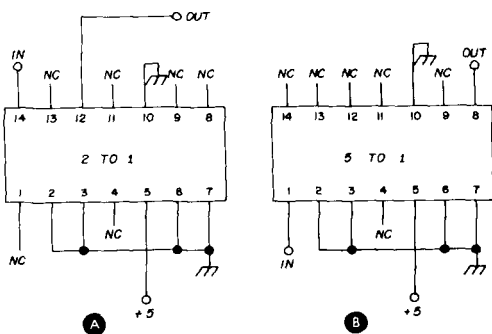
quency is wise because a division of 100 results in a 30-kHz output which is the standard channel separation on the two-meter fm band. The 3-MHz crystal frequency can be adjusted and set on its fifth harmonic which falls on the 15-MHz WWV frequency.

An elaboration of the W1KLK calibrator could include a 3-to-1 counter providing a 1-MHz output, **fig. 7**. In so doing, a calibration point would be available on other WWV frequencies, if and when the 15-MHz signal is not receivable in a given area. Furthermore, the 1-MHz output could also be switched into the divider chain to obtain the same sequence of output frequencies made available from a 1-MHz crystal source.

The dual binary 7473 J-K flip-flop used in last month's experiments can be wired as a 3-to-1 counter as shown in **fig. 8**. Typical waveforms are shown in **fig. 9**. Another 1-MHz circuit using an MRTL NOR gate is shown in **fig. 10**.<sup>3</sup>

### higher-frequency operation

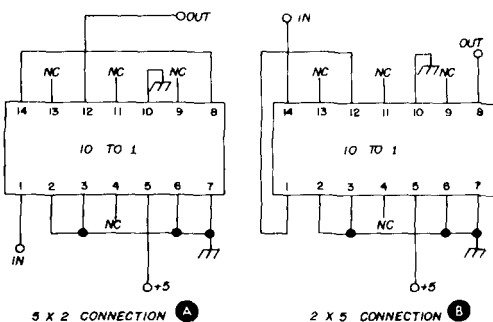
High-speed digital ICs permit crystal-controlled square-wave generation



**fig. 4.** Pin-out diagrams for wiring 7490 decade dividers.

at even higher frequencies. The advantage of an output square wave, of course, is in its high harmonic content. A 5-MHz oscillator of this type produces strong discernible calibration points on the amateur vhf and uhf bands. Typical circuits for an MECL type are given in **fig. 11.4**

Circuit A is for fundamental crystals over a frequency range from approximately 1 MHz to 20 MHz. Circuit B has a broadly resonant output circuit and works with



**fig. 5.** Dividing by ten in 5 x 2 or 2 x 5 sequence.

overtone crystals. I have used this latter circuit successfully with overtone crystals in the 50- to 75-MHz range to produce marks on the 6-, 2-, 1½- and ¾-meter bands. Using a uhf television receiver, calibration points have been checked all the way through the uhf band.

The vhf oscillator can be checked out by using a double calibration technique. I calibrate a 5-MHz crystal oscillator on the highest receivable WWV frequency at the moment. I then tune in its tenth harmonic at 50 MHz on my six-meter receiver. A 50-MHz overtone crystal in the circuit can then be zero-beat with this tenth harmonic of the 5-MHz crystal. Numerous other possibilities exist.

### meteor scatter

I have written before of R.A. Ham and his studies of vhf propagation. Mr. Ham is an amateur radio astronomer. Although I have never met or corresponded with him, I have an exceptional respect for his work because of his skill and his individual interest in basic research. He is interested in a phenomenon because it is presently relevant or practical. Our sently relevant or practical. Our very obsession with "now" relevance and our "so-what" attitudes have become

self-destructive to us as individuals and as a society. One must realize that those things which are non-relevant today might well be of dire relevancy a century from now or even a decade from now; and, furthermore, what we consider relevant today might be most insignificant tomorrow.

Again the following meteor observations of Mr. Ham demonstrate how important radiobeacons can be in basic propagation research. The following is a quote from his coverage in *Electronics Weekly* (British):<sup>5</sup>

"The earth, during its annual orbit of the sun, periodically encounters large swarms of meteor particles which are known to astronomers as meteor showers. Reference to the annual handbook of the

British Astronomical Association will show that the major meteor showers are the Quadrantids in January, the Lyrids in April, the Perseids in August, the Leonids

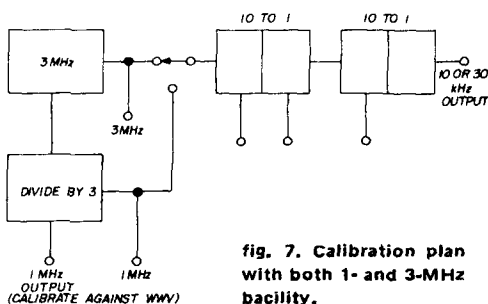


fig. 7. Calibration plan with both 1- and 3-MHz facility.

in November and the Geminids in December.

"The name given to each meteor shower is derived from the constellation of stars from which direction the radiant of the meteors appears to come. For example, one would look toward the constellation of Perseus for the radiants of the Perseid meteors, and toward Leo for the Leonids.

"The visual astronomer hopes for clear skies to enable him to make meteor observations and estimate the number of meteors within the area of the sky which he can see. It is common practice for groups of observers to combine their efforts during a meteor shower to gain as much information as possible. Their reports describe the color, direction and duration of the meteors which they have seen.

"Unfortunately, the visual astronomer has two enemies to contend with; the moonlight, which can make the sky too bright for satisfactory observation, and the weather, which can cloud over the sky at the vital time and prevent any observation at all.

"The amateur radio astronomer can assist his visual colleagues without being hampered by moonlight or overcast skies. The dying ionized trail left by a meteor provides a short-lived passive reflector which can bounce radio signals over a

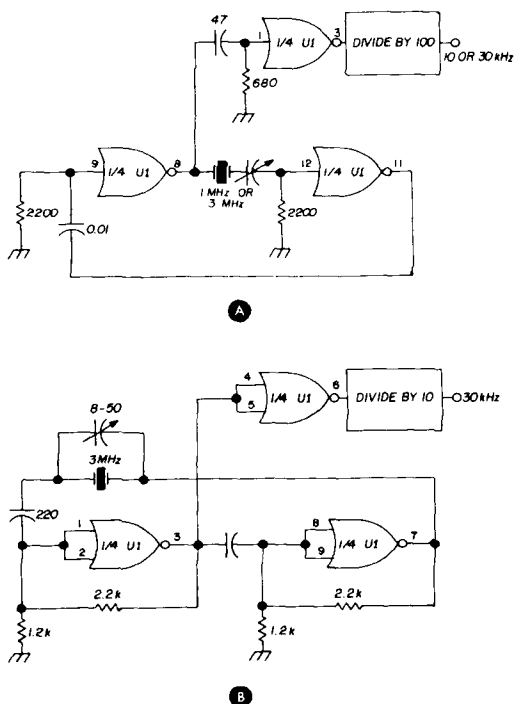


fig. 6. Using the 7400 as a high-frequency oscillator. Use a 7400 with a 1-MHz crystal and a 74H00 with a 3-MHz crystal. Both the 7400 and the 74H00 are quadruple NAND gates, but only three of the gates are used in these oscillators.

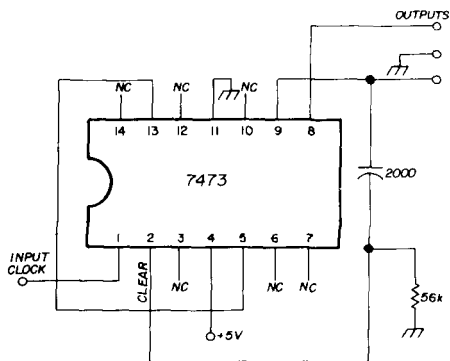


fig. 8. The 7473 3-to-1 divider using two binary flip-flops.

thousand miles. A radio receiving station can tune to the frequency of a distant transmitter and record the number of times that its radio signal is deflected by ionized meteor trails.

"During the life of a meteor shower, there can be a large amount of temporary ionization within the earth's atmosphere, and several United Kingdom radio amateurs have taken advantage of this and established communication, on the vhf bands, over large distances. This means of communication is known as the meteor scatter technique.

"Many amateur radio operators within the past two decades have established two-way meteor-trail communication with fellow amateur stations a thousand miles away. These complicated contacts are confirmed by the exchange of QSL cards between the two stations concerned.

"Attempts to communicate via meteor scatter are usually at pre-arranged times

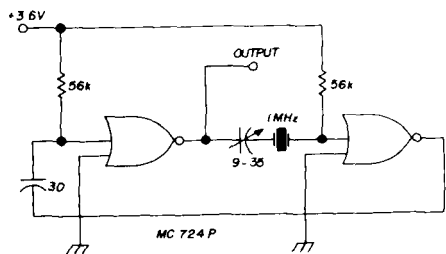


fig. 10. MRTL NOR gate as a 1-MHz oscillator.

and are conducted during a meteor shower when there is a high chance of success.

"Bearing in mind that each meteor trail may only survive for a few seconds, each operator involved in the contact must repeat his call sign and message in high-speed Morse code many times. This intermittent repetition enables the receiving end to piece together tiny bits of the transmitted signal.

"Both operators may have to repeat this procedure at regular intervals for over an hour before a two-way contact can be confirmed.

"Using their CW skill and extreme patience, several United Kingdom ama-

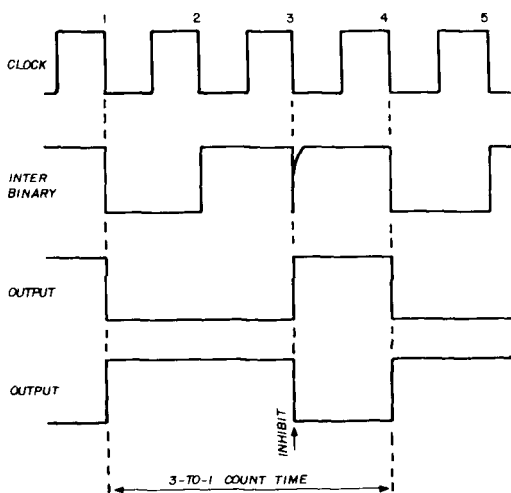


fig. 9. Three-to-one count waveforms.

teurs have confirmed meteor-trail contact with amateur stations in some twenty countries. The majority of this work is with continental amateurs and carried out at the CW end of the two-meter amateur band (144.0 to 144.1 MHz).

"Should a reader hear an amateur during a meteor shower, repeating his own call sign at high speed, take a listen on his frequency when he has finished transmitting and minute parts of the replay may be heard.

"Using meteor trail reflection for communication is a chancy business, but I



personally think that it is a sporting method and a great achievement.

"During the 1968 Geminids, the author (at his Sussex home) conducted an experiment using the RSGB beacon at Thurso in Scotland as the transmitting end. The object of the experiment was to

produced 'GB3' and over 100 minutes later at 2205, another large trail enabled '3GM' to be recorded. The radio signal from a beacon is a continuous tone, which is frequently interrupted with the beacons identification sent in Morse code."

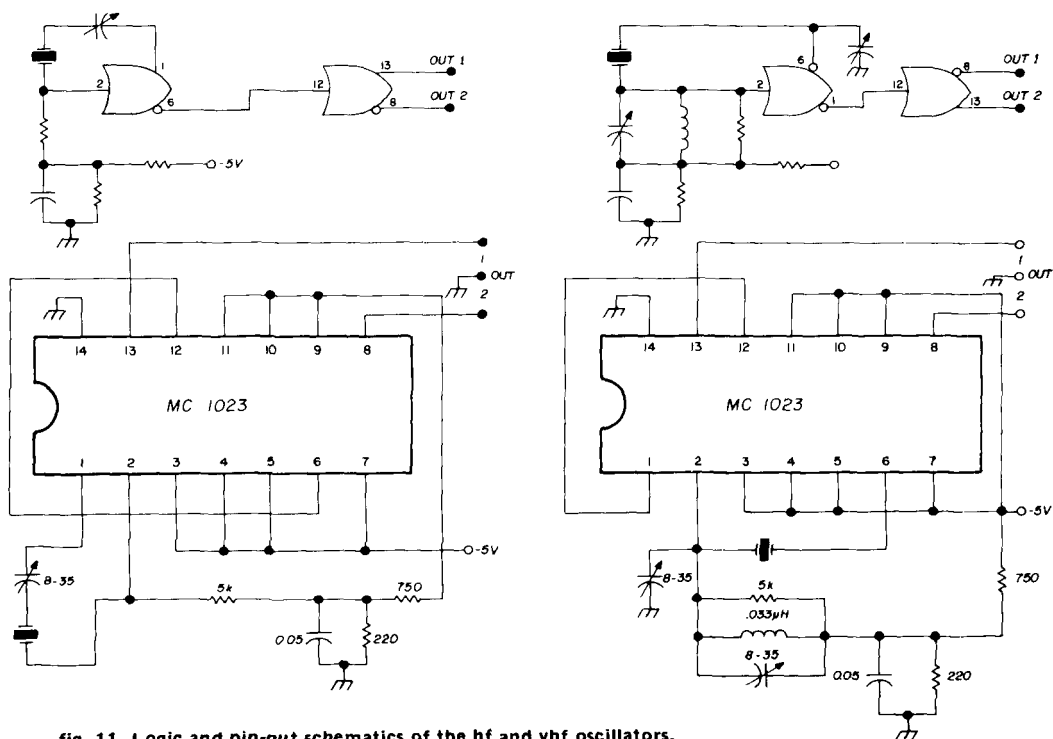


fig. 11. Logic and pin-out schematics of the hf and vhf oscillators.

evaluate the possibility of recording the beacon's identification (GB3GM) by the meteor-scatter technique.

"On December 13, 1968 at 2000 GMT, the author directed his four meter beam toward the north, carefully tuned his receiver to the beacon frequency (70.305 MHz), and accompanied by an interested witness, monitored this frequency until 2300 GMT.

"Throughout the three-hour period, 'pings' of the beacon's tone could be heard bouncing off tiny meteor trails. At 2049 a deflection from a larger trail

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1. Douglas A. Blakeslee, W1KLLK, "Double Standards," *QST*, April, 1972, page 13.
2. Glen E. Zook, K9STH/5, "F.M. Channel Marker Generator," *CQ*, April, 1972, page 41.
3. Bob Botos, "A Frequency Counter using Motorola RTL Integrated Circuits," Application Note AN-151, Motorola Semiconductors, Phoenix, Arizona.
4. Chuck Byers, "IC Crystal-Controlled Oscillators," Application Note AN-417, Motorola Semiconductors, Phoenix, Arizona.
5. R.A. Ham, "Report on Recent 32-Day Meteor Observation," *Electronics Weekly*, November 3, 1971, page 16.

ham radio

# comparison of fm receiver performance

FM equipment manufacturers use two different methods of describing receiver performance — this article discusses the differences between these two systems

Many radio amateurs are confused when they compare the fm sensitivity and performance claims of the various manufacturers. And they have every right to be confused because the methods of measurement are quite different.

Some manufacturers use the 20-dB quieting method, while others use the 12-dB SINAD technique. A comparison of these two very different measurements can be best illustrated with the graph in fig. 1. In this graph the zero dB receiver output reference is the output of an unquieted receiver without an input signal after the volume control has been

adjusted so the receiver delivers rated audio output with an fm generator deviating 0.67 at 1000 Hz.

Curve 1 shows noise quieting, and it is plotted by varying the output of an unmodulated rf generator and measuring receiver output. The 20-dB quieting point occurs at approximately  $0.45 \mu\text{V}$ . This is the most simple sensitivity test which can be performed.

In the SINAD test a modulated rf signal generator is used. In this test, receiver output initially increases with increasing signal level (curve 2). The limiters become saturated at approximately  $0.5 \mu\text{V}$  and receiver output levels out at +4.5 dB. The actual audio in the output signal is signal plus noise plus distortion (S + N + D).

Now curve 3 must be plotted, using an rf signal generator which is modulated with 1000 Hz. A distortion analyzer is connected to the output of the receiver and nulled to reject the 1000-Hz signal. Then the remaining noise and distortion can be measured.

Noise and distortion declines quite rapidly as input signal increases. When the input signal is large, the noise component is practically nil, so curve 3 levels out at the inherent distortion level of the receiver.

The difference between curve 3 and curve 2 is the SINAD ratio. The 12-dB SINAD sensitivity is at approximately  $0.25 \mu\text{V}$  input as shown in fig. 1. The SINAD ratio is computed from the following formula:

$$\text{SINAD ratio} = (S + N + D) - (N + D)$$

I will not discuss the pros and cons of each of these systems, but both have their place in determining the actual performance of an fm receiver. I can only hope that the individual manufacturers will quote both results so amateurs can make realistic comparisons.

Most amateurs talk about receiver sensitivity in terms of microvolts. The graph in fig. 1 is also calibrated in dB relative to 1 watt, or dBW. In a 50-ohm system, a  $0.5 \mu\text{V}$  signal is equivalent to  $-143 \text{ dBW}$ . (When doubling the rf input

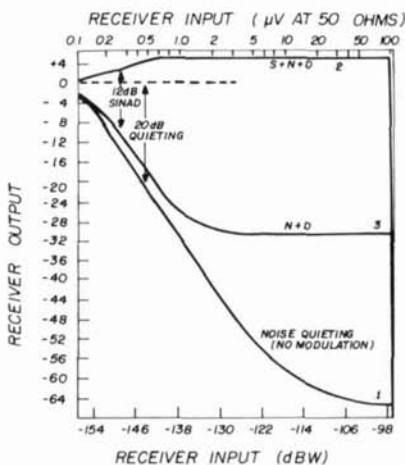


fig. 1. Comparison of fm performance tests. 20-dB quieting is the difference between the zero reference level and noise quieting (curve 1). 12-dB SINAD is the difference between signal plus noise plus distortion (curve 2) and noise plus distortion (curve 3).

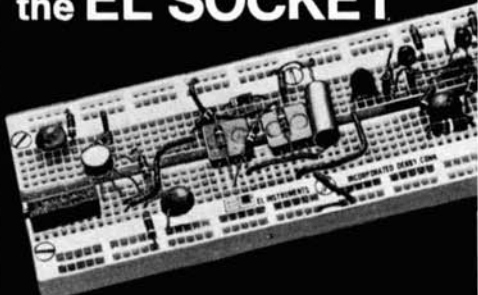
signal, remember that you must increase dB by six.) Another term which is used occasionally is dBm, power in dB relative to 1 milliwatt; dBW can be converted to dBm by subtracting 30 dB. For example,  $-143 \text{ dBW} = -113 \text{ dBm}$ . A complete nomograph of power, impedance and dBm is given in reference 1.

#### reference

1. W. E. Pfister, Jr., W2TQK, "Power, Voltage and Impedance Nomograph," *ham radio*, April 1971, page 32.

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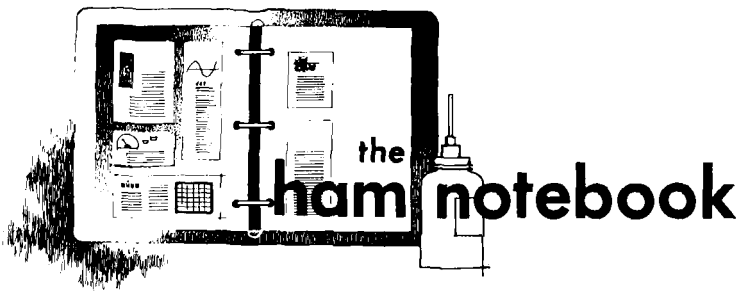
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## solid-state vibrator replacement

With the flood of solid state vhf rigs on the market, the price of the vibrator powered units has dropped to a level where almost anyone can afford to go vhf mobile. This solid-state switch fits in a small minibox and has only two wires that connect to the vibrator socket (not counting the chassis ground). The only

transients on their edges. These transients are bypassed by an RC combination called a buffer circuit. This is a resistor and capacitor series combination in parallel with the secondary high-voltage winding.

The schematic in fig. 2 shows the oscillator. This unit duplicates the electrical function of the vibrator and thus could be likened to a solid-state switch. With power applied, the circuit first grounds terminal A then terminal B. A

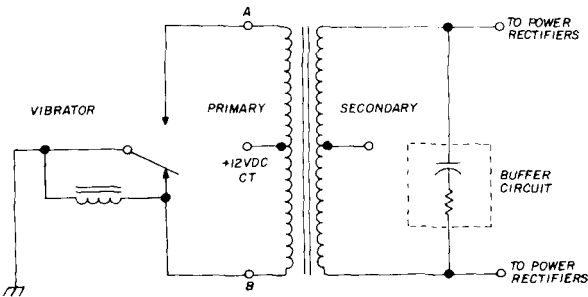


fig. 1. The basic vibrator supply.

addition to the rig is a small spst switch mounted on the power-supply chassis. Yet this simple unit eliminates the hash and mechanical problems common to vibrators.

The schematic, fig. 1, shows a basic vibrator primary circuit and a secondary winding with buffer. It's basically a mechanical switch with the center tap of the primary at plus 12 Vdc. The vibrator alternately grounds point (A) then (B) creating a square wave of dc. This causes current to flow in the secondary winding. The induced currents always have some

brief description would indicate the toroid (similar in dimensions to an 88-mH toroid) is the heart of the circuit. Its function is to feed back current to the bases, alternately saturating Q1 and Q2. CR1 and CR2 are zeners suggested by the transformer company to protect the power transistors from transients. They could be left out but they are inexpensive insurance. The R1, R2 and R3, R4 combinations are resistive current dividers, limiting the base current to a saturated but safe level.

Construction is not critical. The tran-

sistors should be heat sunk as they do run warm under transmit condition. Other components should be mounted with care keeping vibration in mind. A short barrier strip mounted on the outside of the box would permit quick connection and testing without connection to the rig. The switch can be tested without connection to the rig by applying plus 12 Vdc to pin four of the toroid and providing a ground return. The

determined from the schematic of your rig. Run two heavy leads from terminals A and B to the pins that correspond to the ends of the primary. A spst switch should now be mounted on the power-supply chassis, accessible from the top side of the chassis. This switch is wired in series with the buffer circuit. Be sure to label this switch as *buffer in* in the *on* position and *buffer out* in the *off* position. The *buffer out* position is used with

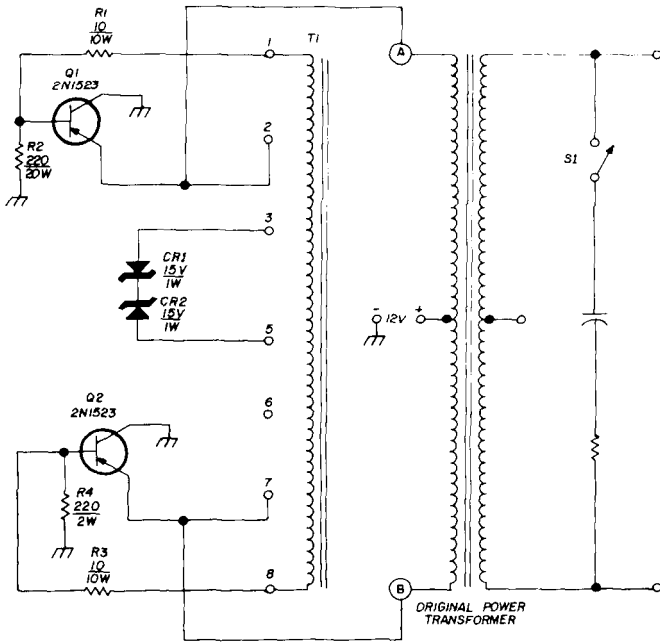


fig. 2. Transistor switch is the heart of the solidstate vibrator replacement. T1 is an Osborne 2709, available from Osborne Transformer Company, 2823 Mitchell Avenue, Detroit, Michigan 48207.

switch will oscillate at a low frequency (you'll hear a buzz from the toroid); however, when a load is connected, the operating frequency of the switch is near 1000 Hz. If the load becomes too great, the circuit will stop oscillating and save the transistors from thermal runaway.

Since the switch has only two connections to the rig plus the chassis ground, the hookup time is minimal. The vibrator is removed from its socket. The base of an old vibrator could be used as a quick connector to the vibrator socket. The two pins to be used should be

the solid-state switch as the reactance of the buffer at the new operation frequency (1000 Hz) is much too low and the power it would pass would burn out the resistor in short order. The rig now is ready to operate with the option of returning to vibrator with the flick of the switch to the *buffer in* position and reinstalling a vibrator.

This same basic circuit appeared in the 1970 edition of *The Radio Amateur's Handbook*, page 336, as an integral part of a more complex mobile power supply.

Ray Kashubosky, K8RAY

## aligning audio filters

The bandpass characteristics of an audio filter may be best displayed and adjusted through the utilization of an audio sweep system with an oscilloscope. However, such systems are expensive and appear only in the best-equipped laboratories. Consequently, filters are sometimes assembled with complete reliance for their tuning entrusted in the labeled values and tolerances stated on the components.

After constructing the excellent 200-Hz filter described by K7UDL in the September, 1967 issue of 73, I questioned whether the four 88-mH toroidal coils were accurately labeled. My suspicions were confirmed when five coils were compared and found to exhibit five different values of inductance.

I found a simple method for precisely matching capacitors, inductors and L-C tank tuning which should be helpful in similar applications. I built the two-terminal inductor oscillator shown in fig. 3 and used a high-impedance earphone to monitor the output tone.

Before assembling a filter, all identical L-C tanks are switched in individually, to

permit easy adjustment, by removing or adding turns. Match capacitors by switching various capacitor combinations across one toroidal coil until each combination produces exactly the same tone in the earphone. Note that the tone monitored in the earphone does not represent the operating frequency of the assembled filter. The test oscillator serves purely as a comparison device.

In simple one-section filters, tuning may be accomplished by adjustment of either inductance or capacitance. However, in a symmetrical multi-section filter, the Q of equivalent sections should be equal. This requires that both inductance and capacitance in those sections be matched.

With components known to be matched, filters accurately aligned to design frequencies may be assembled with confidence that optimum performance will be enjoyed.

Gene Brizendine, W4ATE

## overtone crystals

Do not overtone crystals below 10 Mhz unless you have a frequency meter. The frequency *will not be* three, five, seven or nine times the fundamental. In fact, the frequency can be as much as three percent low. This is shown in fig. 4, a graph of the expected amount of error of low-frequency crystals working on their common overtones. The graph shows average figures taken from tests on many different crystals.

In general, smaller crystals, like the HC-18/U, have bigger errors than larger types, like the HC-6/U or FT-243. However, the older air-mounted types with flat plates (like the 10X) will always be exact multiples.

You can use overtone crystals at their fundamental frequency and at other overtone frequencies. Here is how: Crystal manufacturers can generally make fundamental crystals up to 20 MHz, and they make them for overtones above this. You can easily find out what overtone a crystal works at by dividing the fre-

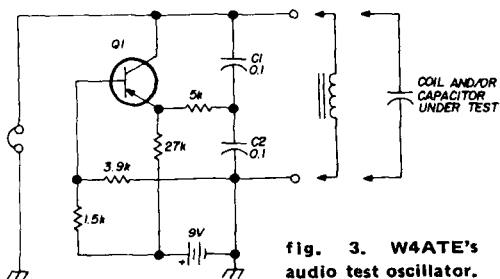


fig. 3. W4ATE's audio test oscillator.

form the frequency-determining elements of the oscillator. If the audio tones produced by all tanks are not identical in frequency, tank adjustments are necessary. Select matched inductances by switching the coils across one capacitor until you find coils producing the same tone.

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quency by odd integers (3, 5, 7, 9) until you get the highest frequency *below* 20 MHz. For example, a 75-MHz crystal cannot be a third overtone because 75 MHz divided by three is 25 MHz. This is too

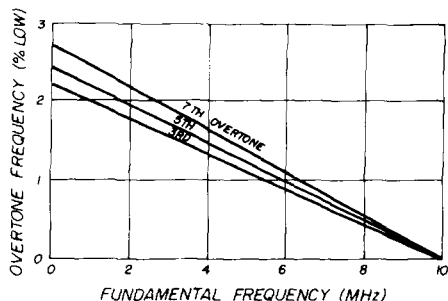


fig. 4. Graph of the projected frequency error when overtoning crystals below 10 MHz.

high, being above 20 MHz. It is a fifth-overtone crystal, as 75 MHz divided by five is 15 MHz. This same crystal can be operated at its third overtone to produce 45 MHz, or on its fundamental to get 15 MHz.

If a particular crystal will overtone at all (some very old ones will not), the Butler circuit will do it. Also use the Butler if you want a high overtone like the ninth. Otherwise the Pierce circuit is fine for good crystals at low overtones.

Martin Mann, G8ABR

## key and vox clicks

A casual review of Official Observer logs suggests that there are about four times as many transmitters with key clicks on the *make* of the contacts, than on the *break*. Often, the amount is not large, but it can be heard several kilohertz to each side of the signal causing unnecessary interference. It may not be noticed by a station tuned to receive the signal because the beat note may obscure it.

Recently, a local amateur telephoned about his signal, so we worked the matter out. He uses a Swan 500C with the keying circuit shown in fig. 5 (except for the plug and key). You will see that capacitor C1507 charges when the key is opened, the rate of charge being limited by R1608. However, when the key is

closed, the capacitor discharges abruptly, possibly through contact-bounce in the key. This would tend to make the click mushier.

We connected a resistor decade box in series with the key. When it was set at about 10 kilohms, the scope pattern showed a very nice sloping rise which was very similar to the sloping fall at the other end of dots and dashes. The click was entirely gone.

There is no modification to the transmitter. The resistor can be mounted on the key or, in many cases, inside of the key plug.

It is probable that this same approach will apply to other exciters, many of which have a filter that is adjustable on one end of dots and dashes, but not at the other end.

Some stations have a *make* click that occurs only after a pause, such as in the first character of a word. Many of these are the result of using a fast vox adjust-

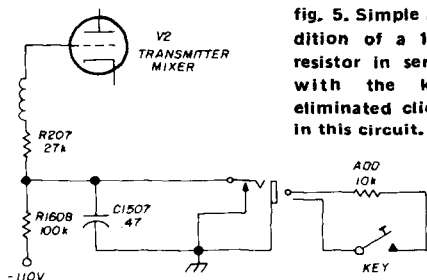


fig. 5. Simple addition of a 10k resistor in series with the key eliminated clicks in this circuit.

ment in an attempt to obtain break-in operation. However, some exciters do not have a key-click filter in the circuit that is operated by the vox. This can cause a click at the *start* of each word sent, but not *within* words unless the keying is *very* slow. Where this occurs, the vox may be slowed down, or a suitable filter can be inserted in the circuit where the vox turns the transmitter on and off. Incidentally, this click also occurs in phone operation.

E. H. Conklin, K6KA



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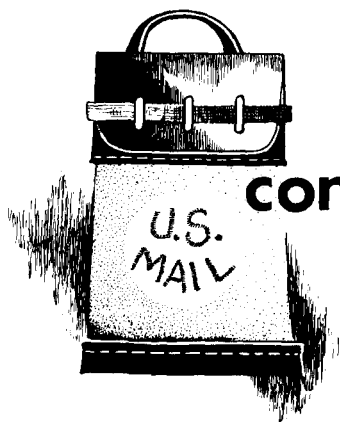
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## comments

### RTTY for the blind

Dear HR:

I recently constructed about a dozen TTL SELCAL units for amateur RTTY. During the course of this activity it has become evident that it would not be too difficult to use similar techniques to build a unit that would make it possible for blind people to read a teletype signal right off the air. If it could be arranged it would open up a new field of interest for our blind friends.

Circuitry can readily be arranged which will recognize any teletype character or letter and, with suitable processing, turn on a light, energize a relay, or as in our case, operate a solenoid. The idea we are presenting is to arrange ten such small solenoids (for use in the case of 5 level teletype) so that 5 of them would be continuously operated by the incoming characters in the LTRS mode of operation while the other five would be operated by the FIGS. The receiving operator could place his left fingers over one of these sets and his right fingers over the other.

It should not take too much practice for the receiving operator to be able to recognize the characters as they came in from the various "highs" and "lows" in each character. Initial practicing could be done at any slow speed such as five or ten words per minute by the simple expedient of having a helper punch occasional letters on a keyboard connected to the

"reader." Possibly punched tapes could be played back at any speed the operator chose through the use of a dc speed-controlled motor belted to the standard tape reader motor. With a few month's practice some operators should be able to copy at the full rate of sixty words per minute because each character would be available both to feel and hear instantly. Any necessary combination of characters and blank spaces could be utilized to assist slower operators.

There is a possibility that a whole new area of interest might develop from these suggestions. Blind people could write back and forth to each other on punched tapes. Articles and even whole books could be made available on punched tape. In some cases ordinary magnetic tape could be used to record the familiar audio tones of the RTTY signals and the receiving operator (with a demodulator, modified SELCAL, and our suggested solenoid combination) could "read" the tape. The most useful purpose we are thinking of for the moment, though, is that blind hams with the proper equipment could communicate with any other ham using teletype.

The diagrams will show briefly the techniques we are suggesting. Possibly other ideas from interested parties would help to realize a satisfactory arrangement. While thinking on this subject, it might be kept in mind that people who are deaf could communicate with each other or with blind persons or with anyone else through RTTY circuits. Anyone working on these ideas would be wise to consider the whole concept.

Larry Walrod, VE7BRK  
Kelowna, B. C., Canada

## diode surge protection

Dear HR:

In the article "Diode Surge Protection" by John Lapham, WA7LUJ, which appeared on page 65 of the March, 1972 issue of *ham radio*, the author is correct when he is talking about high-voltage power supplies using modern diodes — the transformer's secondary dc resistance is usually sufficient to protect the bridge diodes against surge current. However, in low-voltage supplies it is a critical problem which should not be overlooked.

An example is a power supply I was designing for my workbench. I had a 10 V transformer rated at 4 A. It had a secondary dc resistance of 0.33 ohms. With the other components on hand — an epoxy bridge rectifier and a 2000 mF capacitor — I had a 50 A surge current for one duty cycle and I needed 0.6 ohms of surge resistance to limit the surge current. Without the addition of an external resistor, I would be keeping the diode manufacturers in business.

It only takes a few seconds to make the surge calculations which could save lots of headaches and pennies later. I recommend it for any design, but especially in low-voltage designs.

Mr. Lapham stated, "Surge protection is a needless feature in power supplies." This is simply not so in low-voltage designs, and I think this distinction should be made clear.

Joseph P. Bremmer, WB6KXF  
Pomona, California

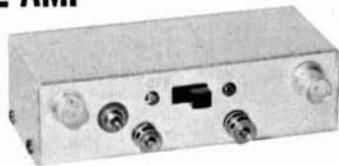
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Dear HR:

Congratulations on your promotion to make Canadian amateurs aware of *ham radio*! Glad to say I am one of the "in crowd" already being a subscriber for several years. Good Luck!

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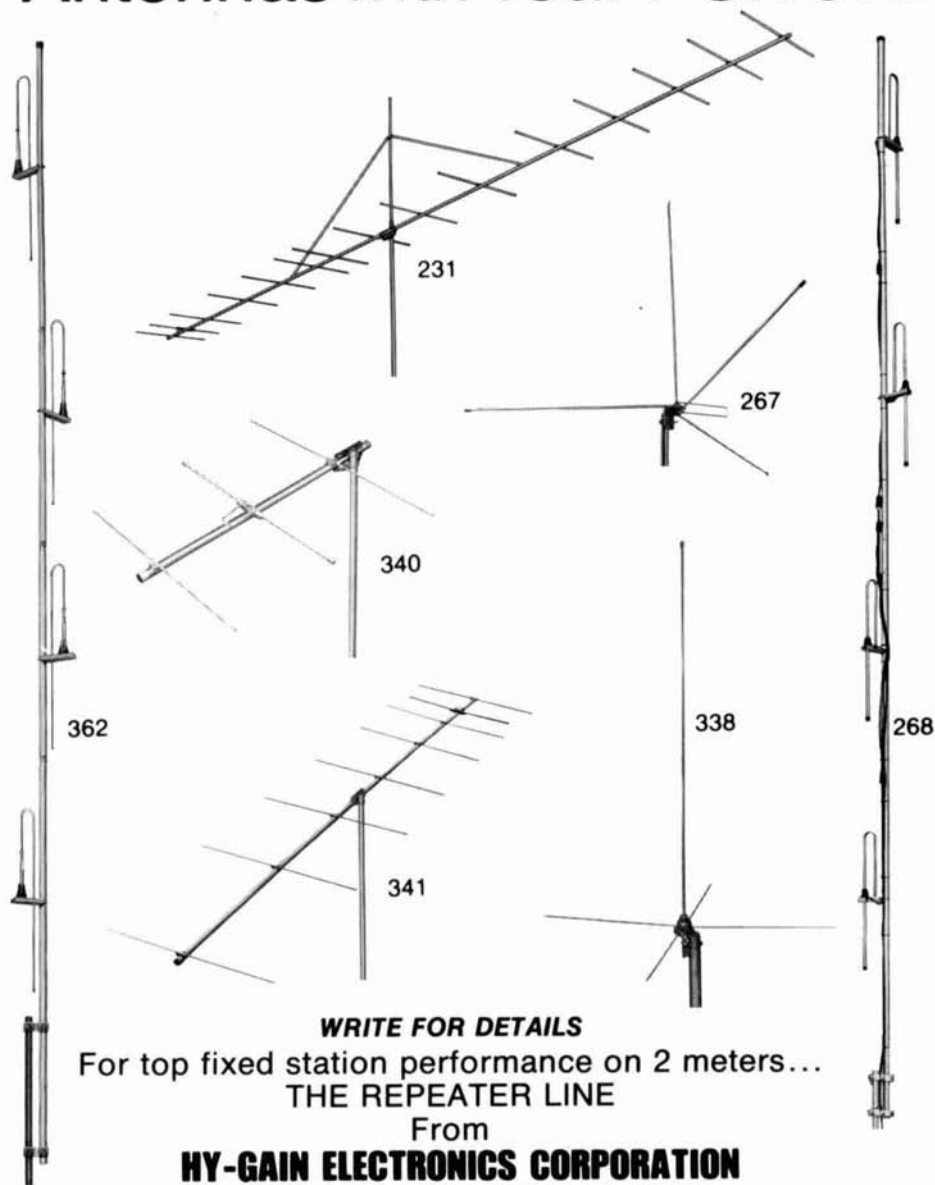
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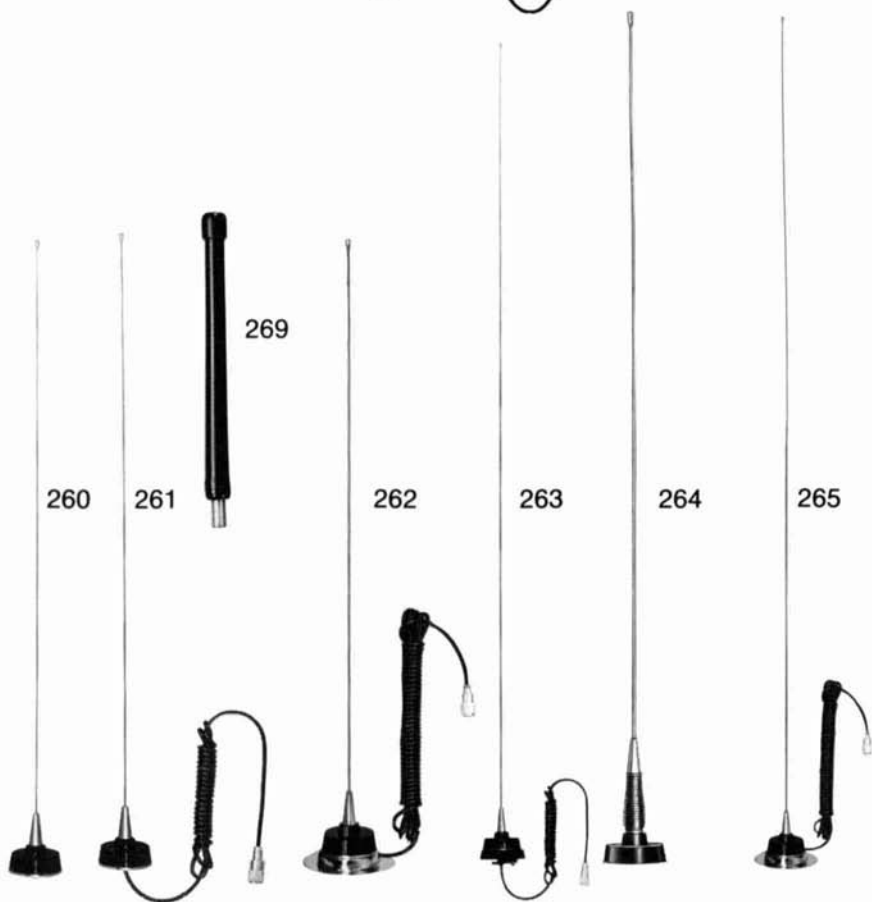
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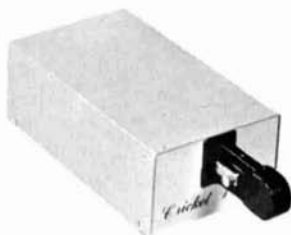
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	1-99	100-999	1000 up	100-999	1000-9990	10000 up		1-99	100-999	1000-999	1000-9990	10000 up	
7400	.26	.25	.23	.22	.21	.20	7474	.50	.48	.45	.43	.40	.38
7401	.26	.25	.23	.22	.21	.20	7475	.80	.76	.72	.68	.64	.60
7402	.26	.25	.23	.22	.21	.20	7476	.56	.53	.50	.48	.45	.42
7403	.26	.25	.23	.22	.21	.20	7480	.76	.72	.68	.65	.61	.57
7404	.28	.27	.25	.24	.22	.21	7482	.99	.94	.88	.83	.78	.72
7405	.28	.27	.25	.24	.22	.21	7483	1.63	1.55	1.46	1.38	1.29	1.20
7406	.52	.50	.47	.44	.42	.39	7485	1.42	1.35	1.27	1.20	1.12	1.05
7407	.52	.50	.47	.44	.42	.39	7486	.58	.55	.52	.49	.46	.44
7408	.32	.30	.29	.27	.26	.24	7490	.80	.76	.72	.68	.64	.60
7409	.32	.30	.29	.27	.26	.24	7491	1.43	1.35	1.28	1.20	1.13	1.05
7410	.26	.25	.23	.22	.21	.20	7492	.80	.76	.72	.68	.64	.60
7411	.28	.27	.25	.24	.22	.21	7493	1.12	1.05	1.00	.93	.87	.80
7413	.58	.58	.52	.49	.46	.44	7494	1.18	1.12	1.05	.99	.93	.87
7416	.52	.50	.47	.44	.42	.39	7495	1.18	1.12	1.05	.99	.93	.87
7417	.52	.50	.47	.44	.42	.39	7496	1.18	1.12	1.05	.99	.93	.87
7420	.26	.25	.23	.22	.21	.20	74100	1.52	1.44	1.36	1.28	1.20	1.12
7421	.26	.25	.23	.22	.21	.20	74107	.52	.49	.47	.44	.42	.39
7423	.80	.76	.72	.68	.64	.60	74121	.56	.53	.50	.48	.45	.42
7425	.50	.48	.45	.43	.40	.38	74122	.70	.67	.63	.60	.56	.53
7426	.34	.32	.31	.29	.27	.26	74123	1.21	1.06	1.00	.94	.89	.83
7430	.26	.25	.23	.22	.21	.20	74141	1.63	1.55	1.46	1.38	1.29	1.20
7437	.56	.53	.50	.48	.45	.42	74145	1.41	1.33	1.26	1.18	1.11	1.04
7438	.56	.53	.50	.48	.45	.42	74150	1.63	1.55	1.46	1.38	1.29	1.20
7440	.26	.25	.23	.22	.21	.20	74151	1.20	1.13	1.07	1.01	.95	.88
7441	1.73	1.64	1.55	1.46	1.37	1.27	74153	1.63	1.55	1.46	1.38	1.29	1.20
7442	1.27	1.21	1.14	1.07	1.01	.94	74154	2.43	2.30	2.16	2.03	1.89	1.76
7443	1.27	1.21	1.14	1.07	1.01	.94	74155	1.46	1.39	1.31	1.23	1.16	1.08
7444	1.27	1.21	1.14	1.07	1.01	.94	74156	1.46	1.39	1.31	1.23	1.16	1.08
7445	1.71	1.62	1.53	1.44	1.35	1.26	74157	1.56	1.48	1.39	1.31	1.23	1.15
7446	1.24	1.17	1.11	1.04	.98	.91	74158	1.56	1.48	1.39	1.31	1.23	1.15
7447	1.16	1.10	1.04	.98	.92	.85	74161	1.89	1.79	1.68	1.58	1.47	1.37
7448	1.44	1.37	1.29	1.22	1.14	1.06	74164	1.89	1.79	1.68	1.58	1.47	1.37
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7451	.26	.25	.23	.22	.21	.20	74180	1.20	1.13	1.07	1.01	.95	.88
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7472	.38	.36	.34	.32	.30	.29	74198	2.81	2.65	2.50	2.34	2.18	2.03
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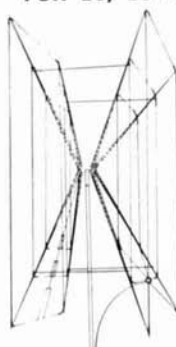
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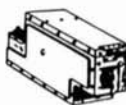
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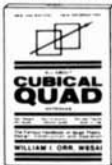
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1-711C Dual diff. comp. (A)	\$1
1-723C Voltage regulator (A)	\$1
1-TVR-2000 Hi-power 723 (A)	\$1
1-747C Dual 741C, TO-5	\$1
1-748C Freq. adj. 741C (A)	\$1
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Buy TO-5 or DIP (DIP in line) case

### IT'S NEW! 12-DIGIT "CALCULATOR ON A CHIP" 24.95

Sorry, can't name U.S. maker! Type SD5051. Similar to Mosek 5012. Outperforms Texas 8-digit TMS-1802. Features: 40-pin DIP; not 5, not 8, but only one "calculator chip"; 12-digit capacity; adds, subtracts, multiplies, divides, fixed point (0, 2, 3 or 4 decimal places), chain operations, leading zero blanking, 7-segment decoded display output, overflow lockout, and negative sign output. Drives incandescent, fluorescent, Nixies, and LED readouts, complete with instruction booklet.

### COUNTING SYSTEM 6.50

Includes SN7490, decade counter, SN7475, SN7441 BCD decoder driver, 0-to-9 Nixie tube, socket & instructions.

### LED READOUT COUNTING SYSTEM 8.88

3 for \$24

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Type	Description	Sale
531	Hi slew rate op amp	3.50
533	Micro power 709	3.88
536	FET input op amp	4.50
540	70W pwr driver amp	2.04
545	Phase loop (A)	3.50
568	Function generator (A)	3.55
567	Tone decoder (A)	3.85
595	Four Quadrant multiplier	3.35
550	Precision 723 V. R. pak	1.19

### INTEGRATED CIRCUIT SOCKETS

Type	Description	Sale
14-Pin, dual in line	5.45	
16-Pin, dual in line	.50	
TO-5, 8 or 10 pins	.29	

### ALLEN BRADLEY 'MICRO-POTS'

Type G, 1/2" dia, 5 1/2" high, Mounts 1 1/2" with shaft, linear, immersion-proof high freq.

Ohms	Price
75	2.9K
100	2.5K
500	5.0K
10K	7.5K
20K	10.0K

5 Meg We stock Locknut Bushing & Screwdriver types at same low prices.

### 35 WATT AUDIO AMPLIFIER BASIC \$3.33

For Class AB use. Basic includes: Signetec 540 30 transistor high power driver TO-5 "IC", with a pair of complimentary 35-watt plastic transistors, i.e. 2N5296 npn and 2N6109 pnp. With schematics, printed circuit and parts board layouts.

PIV	2 Amp	6 Amp
50	\$5.69	\$8.88
100	.79	.99
200	.95	1.25
400	1.19	1.50
600	1.35	1.75
800	1.59	1.95
1000	1.79	2.25

Code: 2 amp TO-6 case  
6 Amp 1/2 x 1/2 x 1/16 sq

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PRV	3 amp	6 amp	10 amp	15 amp	25 amp
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100	.35	.52	.64	.85	1.05
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300	.79	.89	1.10	1.45	1.65
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500	1.25	1.39	—	2.55	2.25
600	1.50	1.69	—	—	2.65

Code: 3 & 6 amp TO-5, 10 amp TO 66, 15 & 25 amp stud.

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Burrhoughs B7971. 0-to-9, A-to-Z. 2 1/2" high char. With socket, 170V.

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1 AMP  
1000 PIV

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SU4	\$1.49
SR4	3.95
866	7.95

### ALLEN BRADLEY 'MICRO-POTS'

Type G, 1/2" dia, 5 1/2" high, Mounts 1 1/2" with shaft, linear, immersion-proof high freq.



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**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.

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

**PRINTED CIRCUIT DRILL BITS.** Trumbull, 833 Balra Drive, El Cerrito, California 94530.

**QSL'S — BROWNIE W3CJ1 —** 3111B Lehigh, Allentown, Pa. 18103. Samples 10¢. Cut catalogue 25¢.

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**TELETYPEERS —** Toroids, tools, supplies, parts, equipment. SASE (16¢) for list. Typetronics, Box 8873, Ft. Lauderdale, Fl. 33310. W4NYF.

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**CV-253/ALR FREQUENCY CONVERTER,** 38-1000 MHz. Continuous tuning, 30 MHz out. New, unopened carton. \$165.00. K4EPW, 1230 Briar Hills, Atlanta, Georgia 30306.

**WANTED: COLLINS 312B-5, K2QDE,** S. Martin, 2011 Ocean Ave., Bklyn, N. Y. 11230. 212-998-2029.

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**RACKED VHF-UHF STATION,** all mode, SB-303 + accessories. L. M. Gessin, 446 Cedarwood Ter., Rochester, N. Y. 14609. "WAZZNC".

**WARREN, OHIO ARA'S FAMILY HAMFEST.** Aug. 20. Giant flea market, swimming, picnicking, all free. Displays, mobile check-in. Camping available. Yankee Lake, Rt. 7 near I-80. Light plane accommodations 1 mi. away at Obermeyer Airport. No fees. Free transport to Hamfest. Details QSL WBVTD.

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

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

**WORLD QSL —** See ad page 108.

**THE ANNUAL IOWA 75 meter** phone net picnic will be held on August 20, 1972 — at Riverview Park in Marshalltown, Iowa. All amateurs and families are cordially invited, or anyone interested in amateur radio. Each should bring a covered dish and his own service. Festivities will begin around noon. Prizes will be offered and a swap table will be available. Nearby camping facilities are available.

**TOPE ENCODERS AND DECODERS —** New line of solid state encoders and decoders compatible with any sub-audible continuous tone system. Small in size, usable from 67-250 Hz, \$8.95 to \$14.95. Send for literature. Communications Specialists, Box 153, Brea, Calif. 92621.

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RECEIVER BOARD KIT \$54.95. 6-CHANNEL ADAPTER \$9.95. DISCOUNTS AVAILABLE TO CLUBS AND LOCAL DEALERS.

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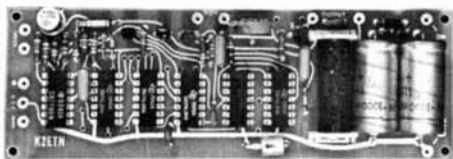
Also available — PREAMPS AS SEEN ON PAGE 25 OF MARCH HAM RADIO. For 6 or 2 Meters or Commercial Freq. \$10 Wired, \$6 Kit. Quantity Prices.

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Now a Keyer with every feature of the most expensive units without the unnecessary frills. Woerner Enterprises has designed a complete ready to use Keyer on a single printed circuit board. Use your own Paddle and incorporate the Keyer in your station console or power supply. Built-in regulated power supply operates from 6.3 VAC or 7½ VDC and only requires 100 Ma.

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

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**THE RADIO CLUB OF TACOMA HAM-FAIR** will be August 19th and 20th at the Sportsmen's Chateau, 164th and Canyon Road, south of Tacoma. Contests, transmitter hunts, technical talks and displays, manufacturer's displays, auction, country store, beauty contest, and door prizes. Grand prize a solid-state 2 meter transceiver. Talk-in frequencies - 146.76 and 3965. Camping - \$1.50 per night with electrical hookup. Advance registration will cost \$5.00 and includes Saturday evening dinner, or \$3.00 without dinner. Contact Emil Koth, K7GPK, 13616 10th Ave. East, Tacoma, Washington 98445.

**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.

**RECIPROCATING DETECTOR PARTS KIT.** Write Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

**CINCY STAG HAMFEST:** The 35th Annual STAG Hamfest will be held on Sunday, September 24, 1972 at the ALL NEW Stricker's Grove, on State Route 128; one mile west of Ross (Venice), Ohio. Check local area map for new location. Door prizes each hour, raffle, lots of food, flea market, model aircraft flying, and contests. Identify Mr. Hamfest and win prize. \$5.00 cost covers everything. For further info. contact: John Bruning, W8DSR, 6307 Fairhurst Avenue, Cincinnati, Ohio 45213.

**WRL** carries a complete line of new gear: Galaxy, Drake, Regency, Hy-Gain, and you can use BankAmericard & Mastercharge, or special prices on cash deals. Ask us to quote your needs or for a catalog. Box 919, Council Bluffs, Iowa 51501.

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**SAVE MONEY** on parts and transmitting-receiving tubes. Foreign-Domestic. Send 25¢ for giant catalog. Refunded first order. United Radio Company, 56-HR Ferry Street, Newark, N.J. 07105.

**NEW JERSEY QSO PARTY** August 19-20. Full details from Englewood Amateur Radio Assoc., 303 Tena-fry Road, Englewood, NJ 07631.

**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.

**OPERATING ACHIEVEMENT AWARD CERTIFICATE** signed by Governor Daniel J. Evans for contacts made during Washington State Amateur Radio Week, September 9th through 17th. Out-of-state hams must contact 10 Washington hams and in-state hams 20 other Washington hams during this week. Send your list of stations worked, their locations, dates worked, and your name, call, and address, along with a legal size SASE to: The Puget Sound Council of Amateur Radio Clubs, 12306 80th Ave. East, Puyallup, Washington 98371.

**NEW! PORTABLE** oxygen unit, breathing aid. 12 minutes oxygen, refillable. \$39.95. GUARANTEED. Write Fire Safety Equipment, 232b Windsor Parkway, Hempstead, N. Y. 11550.

**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 information write Larry Higgins, W5QMU, 2522 Old Hickory Trail, San Antonio, Texas 78230, or Gene Chapline, K5YFL, 2206 La Casa, Austin, Texas 78704.

**GONSET IV** — 2 meter transceiver - \$135.00; Gonset III - \$75.00; Gonset II - \$55.00; Poly Comm IV - \$130.00; Clegg 22er - \$135.00; Gonset III - 6 meter - \$80.00. WB2GKF, Stan Nazimek, Jr., 506 Mount Prospect Avenue, Clifton, N. J. 07012.

**WANTED:** HQ215, SPR4, Cygnet. **SELL:** KWM-2, Lafayette Hi/Lo Monitor Receiver. Darryl L. Dippel, WA5AAO, Box 335, LaGrange, Tex. 78945.

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50	.04	.06	.12	.15
100	.06	.08	.16	.20
200	.08	.10	.20	.25
400	.12	.14	.28	.50
600	.14	.16	.32	.58
800		.20	.40	.65
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1 1/2" Square Speaker. 8 ohm VC  
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18" Patch Cord. Has molded RCA type plug on each end. Very nice. 3 for \$1.00 ppd.

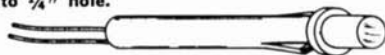


6 Foot Gray Line Cord. UL Approved. 7 Amp Rating. 40¢ ea. ppd.



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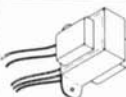
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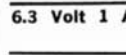
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6.3 Volt 1 Amp Transformer. Fully Shielded. \$1.60 Each ppd.



Rectangular 0 to 1 Mil. Meter. Panel Size Approximately 1 3/8" x 3/4" x 1 1/4" 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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**AM SIGNAL GENERATOR:** AN/URM-25D, 20khz to 50mhz, .5% accuracy, 400 and 1000hz modulation to 50%, crystal calibrator, compact, modern. Restored to original specifications, book included. Signal Generator only, \$150. With all accessories, \$200. Other equipment available, SASE for list. James Walter, 2697 Nickel St., San Pablo, Ca. 94806.

**COLLINS 75S2**, clean, \$270 air mail, HP 425A Microvolt Ammeter, book, probe \$80 air mail. Maj. Don Bohart, Det. 6, 1141 SAS, Box 8895, APO NY 09012.

**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 Raddburn Rd., Fair Lawn, New Jersey 07410 or call 201-796-5414 evenings.

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**FOR SALE — Drake 2B w/Q-mult \$165, HQ180AC w/blanker \$335, Alden 319EA 3-speed fax w/paper \$550. Wanted 51j or R390. WA2DVU, 609-465-5363.**

**FIGHT TVI** with the RSO Low Pass Filter — p115 March 73 — write for brochure — Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada.

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**FOR SALE:** TR-44A rotor \$50.00. NO SHIPPING. W3HZ.

**LINEAR BUILDERS —** Motorola HEP170 diodes 2.5A/1000PIV 30¢ each; Plate Choke 90uh., 5KV/2A \$6.45; Shielded 30A Dual Filament Choke \$6.95; 12,000PIV/2.5A Diode board with RC transient suppressors \$9.95; Diode board unfilled \$2.50. All Postpaid. K.E. Electronics, Box 1279, Tustin, Cal. 92680.

**TELETYPE:** 11/16" perforator tape, 10 rolls/\$3.00, 40 roll case \$10.95; M28 RO base \$12; M14 T.D. \$35; FACSIMILE Converter CV-2A, very clean, \$49; Facsimile Recorder TXC-1, \$110 (as a set, \$145 for both); UHF Receiver (200-400MHz tuneable), R-266A/URR-13, \$60; Newer model with same tuning, R-482C/URR-35, \$75. Jim Cooper, POB 73H, Paramus, NJ 07652.

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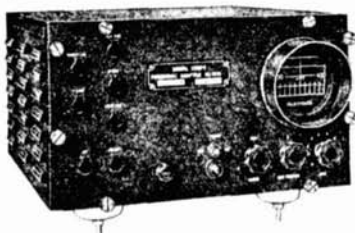
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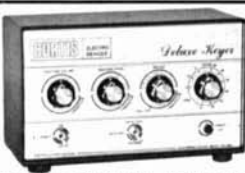
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**SJRA HAMFEST** on Sunday, Sept. 10, 1972, at Mollia Farm, Malaga, N. J., rain or shine. Greatly expanded Swap Shop. Protected pavilion containing tables for Swap Shop and special parking will be provided at the Hamfest for hundreds of cars for those who prefer to display in the trunks of their cars. A variety of contests, games, and displays, plus prizes. Picnic and protected swimming facilities will also be provided to help make a gala event for the entire family. Advance registration at \$2.00 per family (\$3.00 at the gate). Write Fred Holler, W2EKB, 348 Bortons Mill Rd., Cherry Hill, N. J. 08034. Mollia Farm is located just off Route 47 at Malaga Lake, Malaga, N. J.

**THE BEL-ROCK HAMFEST** will be held August 20 at the Boone County Fairgrounds just north of Belvidere, Illinois on highway 76. Free coffee and donuts from 9:30 to 10:00 a.m. There is no set-up fee and plenty of free parking is available. Talk-in on 146.94 MHz. Advance tickets are \$1.00, \$1.50 at the gate. For further information, contact: Lee Gehlausen, WA9WVY, 4610 Cayuga, Rockford, Ill. 61107.

**SSTV, RTTY, CW, WWV, CHU** tone to logic decoder projects. Chassis, plans, hardware, eight 2" x 3" epoxy cards, \$6. Horning, 1630 Bowling Lane, San Jose, Calif. 95118.

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**"MOORY ELECTRONICS COMPANY"** wants "you" as a customer! Write or call "HOSS TRADER" Ed Moory for the best deal on new or used equipment in 1972: New Regency HR-212 two meter FM transceiver, 20 watts, immediate shipment, amateur net \$259.00; New Rohn 50 Ft. foldover tower, prepaid \$239.00; New Mosley CL-33 and demo Ham-M Rotor \$210.00; Used equipment: Drake 2B, \$175.00; HT-37, \$169.00; HQ-180C, \$249.00; HW-16, \$99.00; TR-4 w/34NB, \$499.00; T4-XB, \$359.00; Ham-M, \$85.00; IN STOCK: New Drake, Galaxy, National, Collins, Hallicrafters FPM-300, and lots more. Write for Quotes: Moory Electronics Co., P. O. Box 506, DeWitt, Arkansas 72042. Phone (501) 946-2820.

**TWO METER KILOWATT AMPLIFIERS**, brand new see our ad on page 104 of this issue also Facsimile Transceivers (Western Union Deskfax) see conversion article in QST, May '72. Excellent condition only \$9.95 ea. Fax paper, 2¢ each or pkg. of 250 for \$4.00 (minimum order \$10.00) plus postage. Newsome Electronics, P. O. Box 176-H, Trenton, Michigan 48183.

**AURORA AREA HAMFEST** Sunday, August 27, 1972, Phillips Park Pavilion, Aurora, Illinois. Swap Tables, Picnic Grounds, Playgrounds for the kids, Food, Shelter sponsored by: Fox River Radio League, P. O. Box 443, Aurora, Illinois 60504.

**WX2MAP** will operate at the Miss America Pageant in Atlantic City, NJ, Sept. 1 thru 15, 1972. Will operate SSB 15 kHz inside General Class portion, CW 30 kHz inside bandedge on 80 thru 10 meters. QSL via K2JOX. SASE please.

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HP100D-Freq. stand. w/scope-Acc. 1ppm	165
HP150A - 10mHz Scope w/152B	295
HP160B(USM105)-14mHz dual trace scope	415
HP175A - 50mHz scope w/dual trace	595
HP185B-DC-1gHz samp. scope w/186B	595
HP202B-Audio Gen.-5Hz-50kHz-1-10v	110
HP212A-Pulse Gen. var. width and rate	175
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Nems Clark 1671-FM rcvr. 175-260mHz	210
Polarad R-microwave rcvr (plug-ins avail)	285
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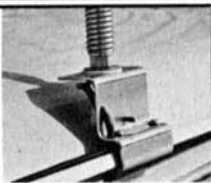
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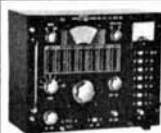


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BE SURE to take advantage of Ham Radio's Summer Sale. Details were sent to you, but if you want another copy of this offer write: Summer Sale, Ham Radio, Greenville, NH 03048.

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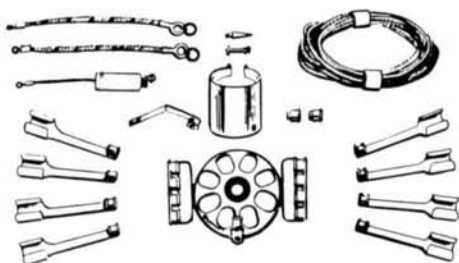
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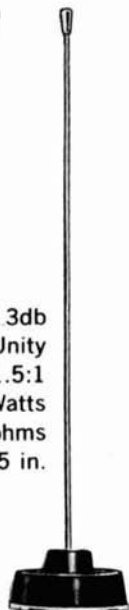
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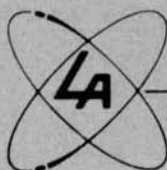
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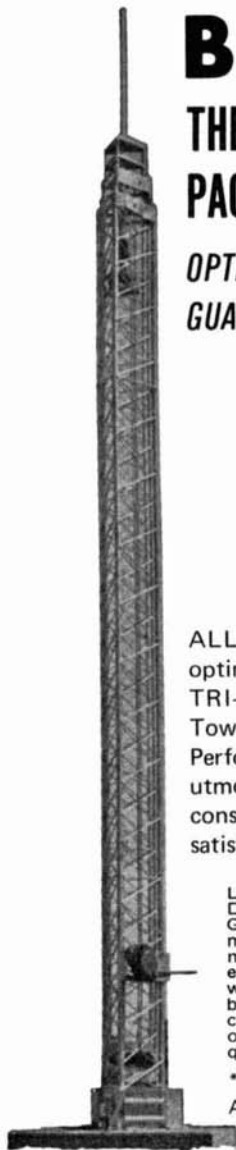
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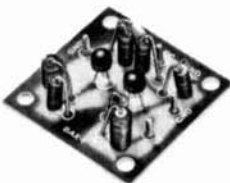
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