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OCTOBER 1973

low-cost electronic keyer with random-access memory

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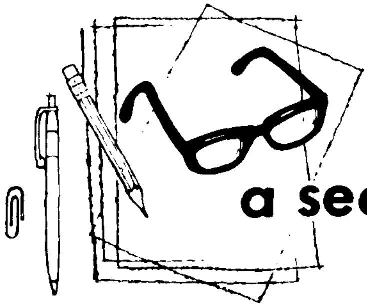


a second look

by jim fisk

The editor of a technically oriented magazine such as *ham radio* must wear several different and diverse hats. In fact, I could use up this entire page describing all the details that need attention to keep the magazine running smoothly. However, I'd like to talk for a moment about one very important editorial task that instills confidence in the reader, and one that doesn't. This task, which I share with the rest of the editorial staff, is that of researcher and seeker of truth.

Most of the articles published in *ham radio* are contributed by readers who want to share an idea or the details of a particularly successful project. Authors range from enthusiastic hams who have never written anything more than a short story for their English professor to fellows with engineering backgrounds who make their livings in front of a typewriter. All want to share an idea and I welcome the output of anyone who is interested in contributing something that will benefit all hams.

Budding authors often ask, "What kind of articles are you looking for?"

That's a difficult question to answer because many new manuscripts come across my desk every day, but generally speaking, I am looking for simple construction projects that the average reader can complete in one or two weekends. Larger projects are also welcome, but most *ham radio* readers must split their spare time between amateur radio and other interests, so they don't have time to build Chinese copies of complex electronics equipment.

Once a month I set one or two days aside to go over all manuscripts that have come in during the previous month. Since I seldom use more than a dozen articles in any issue, I don't accept more than that during any one-month period. This is sometimes a nearly hopeless task since there may be three-dozen or more manuscripts to be considered. The first things I look for are originality and interest value. If the contribution passes this test, the next thing I look for is technical accuracy and attention to detail.

The contributed article doesn't have to be a literary masterpiece to be accepted. If you have a good idea and it's well documented, if the illustrations and technical discussion are clear and accurate — you may have a winner! On the other hand, if the article rambles from one topic to another, covers ground that has been over thousands of times before, or presents inaccurate or misleading information, you will receive a rejection slip.

If your article has been accepted for publication, don't expect to see it published in the very next issue. The production times for a monthly magazine are probably much longer than you ever imagined. The articles for this issue, for example, were being prepared for publication during the month of June. As you are reading this we are putting together the material for the February, 1974, issue of *ham radio.*

Incidentally, my staff and I are fairly adept at ferreting out technical inaccuracies, but despite research and keeping the mailman busy between our editorial office and the author, errors do occasionally creep into the magazine. Contributing authors can help by carefully checking out their facts before submitting the article. Errors can cause considerable misery to the builder, and as publishers *we* are taken to task for the error. So, before you send in that next article, spend some extra time going over the text and illustrations — it might save some later embarrassment.

Jim Fisk, W1DTY
editor

low-cost electronic keyer with random-access memory

Circuit details of an electronic keyer with a 512-bit random-access memory that costs less than $40

Michael J. Gordon, Jr., WB9FHC, 1615 Elmwood, Wilmette, Illinois 60091

Although many circuits for electronic keyers with a memory have appeared in the literature, recent developments in the IC industry have rendered these designs obsolete. These developments include the introduction of MOS random-access memories which are compatible with TTL ICs. It is now possible to build an electronic keyer with 512 bits of storage for a total parts cost of $40.00.* Thanks to the miniature MSI and LSI devices, there are only eight ICs in the unit, which fits on a 2-1/2- by 4-inch circuit board, power supply included!

circuit

The circuit is not very complex and can be divided into two parts: the keyer, which automatically produces dots and dashes, and the memory section and its associated logic.

The keyer circuit is fairly standard and is designed for use with a standard paddle. Speed is variable from roughly 6 to 60 words per minute; the characters are

*A complete kit of parts, including circuit board, is available for $40.00 from Psynexus Systems, P.O. Box 277, Glencoe, Illinois 60022. The drilled glass-epoxy circuit board is available for $4.00. All items post-paid.

self-completing. Details of how this portion of the circuit operates are best understood by looking at the timing diagram, **fig. 1**, and by examining the logic diagram, **fig. 2**. Output from the keyer circuit is taken at two points, labeled DATA and $\overline{\text{DATA}}$ on the logic diagram. The DATA line goes directly to the output gate, where it is inverted and fed to the output keying transistor, a pnp device with a breakdown voltage of 150 volts. The $\overline{\text{DATA}}$ line is fed to the memory inputs.

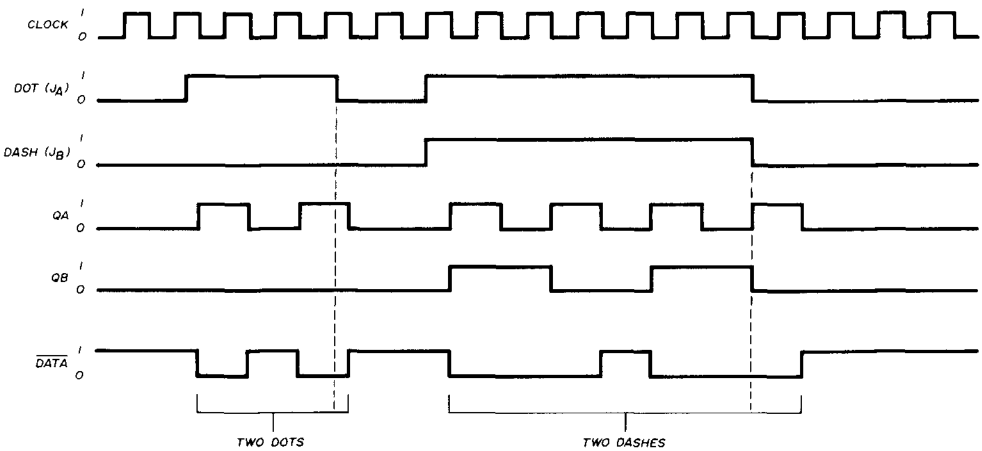


fig. 1. Timing diagram for the electronic keyer. Note that closing the dash contacts puts a logic 1 on the J pins of both flip-flops A and B (fig. 2). Characters are self-completing.

memory

The two memory ICs used in this design are Signetics 25L01B 256-bit random-access memories. Each is an MOS LSI circuit containing thousands of transistors on a single chip. The particular memory cell that is being readout at a given time is determined by the status of the eight address lines. A chip-select control is provided; when a chip is not selected the input buffers are disabled and the output buffers are cut off, effectively taking the device out of the circuit. This allows the two ICs to be connected in parallel, simplifying both circuit board design and the logic requirements. It is this chip-select feature that makes the random-access memory more

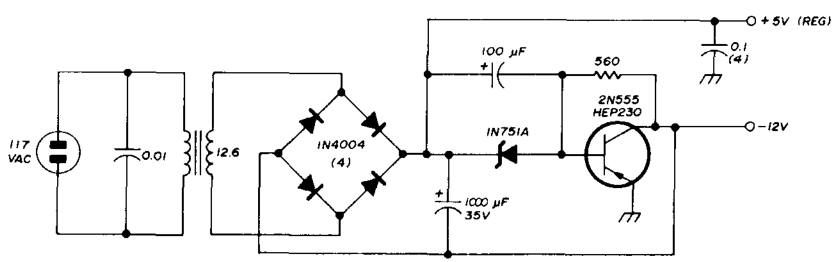


fig. 3. Power supply for the electronic keyer. The negative 12-volt line powers the MOS memory circuits.

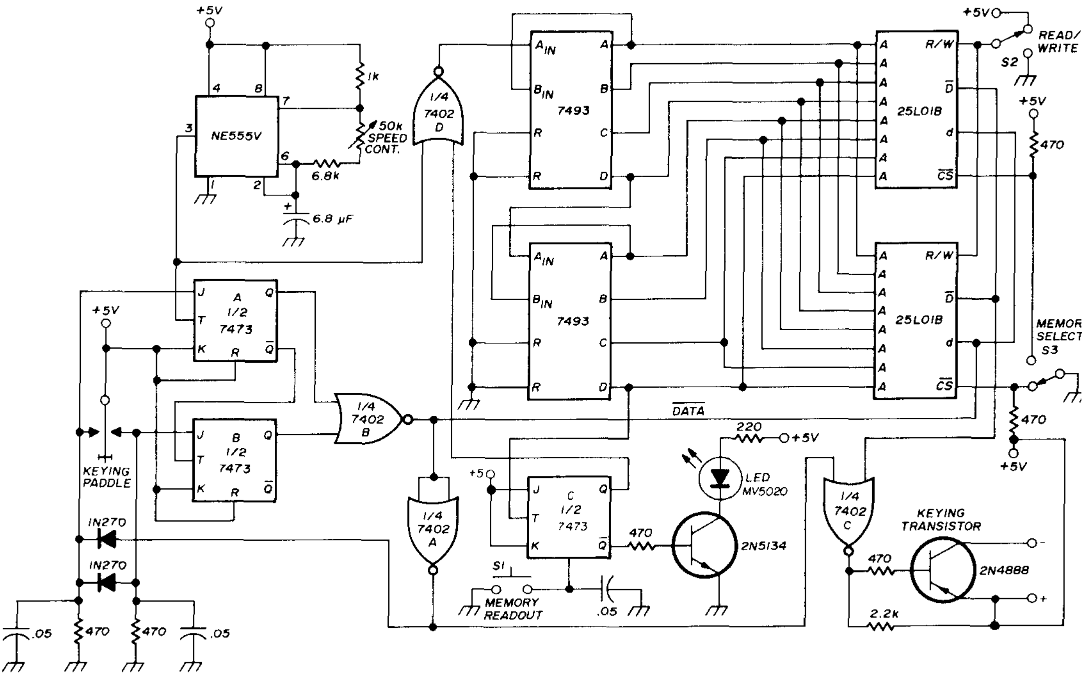
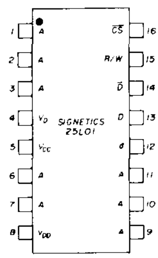


fig. 2. Logic diagram of the low-cost electronic keyer with 512 bits of memory. Memory is provided by two Signetics 25LO1B read-only memory ICs. Breakdown voltage of the 2N4888 keying transistor is 150 volts.



cost-effective than the shift register usually used in this type of circuit.

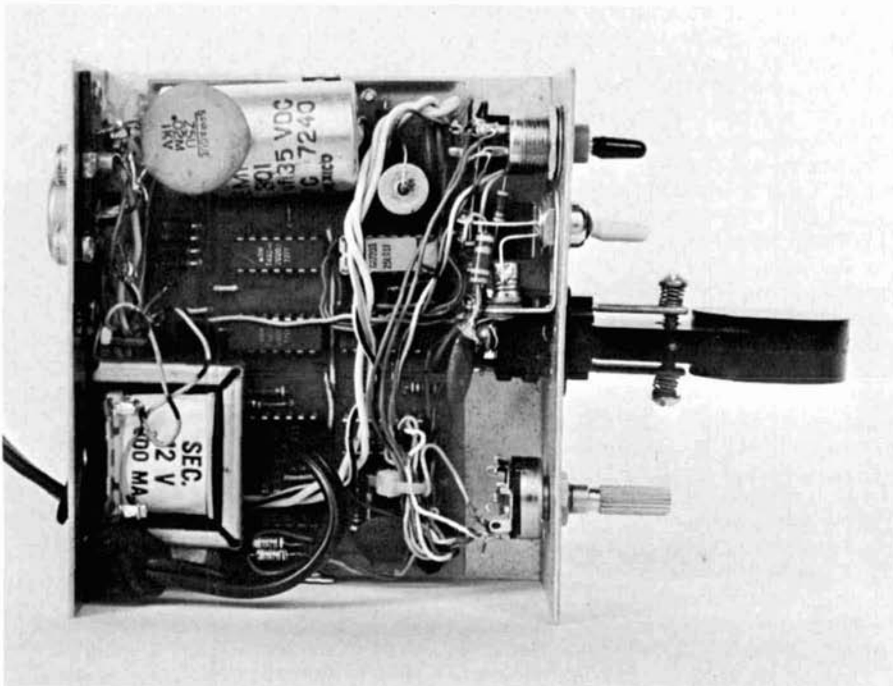
The MSI circuits, both four-bit binary counters, are used to cycle the memory through its 256 bits during the read and write cycle. Normally, the Q output of flip-flop C is in the logic 1 state, and the output of gate D is held low. However, when S1 is closed flip-flop C is reset, Q goes to zero and clock pulses are fed to the memory address counters which cause each of the 256 locations in the selected memory to be accessed.

On the 256th clock pulse, the last stage in the counter triggers flip-flop C – Q once again is returned to the logic 1 state and a memory cycle is complete. During the cycle, Q is in the logic 1 state and the LED is lighted, indicating a memory cycle in progress. If the read/write switch, S2, is in the read position during a cycle, the data in the memory is sent to the output gate and thus keys the transmitter. If this switch is in the write mode, whatever is sent by the paddle is recorded by the selected memory as well as sent to the output gate. Switch S3 selects one memory of the two available.

Clock pulses for the entire circuit are generated by a Signetics NE555V timer IC. The speed is variable over an extremely wide range by a single 50k pot. The NE555 is an excellent choice for this application because of its TTL compatible output, small size and low power consumption.

power supply

The power supply is simple, efficient and effective (see **fig. 3**). It consists of a full-wave bridge feeding a discrete regulator. The voltages necessary for the proper operation of this circuit are +5 volts and −12 volts. The most positive point in the



Construction of the electronic keyer with memory. Timing and memory components are mounted on circuit board. Power supply components are mounted on rear wall of the enclosure.

circuit is called +5, and the regulator keeps ground five volts negative with respect to this point. The most negative point, the other side of the filter capacitor, is roughly 12 volts below ground. Since the 12-volt supply is not really critical, it is not regulated. As long as it provides between −10 and −13 volts, the circuit should operate properly. Four 0.1-μF ceramic bypass capacitors are placed at various points on the circuit board between V_{cc} and ground to absorb current spikes generated by the TTL output logic.

operation

Completely assembled, this electronic keyer is a joy to operate. Programming is simply a matter of pressing the start button and sending the desired message. Readout is even simpler, requiring only the push of a button. There is no need to switch between the manual and automatic mode of operation, since this is done automatically by the logic. Now, you can have your own keyer with 512 bits of storage, at a cost that is a far cry from the \$200 and up that commercial units command.

ham radio

"Look, 'Mr. Ham Operator,' if you want more coffee just ask for it, and stop tapping out -.-. --- ..-. ..-. . . on your cup."

continuous-phase audio-shift keyer for RTTY

Circuit details for an audio-frequency shift keyer which introduces no phase disruptions when switching from mark to space, or vice versa

Eric Kirchner, VE3CTP, Ontario Science Center, Don Mills, Ontario

The desirability for the absence of phase discontinuities in the output signal of audio frequency-shift RTTY keyers has been stressed in previous articles.[1,2] These phase discontinuities appear at that point in time when the AFSK generator frequency is shifted from mark to space, or vice versa. The switch-over from one frequency to the other appears as a disruption along the sinusoidal waveform, as shown in **fig. 1A**.

A phase disruption such as this causes over- and under-shoots which manifest themselves as fast amplitude changes of the transmitted rf envelope. This leads to

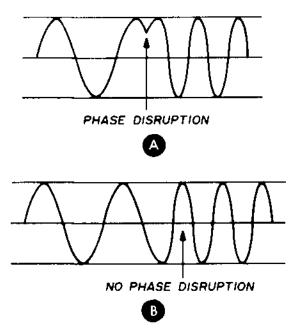


fig. 1. In many AFSK generators severe phase discontinuities are introduced when switching from mark to space, or vice versa, as shown in A. This results in undesirable clicks, similar to CW key clicks. With the AFSK generator circuit described here there is no phase discontinuity when shifting frequencies, B.

clicks, similar to CW clicks, which can be heard on either side of the RTTY signal. Needless to say, these clicks interfere with stations operating on adjacent frequencies.

Although AFSK generator circuits which eliminate these phase discontinuities have been described in the past, they were complex and expensive, and used toroid inductors. The circuit described here was designed with simplicity and state-of-the-art in mind, and uses no toroid inductors.

circuit

The circuit for the continuous-phase AFSK generator is shown in **fig. 2.** Integrated circuits U1 and U2 constitute the audio-frequency oscillator. The output of U2, a National Semiconductor LM311H, is a square wave which is fed through R4 to the input of U1, a National Semiconductor LM301AN. U1 operates as an active filter whose frequency is determined by C1, C2, R1 and R2. The output of the active filter is capacitively coupled to the input of U2. The loop is closed and oscillation occurs.

The sine-wave output is available at pin 6 of U1. This sine wave crosses the zero voltage point at precisely the same time the square wave at pin 7 of U2 changes polarity as shown in **fig. 3.** This square-wave transition can therefore be used to command the switchover from one audio frequency to the other to occur *only* at the zero-voltage crossover of the sine wave. This will provide a phase-continuous output.

The switchover command is accomplished in the following way. The square-wave output from U2 is fed to transistor Q2 which operates as a voltage-level changer. The output swing from U2 is about ±10 volts, while the maximum input requirement for U3, a TTL master-slave J-K flip-flop, is from zero to +5 volts. The square-wave signal at the collector of Q2 toggles the input of U3. When the keyboard contacts connected to the input of gate U4 are opened, the polarity of the voltage at pins 6 and 8 of U4 invert, and the Q output of U3 will change its state then, and only then. This

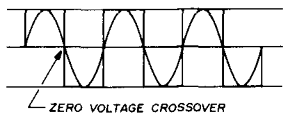


fig. 3. In the circuit of fig. 2 the zero-voltage crossover of the sine-wave output coincides with the zero-voltage crossover of the square wave. The square-wave transition is used to control the precise moment mark-space switching occurs.

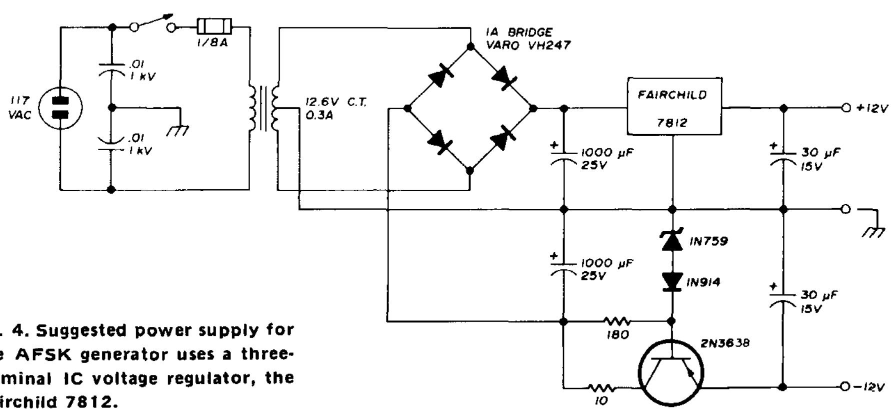


fig. 4. Suggested power supply for the AFSK generator uses a three-terminal IC voltage regulator, the Fairchild 7812.

only happens when U3 is toggled by the negative-going transition of the square wave at Q2, which occurs at the zero-voltage crossover of the sine wave. At that instant the Q output of pin 8 of U3 goes low and brings transistor Q1 into conduction.

Transistor Q1 operates as a switch, effectively paralleling the resistance network R1 and R6 with an additional resistor, increasing the frequency of the audio-frequency generator. When the keyboard contacts are closed again, the same sequence occurs in reverse, switching back to the lower frequency at precisely the zero-voltage crossover point of the sine wave. In this way phase discontinuities are avoided and the AFSK sine wave is sufficiently pure for use with ssb transmitters.

The audio frequencies I chose were dictated by a desire to eliminate the need for a special carrier crystal in my ssb exciter. The lowest audio frequency of this keyer is at 1450 Hz, high enough to place the second harmonic at 2900 Hz, out of the passband of most modern amateur ssb transmitters. The highest audio frequency, at 2300 Hz, lies within the passband of this same equipment.

The AFSK generator requires +12 volts at 70 mA and −12 volts at 15 mA. A suitable power supply circuit is shown in fig. 4.

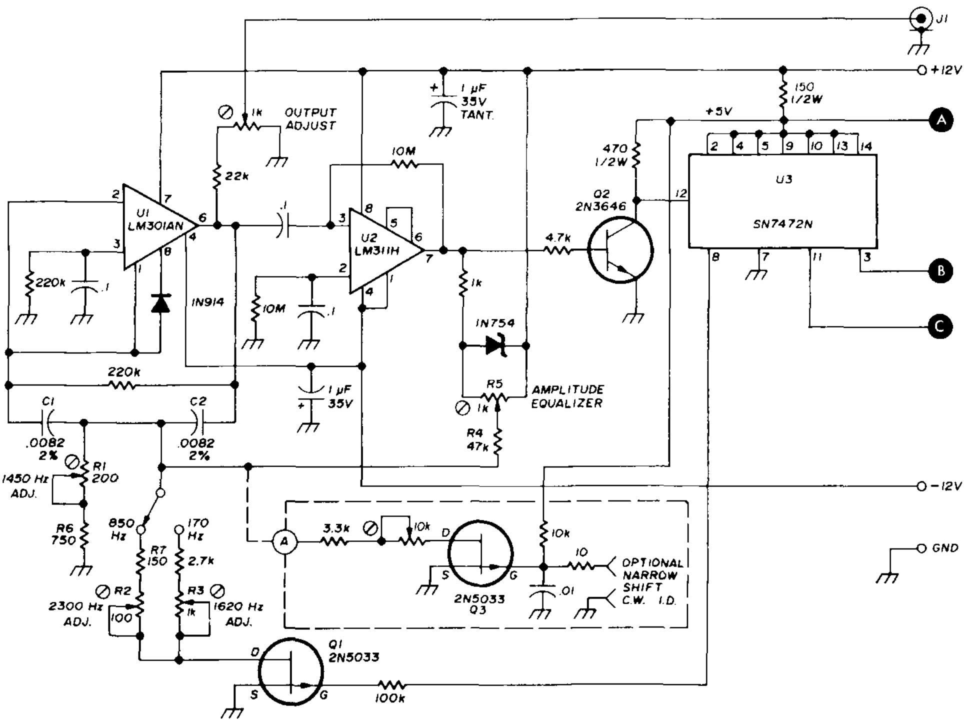


fig. 2. Circuit for the continuous-phase AFSK generator. Capacitors C1 and C2 are polystyrene types. All ICs are manufactured by National Semiconductor.

alignment

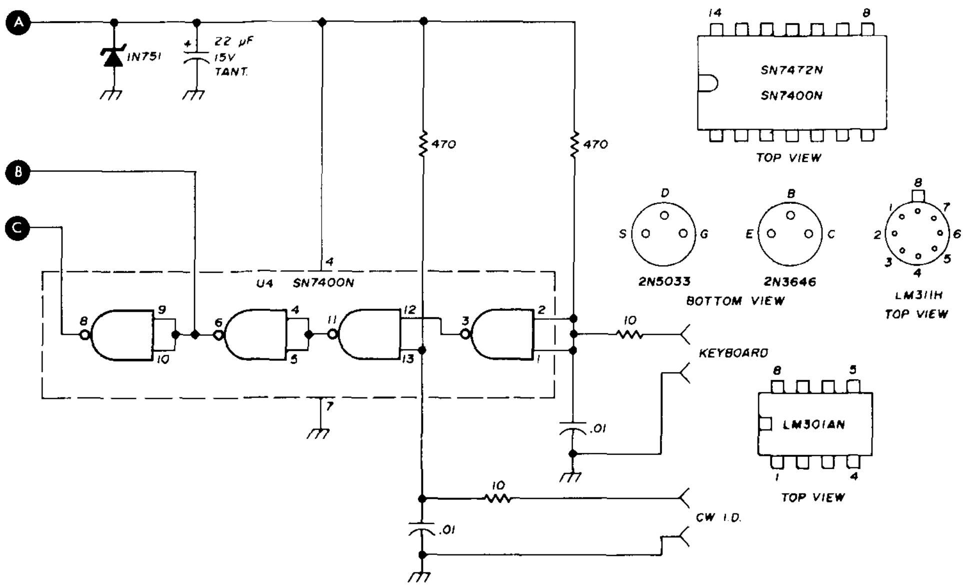
The best way to adjust the frequency and amplitude equalization of this AFSK generator is by using a digital frequency counter and an oscilloscope. These instruments are now owned by many amateurs so you should be able to enlist some help, if you don't personally own this test equipment. Once the alignment is completed it should hold for a long time.

First, connect the frequency counter

to the output jack, J1. Short the keyboard input terminals and adjust R1 for 1450 Hz on the counter. Set the switch to 850-Hz shift and adjust R2 for 2300 Hz. Set the switch to 170-Hz shift and adjust R3 for 1620 Hz. For easier adjustment you may want to install more expensive multi-turn trimming potentiometers at R1, R2 and R3.

optional circuit shown within the dashed lines in **fig. 3.** Simply connect point A of the optional CW identification circuit to point A in the main AFSK generator circuit.

Although the input terminals could be connected directly to the keyboard contacts, in most RTTY station setups they are not. Since it is desirable to copy the



outgoing transmission on the printer, I have a mercury-wetted relay in series with the printer solenoid – the relay contacts are used to key the AFSK generator. A suitable relay for this purpose is the Potter and Brumfield JML-1061-81 with a 33-ohm resistor connected in parallel with the relay coil.

Now, disconnect the counter from the output jack and connect a scope probe to pin 6 of U1. Set the switch to 170-Hz shift and, while simultaneously opening and closing the keyboard input terminals, adjust R5 so that there is no amplitude difference shown on the scope. There may be a slight amplitude difference when switching to 850-Hz shift, but this should be on the order of 2 percent or less, which is not objectionable.

additional notes

The CW identification circuit, as shown, results in the same frequency shift as the selected RTTY shift, 170 or 850 Hz. If you want to use a different CW identification shift, you can add the

references

1. D.H. Phillips, W6FOO, "Radio Teletype using SSB Transceivers," *ham radio,* November, 1970, page 38.
2. D.H. Phillips, W6FOO, "Synchrophase AFSK Oscillator for RTTY," *ham radio*, December, 1970, page 30.

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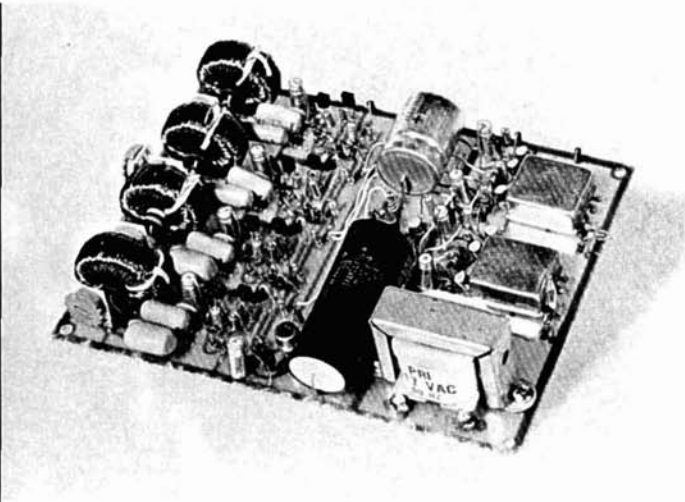


multi-function touch-tone decoder

Complete construction details for a solid-state, multi-function Touch-Tone decoder using latching relays

Robert Heptig, KØPHF, Robert Shriner, WAØUZO, Pueblo, Colorado

With the advent of FCC Docket 18803, all repeaters are faced with the problem of shutdown control that is now defined by the FCC. The decoding system described here was designed to give absolute security in repeater control – *it cannot* be activated by anything other than a true Touch-Tone signal.

Two functions are utilized with latching-type relays. In the event of power failure the relays will remain in the position in which they were placed by the control operator. A time-out timer is also available in the circuit so that one of the functions will time out if desired.

For those readers who do not understand how Touch-Tone works, a short explanation seems to be in order. The Touch-Tone system was originally designed for dialing telephone services. It uses two tones for each digit, zero through 9 as well as * and #. These tones are generated by an oscillator in the

The lead photo shows a complete Touch-Tone decoder built by a Pueblo Ham, Jim Warner WBØ BTA, for use in muting his speaker when his wife is asleep. He's figuring on hooking up the second function to the coffee pot.

Touch-Tone pads. **Fig. 1** lists the frequencies of these tones. Note that only 7 tones are used. The digit 1 is composed of 1209 HZ and 697 Hz, while the digit 2 is composed of 1336 Hz and 697 Hz, and so on, throughout the twelve digits which make up the Touch-Tone pad.

These tones are combined within the pad itself by various switch points. It is the job of the decoder to separate the various tone frequencies on receipt of the signal. If you don't thoroughly understand this go back and read it again, as you must understand how the ten Touch-Tone digits are made up of combinations of 7 tones. Also note that for explanation purposes, each tone is assigned a letter from A thru G. These letters that I have assigned to the various tones will be used for the balance of this discussion.

To use the Touch-Tone system, these tones must be decoded, changed into a dc voltage and combined in an AND gate which will form digits. Then, two of the digits are combined and used to close a relay. What you do with this relay closure is up to you. It may be used to shut down your repeater, disconnect the speaker of your base station, turn on your porch lights, start the coffee perking or whatever your imagination can conjure up.

There are many circuits that can be used to decode tones and cause a relay to open or close. One of the newest circuits is the phase-locked loop (PLL). This circuit was tried in many different configurations but its primary fault is that it will accept tones that are not on the exact tone frequency. Of course, this is an asset in some applications other than Touch-Tone decoders. The broad frequency response of the PLL can also cause problems if the repeater users use a Touch-Tone autopatch system – can you think of a better way to make the telephone company unhappy than to send a batch of off-frequency tones over their lines?

Other circuits using transistor decoders don't work too well as Touch-Tone decoders because of the inherent low impedance of a transistor, necessitating quite tight coupling of the tuned circuit which lowers the Q of the circuit.

The circuit shown in **fig. 2** uses the high impedance of the field-effect transistor. This permits the use of very small coupling capacitors (.01 μF). The resulting Q of the circuit is sufficiently high to assure that the tones are on frequency. The high circuit Q also prevents accidental functions from occuring due to noise or excessive audio on the repeater.

A 1209 Hz	B 1336 Hz	C 1477 Hz	
1	2	3	D 697 Hz
4	5	6	E 770 Hz
7	8	9	F 851 Hz
*	0	#	G 941 Hz

fig. 1. Twelve-button Touch-Tone pad showing tone frequencies and letter designations used for the purpose of this article.

the circuit

Now, let's get down to the nitty gritty of the circuit and see just what happens when the decoder sees a batch of frequencies coming in the front door. First of all, all frequencies are passed through potentiometer R1, which operates as a signal level control, and then fed to potentiometers R2, R3, R4 and R5. These 2.2-meg variables are also used as level controls, but only for the particular frequency of the circuit which they are in. When this multitude of frequencies reaches the 88-mH coil L1, and tuning capacitors C5 and C6 all except the selected frequency are passed straight to ground. The selected frequency is controlled by the tuning of L1, C5 and C6, which will be explained later. This frequency is tone A or 1209 Hz and is fed to the gate of Q1.

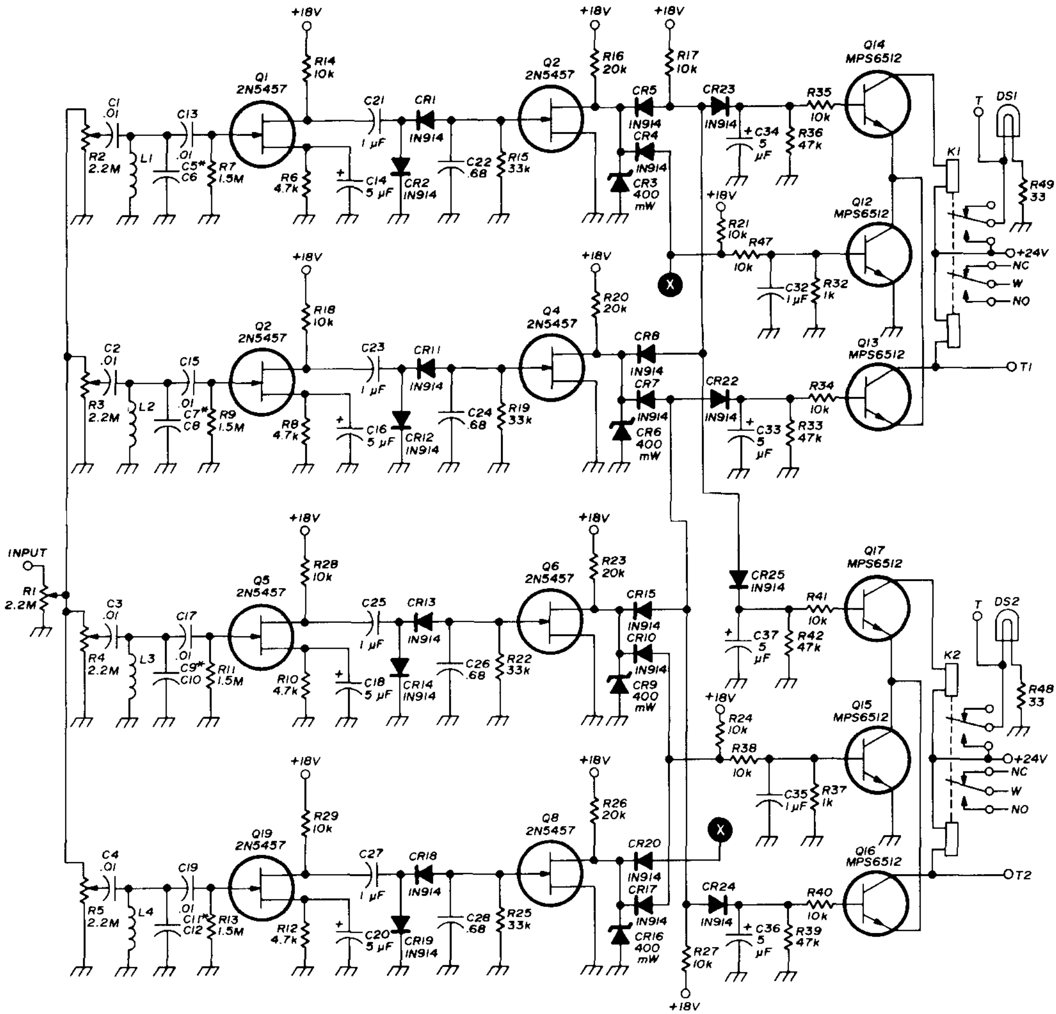


fig. 2. Schematic of the Touch-Tone decoder. Relays K1 and K2 are latching relays such as the Potter & Brumfield FL11D or those available from Circuit Board Specialists.* Capacitors marked with an asterisk are selected for the exact frequency of the desired Touch-Tone tones.

The fet is merely an ac amplifier which boosts the signal up to levels that are easier to work with. The amplified ac signal is rectified by CR1 and CR2, filtered by C22 and is seen on the gate of Q2 as a negative dc voltage. Fet Q2 is connected as a dc amplifier. Therefore, any small change of voltage on the gate will pinch-off the fet which normally conducts the voltage on the drain to ground through dropping resistor R16.

When pinch-off occurs, Q2 ceases conduction and the drain immediately goes positive. This meets one of the requirements of the AND gate made up of CR5 and CR8. Now, go back a little and assume that at the same time tone A was being decoded, tone G was being decoded in the same manner by Q2 and Q4. In this case there would be an immediate rise in

*Etched, drilled, silver-plated printed-circuit boards are available from Circuit Board Specialists, P.O. Box 969, Pueblo, Colorado 81002, $8.50. Relays, latching type, surplus and guaranteed are $2.00.

voltage at the cathode of CR8. This meets the second requirement of the AND gate, allowing the voltage at the junction of R17 and CR23 to go to a positive level. This positive voltage passes through CR23 and is impressed upon the base of Q14 This passes all the current in the relay coil to ground, closing the contacts and performing the function. Since relay K1 is a latching relay, it is not necessary to hold it closed. Once latched, the relay stays closed until another signal releases it. This

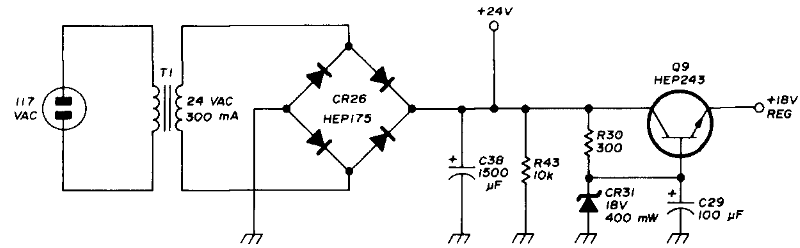


fig. 4. Power supply for the Touch-Tone decoder. Transformer T1 is an Archer 273-1386 (Radio Shack) or equivalent.

and is stored there by capacitor C34 just long enough to keep transistor Q14 turned on.

Although this all happens when you depress one digit on the Touch-Tone pad, the relay still has not been picked up. To do this you must depress another button and decode two more tones. Let's assume that you depressed the digit 7 (tone A and tone F). Tone A would meet one of the requirements of the gate at diode CR4 while tone F would meet the other requirement at diode CR20, providing a positive voltage at the junction of R21 and R47 which is fed to the base of transistor Q12.

Transistor Q12 will turn on heavily and, in effect, ground the emitter of Q14.

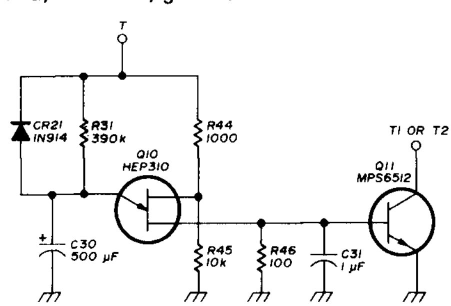


fig. 3. Time-out timer for the Touch-Tone decoder. Resistor R31 is selected for the desired time; value shown is for about 3 minutes.

is done in much the same manner as the initial closure. Through different decoding tones a voltage is imposed on the base of Q13 and stored in capacitor C33. This voltage turns on transistor Q12 again, thereby energizing the off coil or relay K1, restoring the relay and the function back to normal.

time-out timer

Suppose that the function which you wanted to perform with your decoder required a time-out timer, such as 3 minutes for an autopatch hookup. This is done very easily and inexpensively by unijunction transistor Q10 in **fig. 3.** When relay K1 closes a voltage on one set of the contacts it turns on PL1, letting you know that it all worked. This voltage is also fed through point T of the relay to point T of Q10. This voltage is dropped through resistor R31 and allowed to gradually build up across capacitor C30. At a certain predetermined level or time lapse transistor Q10 goes into conduction, turning on Q11 for a split second and energizing the off coil of relay K1 thereby timing out the function. Diode CR21 is used for a fast bleed off of C30 so that the timer will be ready for the next time function.

ham radio

two-band antenna matching with stubs

Complete design details for an antenna stub matching system for two harmonically-related amateur bands

Bill Johnson, W6MUR, Box 803, Ben Lomond, California 95005

With the sun spot cycle going down there will be greater interest in 40, 80 and 160 meters for DX work in the next few years, bands where rotary beams are difficult if not impracticable. Wire antennas still have their place, and it is desirable, if possible, to build antennas that will have reasonable directional characteristics on at least two bands. For example, the familiar dipole fed with open-wire line operates as two half-waves in phase on the second harmonic and exhibits slight gain in the broadside direction. The popular W8JK array can also be operated on two harmonically-related bands.

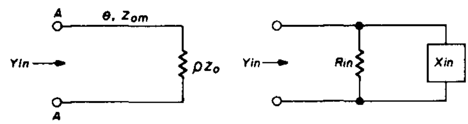
The principal problem with the standard arrangement is that the center impedance of such an antenna at the harmonic becomes quite high, on the order of several thousand ohms, and the bandwidth also narrows; that is, the reactance and resistance change is quite large around resonance. Thus, there is a very high vswr on the open-wire line. While high vswr does not result in appreciable *loss* on a good open-wire line, it does lead to problems in maintaining good balance to ground and in minimizing radiation from the transmission line itself.

High vswr also complicates the antenna tuner in going from one band to another; depending on the length of feeder, it may be necessary to switch from series tuning on one band to parallel tuning on the other, and band changing becomes complicated. Then, there is the matter of the high voltages and currents along the mismatched line. Even a 600-ohm *flat* line at maximum power will have about 600 to 700 volts rms of rf across it; when the vswr is high this voltage will be appropriately higher. High voltages and currents along a transmission line can cause problems in coupling into other lines (such as telephone lines and tv lead-ins) and in arcing to nearby objects. In short, a low vswr is very desirable, even on an open-wire transmission line.

stub matching

It is possible, using stubs and matching sections, to design an antenna that can be operated on two bands, say 80 and 40 or 160 and 80, so that a reasonable match will be achieved to a specified open-wire line without switching. This particular problem hasn't been treated in the various antenna handbooks where stub matching for only one frequency is discussed.[1]

The analytical approach to this problem can get pretty complicated, and in fact, intractable unless it is done in the right way. The solution lies in using the familiar transmission-line equations in admittance, rather than impedance form, and further, in using half-angle rather than double-angle formulas when going between bands. That is, the electrical length of the matching section and stub is defined at θ at the harmonic frequency, say 7 MHz, and the corresponding length at the fundamental, 3.5 MHz in this case, becomes $\theta/2$.



$$\frac{Z_{om}}{Z_{in}} = Y_{in} = G_{in} + jB_{in} \qquad G_{in} = \frac{Z_{om}}{R_{in}} \qquad B_{in} = \frac{Z_{om}}{X_{in}}$$

$$G_{in} = \rho \frac{1 + (\tan\theta)^2}{\rho^2 + (\tan\theta)^2} \quad (1) \qquad B_{in} = \tan\theta \frac{\rho^2 - 1}{\rho^2 + (\tan\theta)^2} \quad (2)$$

$$\text{at } f \quad G_{in1} = \frac{2\rho_1}{\cos\theta\,(\rho_1{}^2 - 1) + (\rho_1{}^2 + 1)} \quad (3) \qquad \text{at } f \quad B_{in1} = \frac{\sin\theta\,(\rho_1{}^2 - 1)}{\cos\theta\,(\rho_1{}^2 - 1) + (\rho_1{}^2 + 1)} \quad (4)$$

$$\text{at } 2f \quad G_{in2} = \frac{\rho_2}{(\cos\theta)^2\,(\rho_2{}^2 - 1) + 1} \quad (5) \qquad \text{at } 2f \quad B_{in2} = \frac{(\sin\theta)(\cos\theta)(\rho_2{}^2 - 1)}{(\cos\theta)^2\,(\rho_2{}^2 - 1) + 1} \quad (6)$$

$$\text{where } \rho_1 = \frac{R1}{Z_{om}} \text{ and } \rho_2 = \frac{R2}{Z_{om}}$$

fig. 1. Basic relationships of open-stub impedance matching at two frequencies, f and 2f.

Fig. 1 gives the basic relationships, the *input admittance of a lossless transmission line having characteristic impedance* Z_o and length θ. In stub matching, the stub (whether shorted or open) is con-

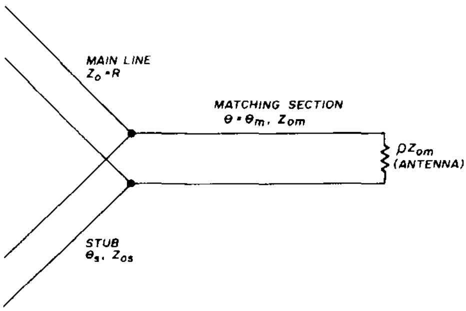


fig. 2. The matching problem.

nected a distance, θ, back from the load or from a known minimum or maximum in the standing wave of the *unmatched* line. The reactance of the stub is made equal and opposite to the input reactance of the matching section at that point.

For two-band matching the problem is depicted in **fig. 2.** The problem is to find a matching section and stub such that the input impedance with the stub connected, at the point of connection, is resistive and equal to $Z_o = R$ of the main transmission line on each of the two bands. To do this, you must also know the antenna impedance at both frequencies. This is assumed to be resistive, justified on the basis that the antenna will be pruned to resonant length or at least carefully calculated.

Resistance R1 is defined as the antenna resistance at frequency f, and R2 as the antenna resistance at frequency 2f.

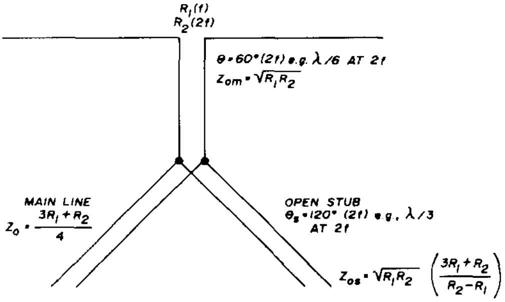


fig. 3. Simplified Johnson match for two-frequency antenna impedance matching.

Using the appropriate trigonometric identities, **eq. 3** through **eq. 6** in **fig. 1** may be derived from **eq. 1** and **eq. 2.** These are the basic relationships for two-band matching. There are four unknowns, Z_{om}, the characteristic impedance of the matching section, Z_{os}, the characteristic impedance of the stub, θ_m, the electrical length of the matching section ($\theta = \theta_m$), and θ_s, the electrical length of the stub.

Resistance R1 has previously been *specified* as the antenna resistance at frequency f. R2 is the antenna resistance at 2f. When the reactance has been tuned out by the stub, the resistance R is equal to

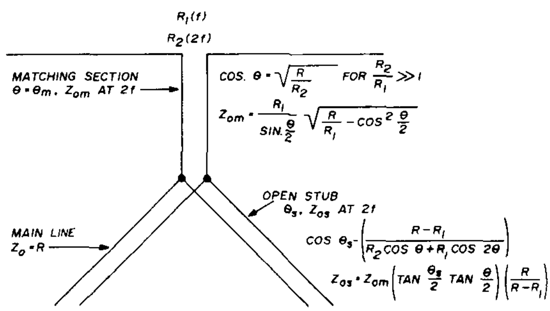


fig. 4. Generalized Johnson match for two bands where R2, the antenna impedance on the higher frequency band, is much greater than R1, the antenna impedance on the lower band, the usual case.

$$R = \frac{Z_{om}}{G_{in1}} = \frac{Z_{om}}{G_{in2}}$$

Where R is the desired match to the main transmission line impedance, Z_o, as shown in **fig. 2.**

special case

Before proceeding to the general case, a most interesting special case is where

$$\rho_1 = \frac{1}{\rho_2} = \rho \text{ or } Z_{om} = \sqrt{R1R2}$$

For this case **eq. 3** is set equal to **eq. 5** to provide the following simple quadratic expression

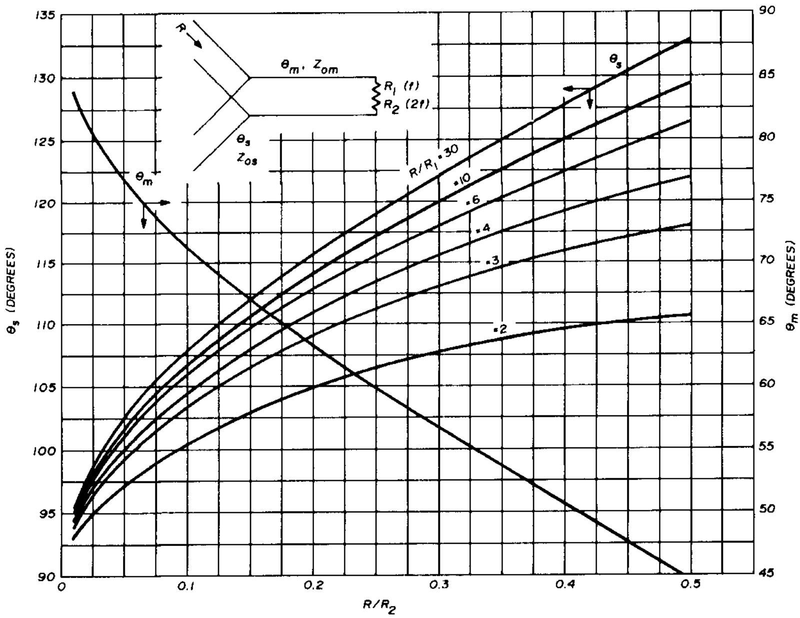


fig. 5. Electrical length of the matching section, θ_m, and the stub, θ_s, at frequency 2f where R2 is much greater than R1.

$$2C^2 + C - 1 = (2C - 1)(C + 1)$$

$$C = \tfrac{1}{2} \text{ or } C = -1$$

$$\text{where } C = \cos\ \theta$$

For this special case $\theta = 60°$ or $180°$ at 2f. The latter case is the familiar quarter-wave section at f, half-wave section at 2f. The input impedance at points A-A in **fig. 1** is simply R2 at both frequencies. The problem here is that R2 is usually high, so the main line impedance to match it is unreasonably high. The other case, θ =

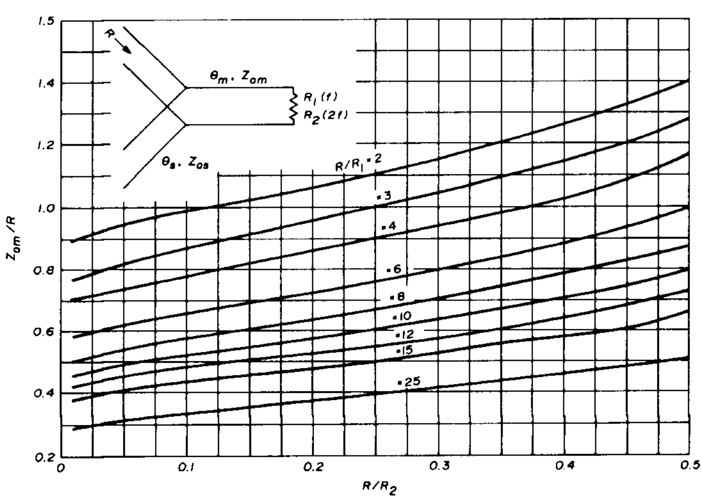


fig. 6. Characteristic impedance of the matching section, Z_{om}, where R2 is much greater than R1.

60° at 2f, is much more useful. For this case the result is

$$R = \frac{3R_1 + R_2}{4} \qquad (7)$$

To tune out the reactance in this $\theta = 60°$ case requires an open stub exactly 120° long at 2f with a characteristic impedance of

$$Z_{os} = 400\left(\frac{3 \times 80 + 2000}{2000 - 80}\right) = 467 \text{ ohms}$$

θ_m = 60° at 2f (1/6 wavelength)

θ_s = 120° at 2f (1/3 wavelength)

You would have a perfect match for a 560-ohm line on both frequencies f and

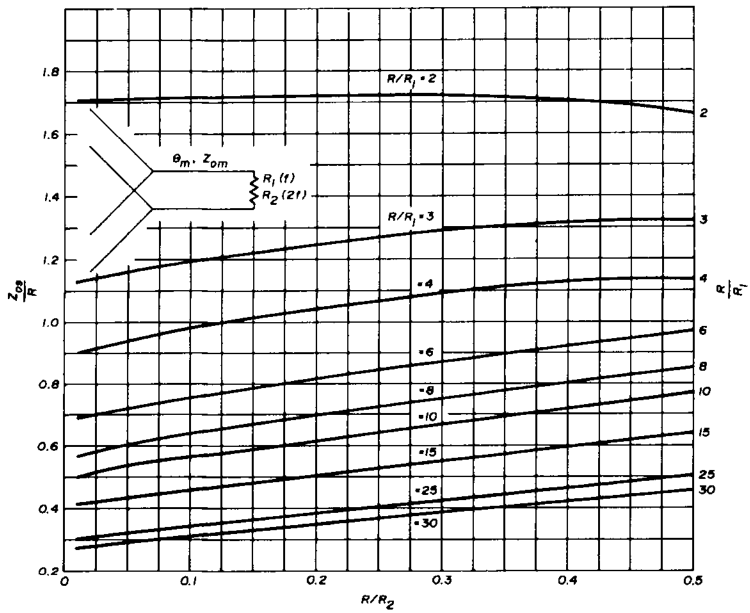


fig. 7. Characteristic impedance of the stub, Z_{os}, where R2 is much greater than R1.

$$Z_{os} = \sqrt{R_1 R_2}\,\frac{3R_1 + R_2}{R_2 - R_1} \qquad (8)$$

The simplified *Johnson Match* is shown in **fig. 3.** Note that the main line characteristic impedance $Z_o = R$ is *not* arbitrary in this case because the quantity Z_{om} has been specified rather than treated as an unknown.

As an example, suppose you had an antenna such that R1 = 80 ohms and R2 = 2000 ohms. Then

$$R = \frac{3 \times 80 + 2000}{4} = 560 \text{ ohms}$$

$$Z_{om} = \sqrt{80 \times 2000} = 400 \text{ ohms}$$

2f by connecting the two 400-ohm and 467-ohm lines as shown in **fig. 3.**

general case

In the special case just considered, you do not have an arbitrary choice of main line characteristic impedance, at least if you want a perfect match. R2 may be considerably higher than the 2000 ohms assumed in the above example, and thus, R may still be unreasonably high, In the general case you first use **eq. 3** and **5** to find Z_{om} and Cos θ (θ and θ_m are used interchangeably here). Then, **eq. 4** and **6** are used together with the stub reactance formula to find θ_s and Z_{os}. The result, when R2 is much larger than R1, the usual case, is given in **fig. 4** as the generalized *Johnson Match.*

To simplify finding the required lengths and characteristic impedances of the line sections, **figs. 5, 6** and **7** have been prepared using the approximate formulas given in **fig. 4** which are valid when R2 is much larger than R1. In **fig. 5** are given the electrical lengths of the matching section and stub in degrees at the harmonic frequency 2f. In **fig. 6** is the Z_{om} family giving the required characteristic impedance of the matching section, and in **fig. 7** is the Z_{os} family giving the required characteristic impedance of the open stub.

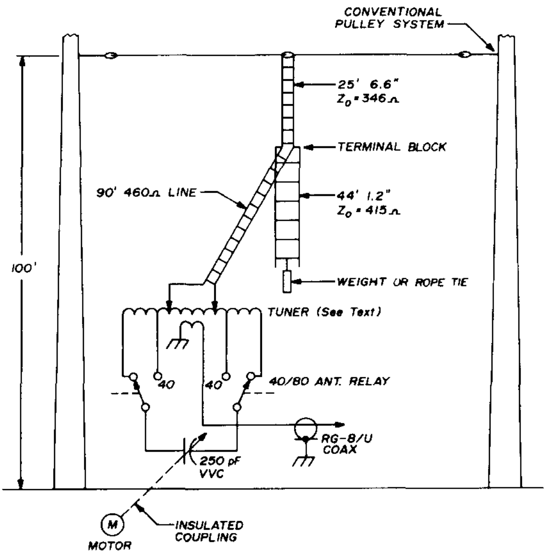


fig. 8. Two-band 3.5- and 7.0-MHz antenna system used at W6MUR incorporates the stub matching system described in the text.

As an example in using these curves, assume an antenna where R2 = 4000 ohms at 2f and R1 = 60 ohms at f. Also assume you want matching on both frequencies to a 600-ohm line, R = 600. Then, compute R/R2 = 0.15 and R/R1 = 10 and enter the curves to find

from **fig. 5**, at R/R2 = .15 and R/R1 = 10, $\theta_m = 67.2°$, $\theta_s = 111°$ at 2f;

from **fig. 6** find Z_{om}/R = 0.55 and compute Z_{om} as 600 x .55 = 330 ohms; and

from **fig. 7** find Z_{os}/R = 0.59 and compute Z_{os} as 600 x .59 = 355 ohms.

antenna input resistance

An excellent, if somewhat obscure, source for accurate information on the center reactance and resistance of antennas in the vicinity of resonance is given in reference 2. This excellent book, by the way, also has sufficient information in graphical form to enable you to design a very good three-element parasitic beam for either best front-to-back ratio or for maximum forward gain (the two do not coincide) without guesswork as to the lengths of driven element, reflector and director. On pages 20 through 25 of this reference will be found some curves giving the center impedance of "nearly half-wave" and "nearly full-wave" center-fed antennas for various conductor thickness. Additionally, the effects of spaced multiple wires are also given.

The effect of ground must also be considered in estimating antenna resistance. With city-lot installations and low heights, this can get pretty indefinite, but if the antenna is reasonably in the clear, curves such as those given by Kraus[3] can be used. It is noted, for example, that the half-wave antenna, when 0.34 wavelength high, has a resistance not of 73 ohms as in free space, but close to 100 ohms. When the half-wave antenna is low, say 0.1 *wavelength high, the resistance drops to* something like 23 ohms.

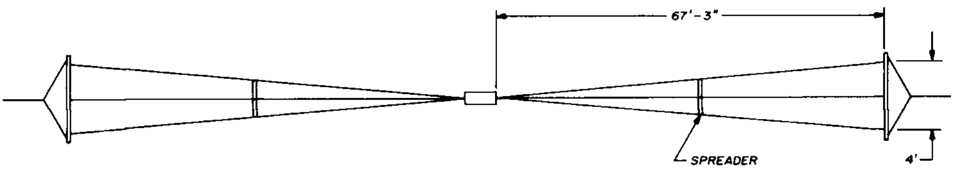
test results

The antenna shown in **fig. 8** and **9** has been installed and tested at W6MUR for over two months with excellent results when this article was written. The antenna resistances were estimated for the multiple-wire arrangement at 88 ohms at 3.5 MHz and 2440 ohms at 7.0 MHz, and the matching section and stub were proportioned accordingly. The antenna tuner is a simple parallel-resonant circuit having an edge-wound ribbon coil (surplus) of 20 turns, 4¼ inches in diameter, 8-inches long (about 18 microhenries), tuned by a 250 pF vacuum-variable capacitor driven through an isulated coupling by a reversible gear motor turning a few rpm. The antenna was tapped on to the coil and a fixed link constructed of three turns of number-10 wire inside the coil, to which was connected the 50-ohm coax. The same coil was used on both 3.5 and 7 MHz, with the capacitor tapped down on the coil for 7 MHz by means of an antenna switching relay.

Using a grid-dipper signal source and an antenna impedance bridge looking into the link, the points where the antenna feedline was tapped on to the coil were varied until, at resonance, the bridge read 50 ohms at both frequencies. This worked out to be 3 turns between the feeders. The vswr measured into the coaxial line at the transmitter is 1.05:1 at 3500 kHz, 1.6:1 at 3800 kHz, 1.5:1 at 7000 kHz, and 1.9:1 at 7250 kHz. In tuning, once the tuner is very near resonance, you simply watch the reflected power as the motor is energized, and set the variable capacitor for a minimum reading.

This antenna, at 100-feet high, has proved remarkably good for DX work in the preferred direction, which is toward Asia and South America. In the CQ DX contest, November, 1972, on 3.5 MHz 34 countries, 19 zones and 172 stations were worked. The antenna proved itself in the pile-ups for ZD3Z, 9L1GG and 5T5CJ, all worked through eastern-U.S. QRM on 3.5 MHz. Both VU and UL7 have been worked with this antenna on 7-MHz CW, via the long path around sundown.

The main objective of achieving a good match on both bands to the open-wire feedline without excessive switching has been realized. It is hoped that the approach presented here will be useful to others wanting two-band antennas, even if they don't happen to have 135-foot redwood trees to tie them to!



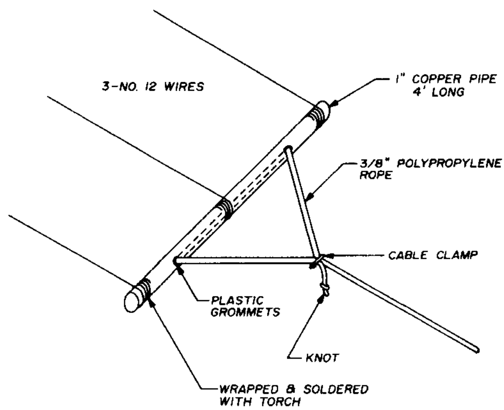


fig. 9. Construction details of the 3.5- and 7.0-MHz antenna used at W6MUR.

references

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2. Shintaro Uda and Yasuto Mushiake, *Yagi-Uda Antenna,* Sasaki Printing and Publishing Co., Ltd., 27 Tsutsumi-dori, Sendai, Japan, distributed by Maruzen Co., Ltd., 2-chome, Nithombashi-dori, Chuc-ku, Tokyo, Japan, 1954, pages 20-25.
3. John D. Kraus, *Antennas,* McGraw-Hill, New York, 1950, page 305.

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$$\text{*SINAD} = \frac{\text{Signal + noise + distortion}}{\text{Noise + distortion}}$$

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simple and accurate rf power meter

A useful device for measuring transmitter power output

Adrian Weiss, K8EEG, 213 Forest, Vermillion, South Dakota 57069

The increasing popularity of QRP operation has had one beneficial effect on amateur radio – an increased awareness of the importance of output power versus input power as a measure of performance. It has been generally accepted practice to state the performance of a given circuit in terms of input power capability, but we have long known that input power is no criteria for judging the important aspect of performance; i.e., what the circuit can deliver to the antenna. Our continued use of this archaic approach is perhaps due to the odd FCC practice of defining amateur service power limits in terms of input, while most other services are defined in terms of *effective radiated power.*

efficiency

Whatever the cause, a rather unscientific practice has occurred, namely the rule of thumb assumption that a given circuit will put out roughly 50% of the input power. But such an assumption is in defiance of all known principles of scientific methodology. One simply cannot *assume* that the efficiency of a given circuit is 50%. In fact, the theoretical ideal of a 60% efficiency factor may be achieved in practice in some circuits, but more likely the efficiency achieved will be considerably below that level. Measurements I have made over the years have revealed that some commercially produced "novice" kit transmitters hardly came out above 25% efficiency, and some were as low as 10%! These results have been corroborated by other experimenters. The lesson here, then, is that input power is no index of what really matters — power delivered to the antenna.

power output meter

Numerous devices for measuring power output have appeared in the literature. The circuit of **fig. 1** consists of two parts. First, a dummy load constructed of non-inductive, carbon composition resistors is designed for a certain level of power dissipation. Secondly, a sensitive metering circuit samples the peak rf voltage developed across the dummy load through CR1. That voltage is measured through a series dropping resistor on a 200-μA meter. The data is plotted to show μA versus watts. The formula of **fig. 1** is used to convert the voltage developed to average power. Accuracy of a high order can be achieved by close attention to accurate measurements of the calibrating voltage, the actual resistance of the dummy load, and the use of a high-quality microammeter.

Author Weiss' opening remarks are well put. Tuning across the ham bands, one often hears remarks such as, "The power here is X watts into an X antenna up X feet." If the power is stated as 1 kW, for example, the fellow on the receiving end might well raise his eyebrows. If the power is stated as dc input (which is often the case), the fellow making the statement is either (a) exceeding the legal power limit for amateur service, (b) his rig is 100% efficient, (c) he's expressing wishful thinking, or (d) he really doesn't understand the difference between input and output power.

True output power is difficult to measure in a radio transmitter, especially if high power is used. This article will permit the measurement of output power with reasonable accuracy for rigs of moderate power. **editor**

construction

The photos illustrate the final constructional approach used in this output meter, which is designed for about 60 watts continuous, and perhaps double that for intermittent service. During experimentation, it was discovered that with power levels above a few hundred

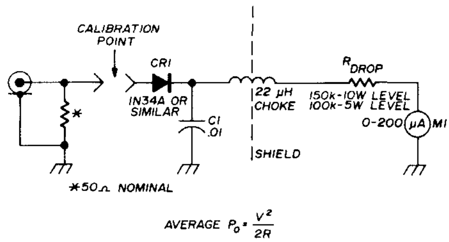


fig. 1. Output power meter schematic. Do not attempt to use a voltmeter at M1. Voltmeters exhibit an impedance which will distort the load seen by the transmitter at the input port at rf, resulting in inaccurate readings due to high swr. However, a high-impedance input meter such as a vtvm may be used instead of the meter shown.

milliwatts, a sufficiently intense rf field was generated, which upset the meter indication accuracy. Hence, a shield was inserted between the meter and the dummy load compartments. Next, it was found that the lead from the rectifying diode provided an rf path to the meter. A small, low-resistance 22 μH rf choke was installed in the lead and positioned halfway through the shield. This eliminated the last traces of rf feedthrough.

If the spirit of this approach is followed — shielding and filtering for rf — an accurate indicating meter will result.

dummy load

The dummy load should be designed for the power levels to be measured. In the meter shown, the dummy load is constructed from 21 two-watt, 1000-ohm resistors in parallel. This load will allow for about 60 watts continuous duty, provided that at least 1/8-inch clearance around each resistor is allowed, and ventilating holes are drilled above and below the load. For power levels under 4 watts, three 180-ohm, two-watt resistors may be paralleled. Whatever setup is devised for the dummy load, an accurate measurement of the resulting actual resistance is necessary to determine the quantity R

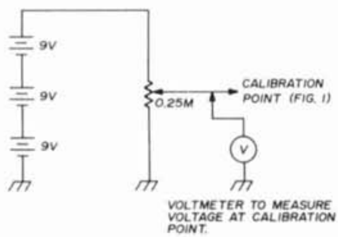


fig. 2. Suggested power supply.

shown in the formula. The dummy load resistors are sandwiched between two sheets of PC board. The lead from one resistor at the center of the board is left about 1/2-inch long, so that it may be inserted directly into the center conductor of the coax receptacle. The ground side of the load is connected to the receptacle outer shield by two no. 18 wires, which pass through both PC boards and go directly to the receptacle.

Even though the dummy load shown in the photo is approximately 3 x 2-5/8 inches, reactance is negligible. When fed through an odd length piece of coax, the swr is not above 1.05:1.

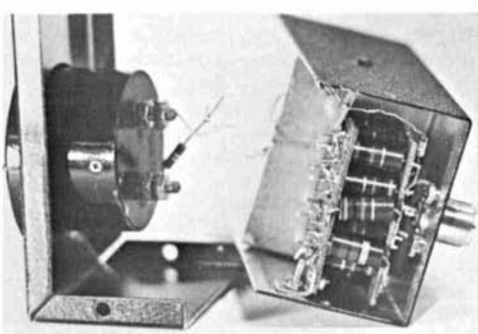
calibration

As noted above, meter accuracy depends upon care in calibration and using a high-quality meter. The best method of calibration is at dc using a variable-voltage power supply. If no such power supply is available, a suitable substitute is shown in **fig. 2,** and consists of four 9-volt transistor batteries in series to provide approximately 37.5 volts. This supply will calibrate up to about 14 watts. In calibrating, the voltage drop across CR1, the rf choke, and the dropping resistor is automatically accounted for by connecting the calibrating voltage source to the dummy-load side of CR1 – disconnect CR1 from the load first, of course!

Our objective in calibration is to plot the reading in microamps against the voltage at the calibration point **(fig. 1).** The series dropping resistor value can be determined by experimentation. With the meter used here, a 100k resistor provided just over full-scale meter deflection at 10 watts; a 150k resistor at 5 watts. **Table 1** lists the calibration points in the meter shown, and the formula used to calculate the voltage versus watts at chosen power levels. Once the complete number of power levels has been calibrated against microamp readings, the accuracy can be cross checked by attempting to reset the calibrating voltage to given levels by using the newly calibrated microammeter. Results will give you a practical idea of what to expect in the way of accuracy. With care, at least ±5% should be easily achieved. Remember we are ultimately dealing in terms of signal levels, where 1 dB is the lowest possible noticeable difference in signal strength. Accuracy of ±15% is acceptable in these terms.

other applications

The instrument is quite simple and is a valuable addition to any station. Its uses go beyond simple power output measure-



Inside view. Dummy load, coax receptacle and shield. CR1 is to the right of the dummy load PC board. C1 is partially obscured at the rear of the load. Meter and series dropping resistor are at left.

ment. It can be used to calibrate the popular Breune type in-line power output meter, such as was described by DeMaw.[1] Also, it can be used to measure the output of an exciter to determine exactly how much attenuation is needed to provide proper excitation to a linear. For the QRPP operator, such a device is absolutely essential. For the QRO gang, it can provide a much needed moment of truth

table 1. Calibration points calculated from formula of fig. 1 transposed to solve for V^2, i.e.: $V = \sqrt{P_o \times 2R}$ (102 ohms)

P_o (watts)	V (volts)	µA (my meter)
10	31.93	203
9	30.02	194
8	28.50	185
7	26.74	169
6	24.7	158
5	22.5	145
4	20.17	129
3	17.56	113
2	14.28	93
1	10.99	67
0.9	9.58	63
0.8	9.03	59
0.7	8.44	55
0.6	7.82	51
0.5	7.14	48
0.4	6.38	43
0.3	5.53	38
0.2	4.51	30
0.1	3.19	22
0.075	2.76	17
0.05	2.25	14
0.025	1.59	10
0.01	1.01	5

when connected to the pride and joy that's supposed to be putting out (assuming 50% efficiency, that is) 90 watts. The device requires such little time and effort to construct and calibrate, there is little reason why any station should be without it. It provides an indication of the only significant performance factor in the transmitting system – the amount of power the transmitter can deliver to the antenna.

reference

1. Doug MeMaw, "In-Line RF Power Metering," *QST*, December, 1969, page 11; "The QRP 80-40 CW Transmitter," *QST*, June, 1969, page 16.

ham radio

circuit improvements for the advanced frequency scaler

Several advanced circuit improvements for the vhf divide-by-ten frequency scaler previously described in *ham radio*

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

This is a follow-up to the divide-by-ten frequency scaler described in the September, 1972, issue of *ham radio.*[1] That article was originally prepared in late 1971 when the Fairchild 95H90 IC was relatively new, and experience with it, ham-wise, was quite new indeed. More than a year's experience with the 95H90 is now behind us, so a later look seems appropriate.

To the serious experimenter, the original article presented no real problems and many letters I received indicated that the device was a worthwhile project and performed as described. Others, being perhaps less knowledgeable, have had some difficulties. Drawing upon the experience of *Belmont Spectrum Research* in its commercial manufacture of scalers, and its experience wtih a great many ICs, as well as letters from users, indicates that attention to the following points should greatly assist in smoothing out difficulties and in making your scaler a truly useful device.

power supply

Experience has shown that *most* 95H90s have optimized at between 4.75 and 4.85 volts. This means that the power supply should provide this voltage to the IC. Accordingly, and this is done in the commercial version, a different power supply is now recommended for new construction (see **fig. 1**). This circuit is much more simple than the original and can easily be optimized. The circuit used Fairchild's 7805 voltage regulator IC (National's LM309 is equally suitable) which is rated at 5.0 volts output. These regulators may vary slightly from the "typical" 5.0-volt rating specified on the data sheets, so measure their output voltage. When using them, shoot for 4.85

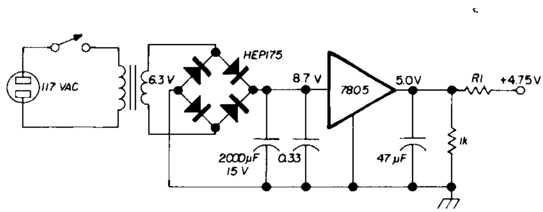


fig. 1. Recommended power supply for the 95H90 frequency-scaler IC. The supply voltage is connected to pins 4 and 5 of the 95H90. If the voltage is greater than 4.75 to 4.85 volts, install resistor R1 (2.7 to 3.9 ohms) and adjust value, as required.

volts at pins 4 and 5 of the 95H90 by inserting, if necessary, a 2.7- to 3.9-ohm resistor in series (R1 in **fig. 1**).

frequency limit

Experience with many manufactured scalers has shown that all 95H90s are far from being identical as far as their *upper* frequency limits are concerned. Some will go to 325 MHz, many will go to 300 MHz, but some will not go above 260 MHz. They are specified to 220 MHz as a minimum and to 270 MHz as "typical" by Fairchild.

The reason for the wide upper frequency range found in practice is that this IC is heat sensitive. That is, a large temperature rise lowers the upper frequency limit. Hence, the 95H90 should have a good heat sink if its maximum possibilities are to be exploited. The grounding of all unused pins (except pin 14, more on this later) will help. A good commercially made heat sink is made by IERC (their part numbers DC000080B and LIC 214A2WCB). You can make your own heat sink, however, by using a two-inch square piece of aluminum in contact with the top of the IC and bolted to the circuit board.

circuit-board design

A good ground plane is essential in the frequency scaler. This is why, in the original article, lines were removed from a "ground plane" copper board rather than using interconnecting traces. Don't forget you are dealing with very high frequencies where miscellaneous circuit paths may lead to feedback or ground loops and cause instability. Those builders who have used "traces" have had real problems and have ended up remaking their circuit boards.

preamp

If a preamp is used, and one is very worthwhile (the commercial version uses two stages), the 95H90 must be adequately decoupled from the preamp. This is rather simply accomplished by the 22-ohm resistor between the 5-volt supply and the junction of the 180-ohm resistor and the peaking coil shown in **fig. 5** of the original article. Do not omit this precaution against feedback or you will end up with an oscillator, not a preamplifier.

low-impedance input

Experience has shown that *most* 95H90s respond best to a low-impedance input. This was not so with the prototype ICs I used for the original article. Thus, it is now recommended that a 68- and

200-ohm bias divider be used for the 95H90. Also, if you would like a further refinement (the original was somewhat of a compromise in order to make it work for the vast majority of cases) the use of a *threshold* control, as shown in **fig. 2**, may be more acceptable for your applications. Try it, you may like it.

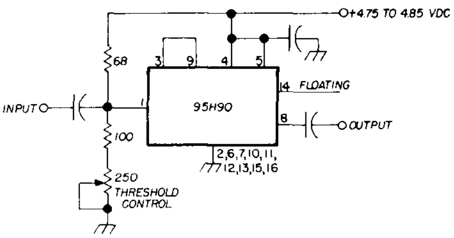


fig. 2. Improved low-impedance input circuit for the 95H90 IC. The 250-ohm threshold control is optional but should prove useful for most applications (see text).

bypassing

Bypassing pins 4 and 5 of the 95H90 (assuming you followed **fig. 3B** of the original article) is very important. The by-pass capacitor should be connected as close as possible to pins 4 and 5. In other words, *zero* lead length!

ground unused pins

Experience has shown that *all* unused pins of the 95H90 with the exception of pin 14, should be grounded. Thus, if you use a negative grounded board as in **fig. 3B** of the original article and use pin 1 as the input, pins 2 and 16 should be grounded. Pin 14 should be left floating (ungrounded). The first few ICs I tried did not require this. However, subsequent experience has proved the efficacy of the grounding of pins 2 and 16. Ground them!

heat dissipation

The 95H90, which dissipates approximately one-half watt continuously, should feel only slightly warm to the touch if operating properly. If it runs hot you have something wrong. Check the voltage at pins 4 and 5. It should not be over 5.0 volts and preferably, only 4.75 to 4.85 volts. Also, check your output resistance (with the unit turned *off*). It should not be much less than 800 ohms at pin 8. If it is less, you undoubtedly have a circuit error.

base diagrams

Be sure you have not somehow confused top and bottom views of the 95H90 and the 2N5179. A number of builders have interchanged them, much to their sorrow. *Always* check a data sheet which shows the basing diagram. In fact, double check it.

TTL interface

Unless you are sure that you need a TTL interface (as shown in **fig. 6** of the original article) omit it. There are presently no known counters which require it. If you do use it, however, pay no attention to the V_{cc} and V_{ee} markings on **fig. 6.** The proper voltage is already taken care of in the 95H90.

cable termination

In using a scaler with a highly sensitive high-impedance frequency counter, false counting can be experienced if the interconnecting cable between the scaler and the counter is not properly terminated at the counter end of the line. A line termination consisting of a carbon resistor, equal to the characteristic impedance of the line, should be connected across the line at the counter end. This precaution is often overlooked by hams but is standard practice in industry. Tektronix makes a fine termination adaptor for use with 50-ohm coax (their part no. 001-0049-01).

reference

1. F. Everett Emerson, W6PBC, "Advanced Divide-by-Ten Frequency Scaler," *ham radio*, September, 1972, page 41.

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improved half-wave rectifier

Circuit details of an improved half-wave, choke-input power supply circuit that offers several advantages over conventional circuits

John T. Bailey, 86 Great Hills Road, Short Hills, New Jersey 07078

Suppose you needed a simple power supply in a hurry that would give you 50 volts dc at, say, 100 mA, well filtered, with load regulation at least equal to that of a choke-input filter, and without excessive peak rectifier currents. You might say, "That's not difficult." A conventional full-wave center-tap circuit feeding a choke-input filter would meet these requirements and the parts should be available in the junk box.

You start looking for a transformer with a 120-volt center-tapped secondary in order to get 50 volts dc at the input to the filter choke. Or, if you used a bridge rectifier, you would need a 60-volt untapped secondary. My guess is that you won't find either transformer in your junk box.

half-wave circuit

That being a dead end, let's make it more difficult. How would you design for the same requirements without using any transformer? I'll give you a hint. Try a half-wave rectifier feeding a choke-input filter. But, you say, "This is never done. The regulation would be horrible and the output voltage would be practically unpredictable."

One simple diode added in the right place can fix all that. Simply wire a diode, reversed polarity connection, from the choke input to the common line. That's all. The load regulation will be

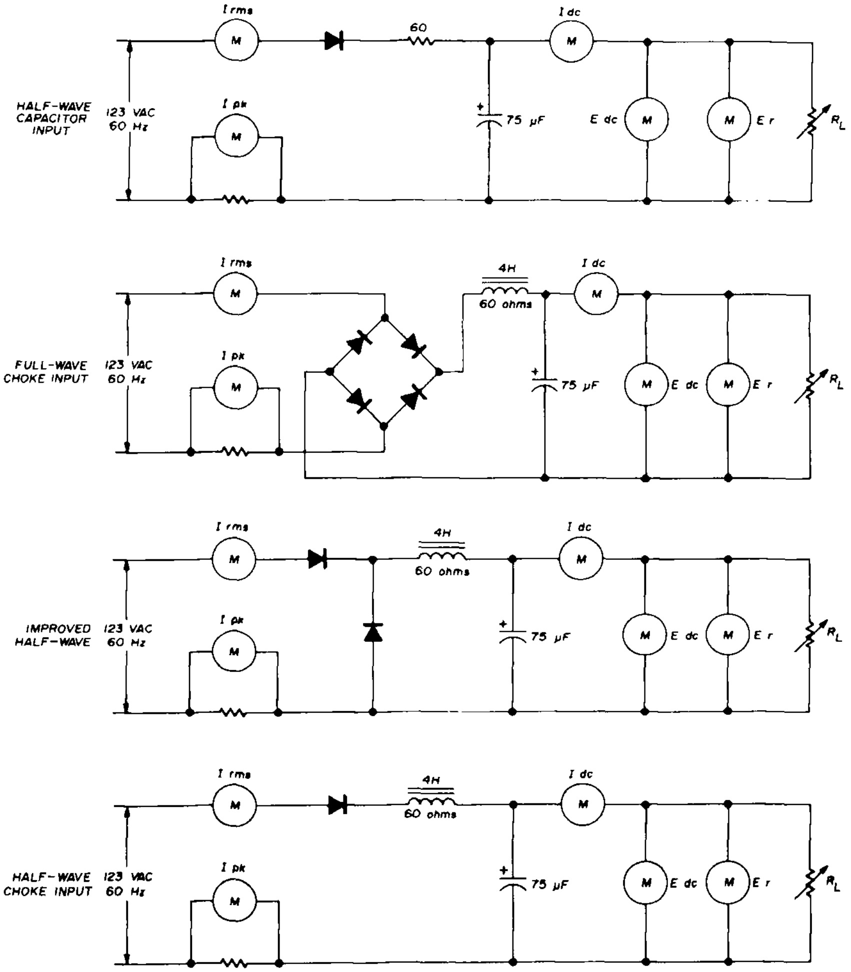

fig. 1. Power-supply circuits measured to show performance of improved half-wave circuit relative to other popular rectifier circuits.

equal to that of a full-wave, choke-input circuit. The output voltage will be one-half that of a full-wave, choke-input circuit or 45% of the rms ac input.

What about disadvantages? There must be some. A minor one is that when the dc output current must be varied from full load to a much lower output current the minimum load current must be greater than that for a full-wave choke-input circuit to prevent the circuit from reaching criticality.

Ripple in the output voltage will be the same frequency as the supply. Its magnitude will be more than for the full-wave, choke-input circuit but much less than for the conventional half-wave, capacitor-input circuit. The ripple voltage waveform will resemble the "full sine wave" ripple of the full-wave, choke-input circuit rather than the "triangular" waveform of the conventional half-wave circuit.

tests

To delineate the performance of the improved half-wave circuit, measurements were made on the four circuits shown in **fig. 1**. *The results are plotted in* **fig. 2**. **Table 1** lists other data applicable to the four circuits using the component values shown for each. If some of these data appear to differ from those derived from theory remember that the data here are measured values on circuits whose components are not the *ideal* components frequently assumed in textbook analyses.

In these circuits the I_{rms} line currents were measured with a thermocouple milliammeter. The peak line currents were measured with an oscilloscope across a 1-ohm resistor in the line. The ripple waveforms were observed on a scope across the load resistor and the rms values were calculated.

table 1. Data applicable to the four rectifier circuits shown in fig. 1.

	half-wave capacitor input	full-wave choke-input	improved half-wave	half-wave choke-input
Ripple voltage, rms				
at full load	9.6	.33	1.2	1.9
at 50% full load	5.5	.31	1.1	1.2
at 20% full load	2.7	.30	1.0	.7
Internal resistance, ohms				
at full load	250	66	66	1,100
at 50% full load	300	66	66	1,700
at 20% full load	500	66	66	3,400
I_{rms}/I_{dc}				
at full load	2.0	1.00	.8	1.3
at 50% full load	2.3	1.05	.8	1.5
at 20% full load	2.6	1.10	1.0	2.0
I_{pk}/I_{dc}				
at full load	5.0	1.10	1.3	2.2
at 50% full load	6.3	1.23	1.5	2.8
at 20% full load	9.2	1.67	2.3	3.8

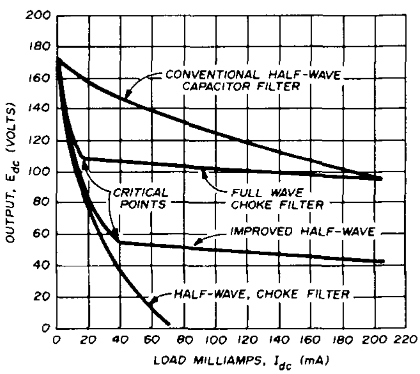

fig. 2. Voltage regulation curves of the rectifier circuits shown in fig. 1. Note the superior regulation of the improved half-wave circuit ***relative to the other half-wave circuits.***

Note particularly that the slopes of the curves for the improved half-wave circuit and the full-wave, choke-input circuit are straight and parallel over their useful spans indicating the internal resistance of each is constant and equal to the other.

summary

Of course, this improved half-wave circuit is not limited to line operation without a transformer. It is a handy circuit to keep in mind whenever an available secondary transformer voltage is about twice as high as you need to get a certain dc voltage when using conventional circuitry. For instance, a 24-volt filament transformer feeding this circuit will give you a handy 10 volts for those transistor projects where better regulation isn't justified.

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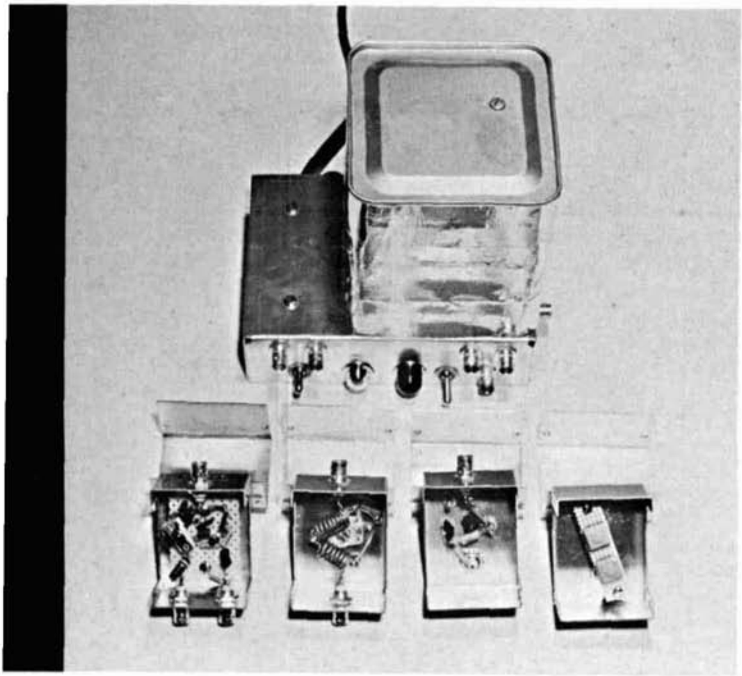


photos by WA3MEJ

accurate frequency measurement of received signals

How to use an offset frequency measurement system that does not impair the performance of your receiver

Jeff Walker, W4AAD, Pasadena, Maryland 21122

Several articles dealing with frequency measurement of received signals have appeared in the amateur publications.[1-6] The techniques described have been essentially of four types, manual analog comparisons, *receiver dial readouts, received* signal synthesis and multiplexed *up/down counting.* Each of these techniques has advantages and disadvantages.

Manual analog comparisons (zero beating and interpolation) are easy and inexpensive to implement but require considerable operator skill and are subject to human error at best. Receiver dial readouts are fine except that they read the frequency to which the receiver is tuned and not that of the received signal. Also, most dial readout systems described to date have a resolution of only 100 Hertz.

Received signal synthesis is good except that the synthesized signal tends to re-enter the receiver front end and cause oscillation problems. The multiplexed up/down counter approach works well but requires some moderately complex digital design and pretty well eliminates the use of a general-purpose frequency counter.

My goals when designing this measurement system were to obtain precise readout of the received signal carrier frequency (within counter resolution) with complete independence from receiver tuning. The system would offer easy calibration and require no operator-performed analog comparisons, such as zero beating. It would provide stable, non-oscillatory operation and use an existing general-purpose frequency counter. Also, it would have high sensitivity and be capable of operating on all the hf and vhf amateur bands within the limitations of the counter.

To meet all the requirements, I decided to use a modified form of received signal synthesis. Henceforth, I will refer to this technique as *offset counting.* Signal synthesis satisfies most of the design goals, however, it does not satisfy the requirement for stable, non-oscillatory operation. This problem is worsened when a general-purpose counter is used because the synthesized signal must be fed into the counter with coaxial cable, thus allowing some signal leakage. Furthermore, most counters generate at least 1 volt, peak-to-peak, of counted signal and radiate some portion of that. The net result is anything but a stable, non-oscillatory system. The only way to make such a system work is to desensitize it. This compromises the goal of high sensitivity.

However, if the counted signal is on some frequency other than the received frequency, high sensitivity is not compro-

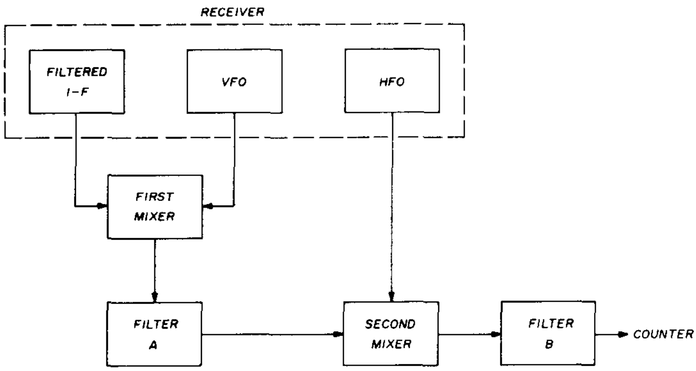


fig. 1. Block diagram of a system to synthesize a received signal for frequency counting.

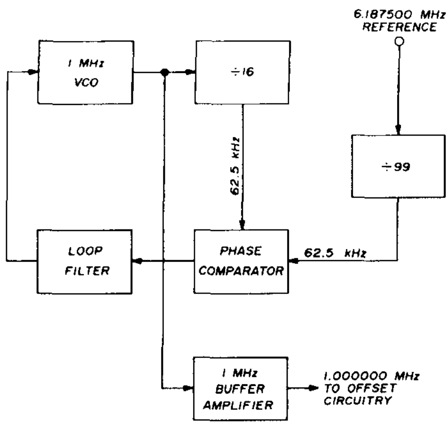


fig. 2. Block diagram of a circuit to phase lock an offset oscillator to an "odd-ball" reference frequency.

mised, and the system provides stable, non-oscillatory operation. The problem here is that you must know *exactly* how the counted signal relates to the received frequency. I accomplish this in the offset counting system by introducing an offset which is derived from the time base used to control the counter gate.

This system was initially designed and used for the high-frequency bands. With the addition of just one mixer and a filter, the technique is expandable to vhf.

theory of operation

To generate a synthesized received signal for counting, the system shown in **fig. 1** would be used. Such a system, typically, would be prone to oscillation. I might point out that this system can be made to work by enclosing the mixers, filters and counter section in the same small, well shielded and filtered enclosure. For this to work, after the signal has been synthesized, absolutely no rf can be allowed to escape from the enclosure.

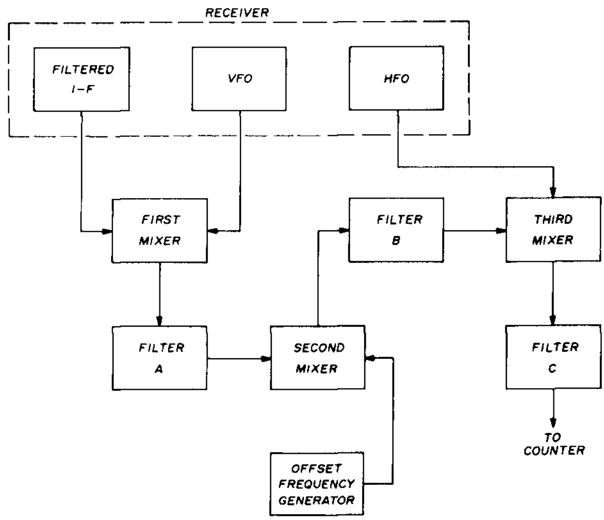


fig. 3. Simplified block diagram showing how an offset counting system may be used to measure the frequency of a received signal.

This is not an easy task to accomplish; shielding and filtering well in excess of usual amateur practice would be required to build a truly stable, high sensitivity system. In **fig. 1,** the first mixer combines the filtered i-f output with the variable-frequency oscillator to produce sum, difference and spurious products.

Depending on the mixing scheme of the receiver, filter A is designed to pass only the desired product and attenuate all others. The output of filter A is then fed to the second mixer where it is combined with the output of the high-frequency oscillator. Either the sum or difference component is selected by filter B (depending on the receiver mixing scheme) and fed to the counter. This diagram is somewhat simplified in that buffer amplifiers may be required between the receiver oscillators and external mixers as well as between the first and second external mixers.

For example, when designing a synthesis system around a Collins 75S-3 receiver, the hfo operates 3.155 MHz above the low end of the band in use, the first i-f is a passband from 2.955 to 3.155 MHz, the vfo tunes from 2.7 to 2.5 MHz, and the second i-f is 455 kHz. The first mixer uses difference mixing and the second mixer uses summation mixing.

Thus, in **fig. 1,** the first mixer sums 455 kHz and 2.7-2.5 MHz. The desired output of this mixer is 3.155-2.955 MHz (3.155 MHz is at the low end of the band being tuned). This range is selected by filter A which may be either a bandpass or highpass type.

The second mixer takes the difference between 3.155-2.955 MHz and (F_1 + 3.155 MHz) (F_1 is the frequency of the low end of the band being used, for the 7.0-7.2-MHz band [F_1 + 3.155 MHz] is 10.155 MHz). The difference output of the second mixer is then exactly the same frequency as the received signal. Filter B selects this product.

To modify the synthesis system to offset counting, an offset frequency must be inserted which will displace the counted frequency sufficiently from the received frequency that the receiver will not respond to it. This offset can be any value large enough to get the counted frequency out of the i-f passband, or preferably, out of the i-f *and* rf passbands. The two absolute conditions which must be placed on the offset frequency are that it be stable (preferably locked to the counter time base) and that it be known to an accuracy at least as good as the highest resolution expected from the counter.

For example, if some multiple of the counter time base is 6.1875 MHz, this would be a perfectly acceptable offset frequency. However, it would be inconvenient to have to add or subtract this number from every frequency you measure. A much more convenient number would be 1.000 or 5.000 MHz. If a nice, round offset frequency is not directly available, it may be possible to phase lock a lower stability signal.

In the case of the 6.1875-MHz signal mentioned above, it is possible to use the circuit block diagrammed in **fig. 2** to phase lock a 1-MHz oscillator to the 6.1875-MHz signal. The phase locked 1.000-MHz signal may then be used for the offset frequency. The reference frequency may be any value that has a common denominator with the offset frequency with which you desire to phase lock.[7]

To see how the offset frequency would be added to the basic synthesis system, refer to **fig. 3.** The offset counting system requires one additional mixer, another filter and an offset frequency source. The signal flow is the same as in **fig. 1** up to the second mixer. At this point the offset frequency is combined with i-f/vfo sum. Either the sum or difference product of the second mixer may be selected by filter B.

In general, the sum product would be preferred as this will ease the design requirements on filter C (in the case of the 75S-3 mixing scheme, it also allows 10-MHz counters to operate up through 20 and 15 meters). The output of filter B is combined with the receiver hfo in the third mixer. Either the sum or difference

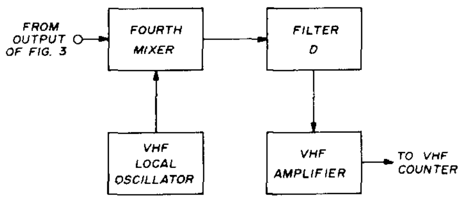


fig. 4. Offset frequency measuring system may be extended to the vhf bands with the setup shown here.

may be selected by filter C; the difference product would be preferred as it will result in a lower frequency to be counted and, thus, be within the range of more general-purpose counters.

Note that I did not combine the receiver i-f, vfo and hfo in sequence; to have done so would have resulted in the generation of the undesired received signal. Instead, the offset was entered before the hfo signal.

The offset counting system may be extended to include the vhf bands by adding an additional mixer, filter and source of vhf local-oscillator signal. Refer to the block diagram in **fig. 4** to see how this is accomplished. The output from filter C in **fig. 3** is fed to the fourth mixer along with a portion of the vhf converter local oscillator signal. Depending on whether the local oscillator is above or below the received vhf signal, filter D should be designed to select the difference or sum output, respectively, of the fourth mixer.

It is likely that the output from filter D will be inadequate to drive your vhf counter so a stage of amplification will probably be required to raise the signal level. Don't forget that the offset frequency is still in the system and the vhf counter reading will include this factor.

At this point the advantage of using 5.000 MHz or 1.000 MHz for the offset frequency should be realized. Not only will the counter correction be easy to manipulate mentally, filtering out the image from the fourth mixer will be much easier if it is 10 or 20 MHz from the desired signal, rather than some considerably smaller value.

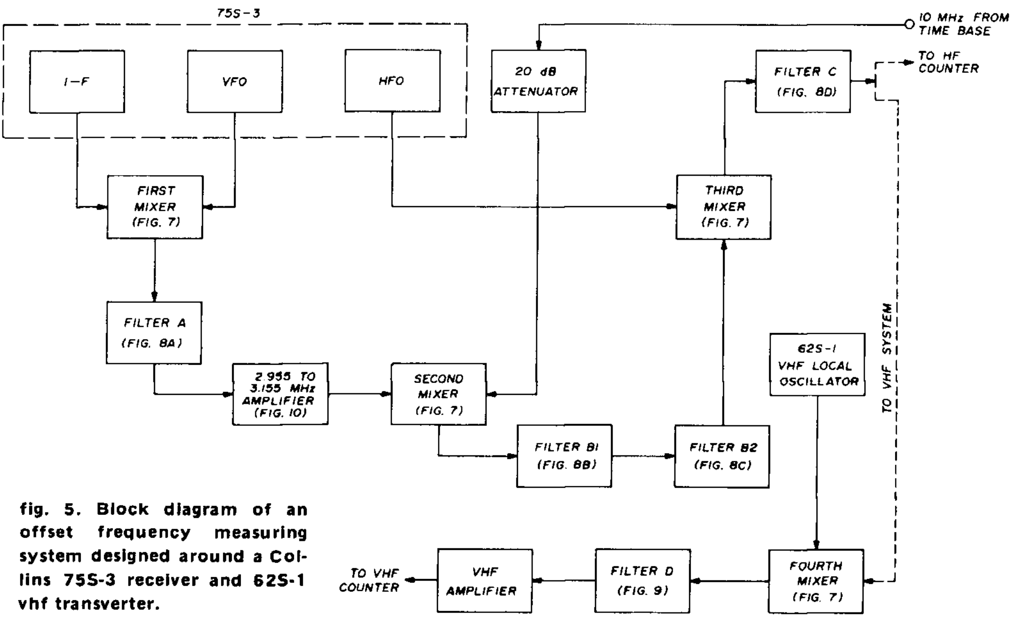


fig. 5. Block diagram of an offset frequency measuring system designed around a Collins 75S-3 receiver and 62S-1 vhf transverter.

construction

As can be seen in the photographs, I used a modular format in building this system. Each block of **figs. 3** and **4** is a separate module with the exception of the offset frequency source which is a part of my counter time base. This modular approach allowed me to mount the entire frequency measuring system in a convenient, long, narrow space on my operating table, behind the transmitter and receiver.

This is desirable since the cables feeding the oscillator signals from the receiver and vhf converter can be kept to a minimum length. However, there is no reason why the system can't be built into one large enclosure. If your operating table will permit the inclusion of an additional moderately sized unit you may wish to follow this approach. Just be sure to provide good shielding between the various mixers; otherwise, undesired signal leakage around the mixers will occur.

The modular approach also allows the builder to tackle construction of one distinct portion of the system at a time. Each module can be built, tested and set aside for later inclusion in the system. A good compromise between the modular and single unit approaches would be to build the separate modules and then assemble them in a mainframe or chassis; that way adequate inter-module shielding will be assured.

system description

My frequency measuring system was designed around a Collins 75S-3 receiver,

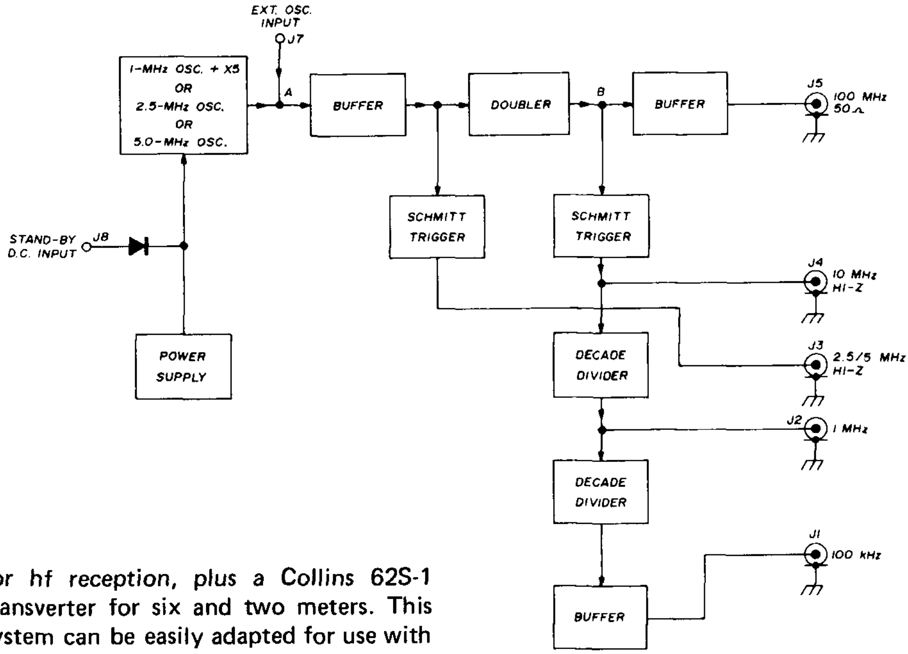


fig. 6. Block diagram of the frequency-counter time base and offset frequency source.

for hf reception, plus a Collins 62S-1 transverter for six and two meters. This system can be easily adapted for use with most hf communications receivers, irrespective of mixing schemes or number of conversions. Single-conversion receivers will require one less mixer, and triple conversion units, one more.

Assuming the modular construction approach is to be used, it is obvious that there is no provision for bandswitching, at least in the classic sense. To change bands requires changing one or more of the filter modules. This is no handicap unless you are an ardent band-hopper or want to measure the exact frequency of every station you work during a DX contest! Even then, you are in luck, providing you hop around the right bands; more on this later.

Complete block diagrams of the frequency measuring system implemented around the 75S-3/62S-1 combination are shown in **figs. 5** and **6**. In **fig. 5**, the 75S-3 i-f and vfo signals are combined in the first mixer. The desired output from this mixer is 2.955-3.155 MHz. This is selected by filter A which is a nine-section highpass filter with a cutoff frequency of 2.9 MHz. The output of this filter is fed to a tuned rf amplifier providing approximately 25-dB gain over the

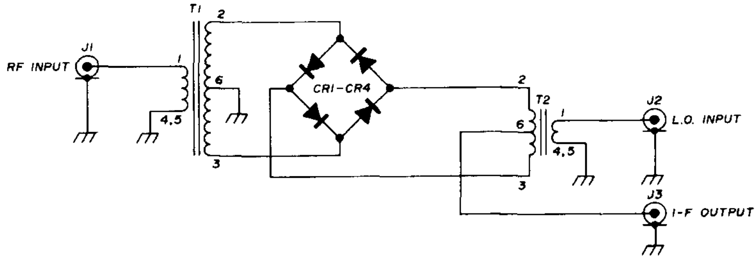


fig. 7. Double-balanced mixer circuit for use in the offset frequency measuring system. Transformers T1 and T2 are Vari-L wideband transformers, model HYB-1. Diodes CR1-CR4 are matched hot-carrier diodes, Hewlett Packard 5082-2805.

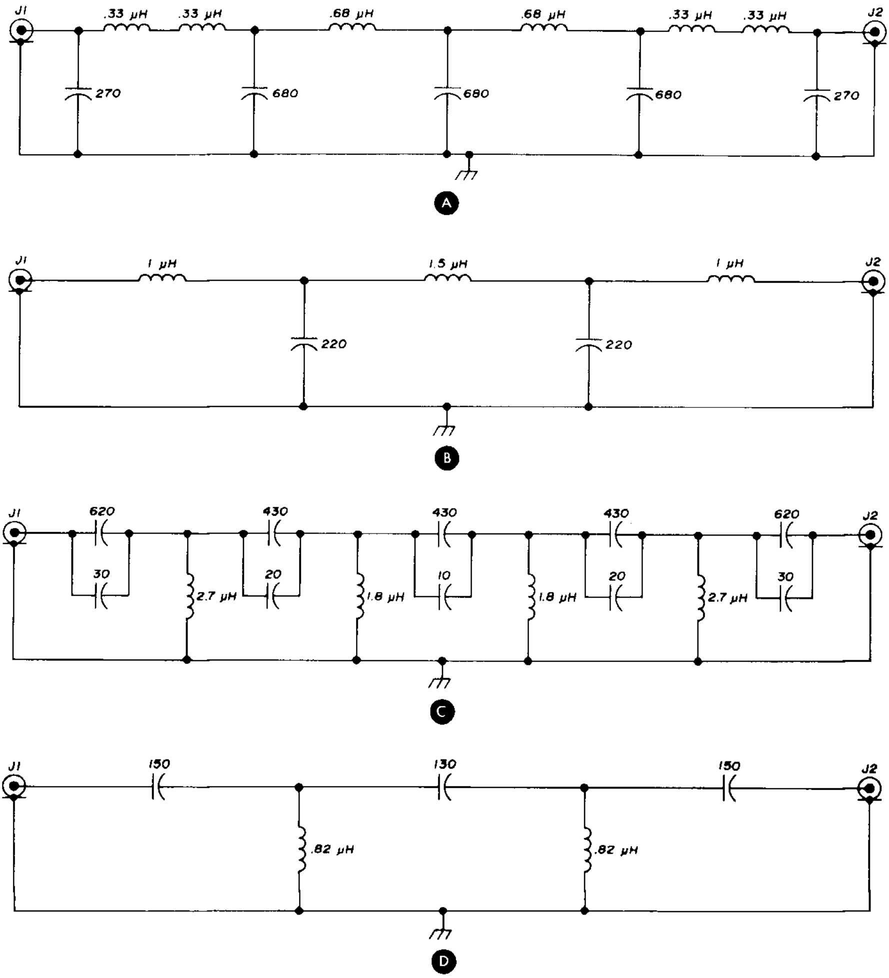


fig. 8. Filters for the offset frequency-measuring system. The filter in (A) is a highpass unit with a 2.9-MHz cutoff (filter A). The highpass filter in (B) has a cutoff frequency of 10 MHz (filter B1). The filter circuit in (C) is a lowpass unit with a 14.2-MHz cutoff frequency (filter B2). The lowpass filter in (C) has a cutoff frequency of 15 MHz (filter C). All capacitors are dipped mica; inductors are Nytronics Wee-ductors.

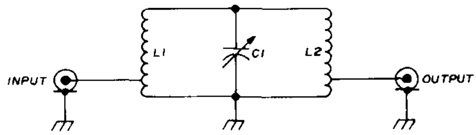
2.955-3.155-MHz range. The amplifier output is combined in the second mixer with the attenuated 10-MHz output from the time base – the offset frequency.

I elected to take the sum product of the second mixer (12.955-13.155 MHz) for further processing. This frequency range is filtered through filters B1 and B2 in tandem. Filter B1 is a 10-MHz highpass filter and B2 is a 14.2-MHz lowpass design. These two filters may be combined into a single bandpass unit. My junk box dictated the construction of the two filters in tandem. The output of filter B2 is combined in the third mixer with the 75S-3 hfo signal. The difference product is selected by filter C which is a 15-MHz lowpass design.

As indicated, the output of filter C may be fed directly to a counter for hf measurements or on to the vhf portion of the system. Assuming vhf operation, the offset hf signal is fed to the fourth mixer where it is combined with the 62S-1 local-oscillator signal (36-40 MHz for six meters or 130-134 MHz for two meters). Since the 62S-1 local oscillator is on the low side of the received signal, the sum product of the fourth mixer is selected by filter D and subsequently amplified in the vhf amplifier.

Fig. 6 is a block diagram of the 100-kHz counter time base and offset frequency source. It is not anticipated that many people will be interested in duplicating this unit exactly because it includes a surplus 2.5-MHz oven-controlled oscillator. The method and remainder of the circuitry are applicable, however, no matter what oscillator you use.

If you have a good 1-MHz oscillator, you could add a times-five multiplier and pick up **fig. 6** at point A. If you have a 2.5-MHz oscillator, use the entire block diagram. If you have a 5-MHz oscillator, again enter the diagram at point A. If you have a 10-MHz oscillator, enter the diagram at point B and delete the high-impedance 5-MHz output. If you don't have an existing oscillator available, I recommend a 5-MHz version of the oscillator described by Irv Hoff.[8]



40 - 44 MHz

L1,L2	13 turns no. 16, 5/8" diameter, 1-3/8" long, tapped at 2 turns from ground end
C1	9-180 pF mica trimmer (Elmenco 463)

134 - 138 MHz

L1,L2	7 turns no. 16, 3/8" diameter, 3/4" long, tapped 1-1/4 turns from ground end
C1	7-45 pF mica trimmer (CRL 822BN)

210 - 215 MHz

L1,L2	5 turns no. 16, 3/8" diameter, 1/2" long, tapped 1 turn from ground end
C1	4.5-25 pF mica trimmer (CRL 822AZ)

fig. 9. Schematic of the bandpass filter required for vhf offset-frequency measurements (filter D).

bandswitching

Earlier, I mentioned that bandswitching essentially consisted of changing filters. With the proper selection of filters, more than one band can be covered with the same units. Refer to **table 1** for a list of offset frequency ranges for each band when using the 75S-3 receiver and a 10-MHz offset frequency source. It will be recalled that filter C selects the difference product of the third mixer; this is the offset range column in **table 1.**

If the sum product were selected instead, the lowest frequency of interest would be 19.510 MHz (80 meters), filter C would have to be a highpass type and the hf counter would have to have a much greater operating range. Since 19.510 MHz is the lowest sum product which can cause trouble, filter C should have a cutoff frequency slightly below this value. It can be seen from the offset range column of **table 1** (or counted frequency) that if a 15-MHz cutoff lowpass filter is selected for filter C, then all hf bands except 10 meters can be measured with no filter changes.

To measure 10 meters, a 25-MHz cutoff lowpass filter should be substituted at filter C. If you often operate on (and want to measure) 10 meters, you may place both filters in the circuit with a coax

table 1. Offset frequency ranges for each amateur band when using a Collins 75S-3 receiver and a 10-MHz offset frequency source (all frequencies in MHz).

band	hfo	vhf LO	offset range
3.5 - 4.0	6.555 - 6.955	—	6.0 - 6.5
7.0 - 7.3	10.155 - 10.355	—	2.7 - 3.0
14.0 - 14.4	17.155 - 17.355	—	4.0 - 4.4
21.0 - 21.6	24.155 - 24.555	—	11.0 - 11.6
28.0 - 29.7	31.155 - 32.755	—	18.0 - 19.7
50.0 - 54.0	17.155	36 - 40	40.0 - 44.0
144-148	17.155	130 - 134	134.0 - 138.0
220-225	17.155	206 - 211	210.0 - 215.0

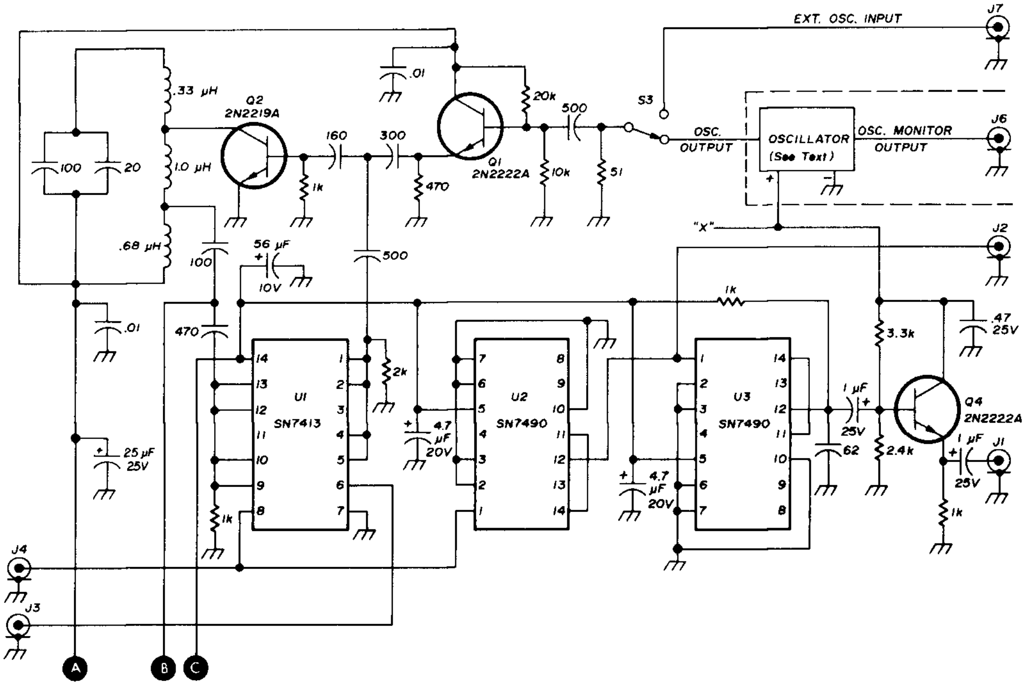


fig. 11. Time-base module used in the offset frequency-measuring system. Transformer T1 is a Stancor P-8180.

switch to short out the 15-MHz filter when operating on 10 meters.

It should be noted that if a 7-, 14- or 21-MHz i-f range is chosen for your vhf converter, then filter C may remain unchanged from hf band operation. Just drive the fourth mixer directly with the output of filter C. Filter D, of course, is selected for the vhf band to be used according to the offset range column of **table 1.**

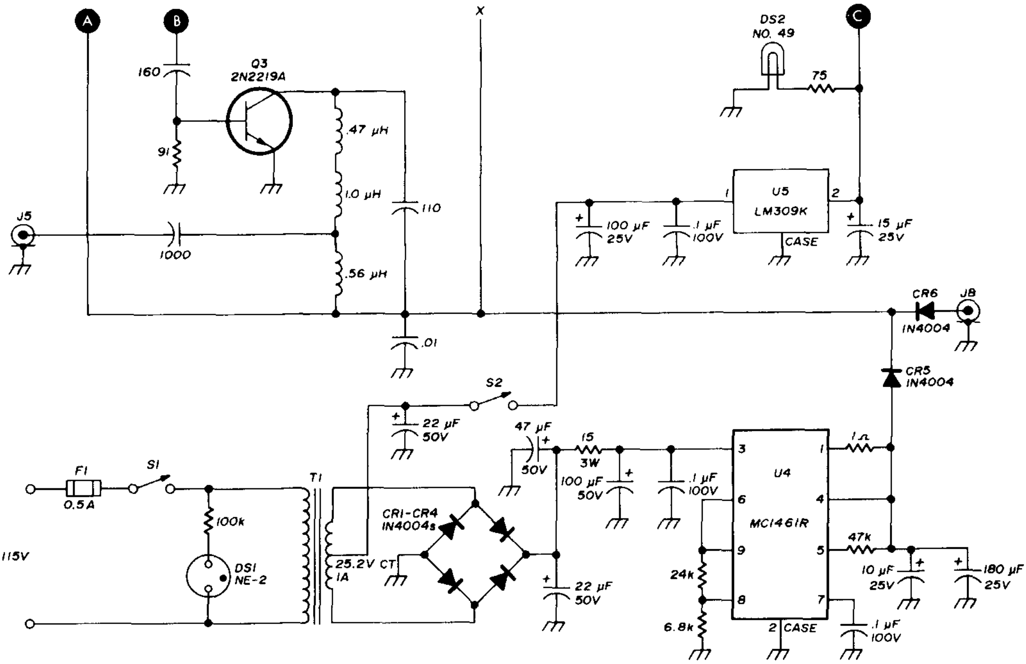
circuit description

The module circuits will be described in the following order; mixers, filter A, filter B1, filter B2, filter C, filter D, 2.955-3.155 MHz amplifier and power supply, vhf amplifier and time base.

All mixers are passive double-balanced mixers using hot-carrier diodes. Double-balanced mixers were chosen because they provide excellent suppression of mixer input signals at the output port. Passive mixers were chosen over active circuits because of their superior performance at vhf. Also, it is easier to provide the higher LO injection levels required by passive mixers than to put together the additional circuitry required by active devices (power supply, input/output coupling, etc.).

Several approaches may be taken to acquire the four mixers required for the system. Double-balanced mixers enclosed in shielded containers with female BNC connectors are commercially available from sources such as Hewlett-Packard, Relcom and Vari-L; prices start at around $40. Mixers with similar specifications but designed for PC board mounting are available from Vari-L, Mini-Circuits Laboratory, Merrimack Industries and others. Prices start at under $10 for quantities greater than five.

If you prefer to build you own mixers, one method has been described by Ress.[9] This unit will cost about $10 plus the PC board. If you can fabricate your own board, this is the most economical approach. I also built a double-balanced mixer using commercial wideband transformers and hotcarrier diodes mounted



on perf-board with adhesive copper foil used as the ground plane. The schematic for this unit is shown in **fig. 7.**

Filter A selects the 2.955-3.155 MHz output of the first mixer. This filter must have a sharp cutoff characteristic to attenuate the undesired 2.045-2.245 MHz image. I decided to use a nine section Chebyshev highpass filter with 1-dB passband ripple **(fig. 8A)**. This type filter has a sharper cutoff characteristic, for a given number of sections, than either the Butterworth or older image parameter designs. It provides a minimum of 16-dB rejection to the image. If your receiver has different i-f or vfo frequencies, this and subsequent filters may be designed using references 10 and 11.

Filter B1 is a 10-MHz highpass design. It is similar to filter A except it has only five sections **(fig. 8B)**. Filter B2 is a nine-section Chebyshev lowpass design with a 14.2-MHz cutoff frequency **(fig. 8C)**. Filter C is a five-section Chebyshev low-pass design with a cutoff frequency of 15 MHz **(fig. 8D)**. Filter D is a lumped-constant bandpass design which yields 3-dB bandwidths in the vicinity of 5% of F_o and ultimate rejection of at least 40 dB **(fig. 9)**.[12]

wideband amplifier

The 2.955-3.155 MHz amplifier is built into a small Minibox (Bud CU-2100-A). The power supply was built in a separate enclosure for more convenient placement. The amplifier is a single-stage common-emitter design **(fig. 10)**, only the collector circuit is tuned. The output is tapped from the collector tank capacitor rather than the inductor. This was done so I could use a miniature molded inductor in the circuit. Gain is approximately 25 dB.

The combination of filter D and the vhf amplifier, in **fig. 5**, provides approximately 20-dB gain at the selected vhf offset frequency. The most versatile way to achieve this is to follow filter D with a wideband 20- to 25-dB gain amplifier having a high-frequency cutoff at least as high as the maximum frequency to be counted. An alternative is to replace filter D and the wideband amplifier with a multi-stage tuned vhf amplifier. I have used both approaches successfully.

A disadvantage of using the alternative is that every time you go to a different position of a vhf band, you must retune the amplifier or replace it with another pretuned amplifier. If your interest lies in only one portion of one vhf band, then this approach is the most economical. As a matter of fact, if you have a preamplifier for the band(s) of interest, and it will put out 1 milliwatt before gain compression, you can use this in place of filter D and the vhf amplifier.

The preamp must be capable of being tuned to the offset frequency 10 MHz below the band being used. This is most likely possible on 144 and 220 MHz. If you use a broadband preamplifier (10 - 20 MHz) be sure to use filter D to prevent the image from the fourth mixer from getting through.

If you prefer to use the filter and wideband amplifier, commercial models are available from Hewlett-Packard, C-Cor, Avantek, Optimax and others. Prices of modular wideband amplifiers suitable for PC board mounting begin around $40. If you are interested in building your own wideband amplifier, see references 13 and 14 for ideas. Also, the International BAX-1 amplifier module should be useful in this application.

time-base module

The time base is the largest module in the system. Indeed, it could be considered a sub-system (**fig. 11**). The schematic does not include an oscillator, which may be considered as a separate module. A recommended circuit is given in **fig. 12**.

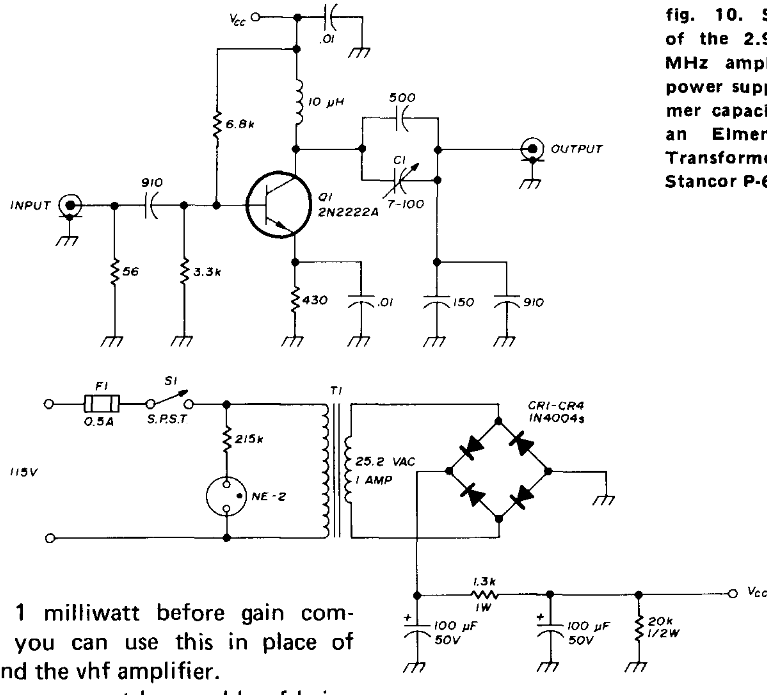


fig. 10. Schematic of the 2.955-3.155 MHz amplifier and power supply. Trimmer capacitor C1 is an Elmenco 423. Transformer T1 is a Stancor P-6469.

Referring to **fig. 11**, the input stage, Q1, is an emitter follower which establishes a resistive 50-ohm input impedance. This stage drives an SN7413N Schmitt trigger which will give a 2.5 or 5.0-MHz, TTL compatible, output (J3) depending on the oscillator you use. Transistor Q1 also drives a class-C doubler/amplifier stage (Q2) which operates as a doubler when driven at 2.5 MHz and as an amplifier when driven at 5 MHz.

The doubler/amplifier drives the second half of U1 (SN7413N), a Schmitt trigger, which has a TTL compatible 10-MHz output (J4). Transistor Q2 also drives Q3 which is a class-C, 10-MHz output amplifier; Q3 provides +25 dBm (315 mW) output at 10 MHz (J5). A portion of this signal is fed to the external second mixer for the offset frequency.

The 10-MHz output of U1 drives U2, an SN7490N decade divider, which provides a TTL compatible 1-MHz output (J2) and drives U3, an SN7490N decade divider, which drives Q4, an emitter follower. Q4 provides a 100-kHz output (J1) for the counter external time base input.

The power supply uses a center-tapped bridge circuit for dual output voltages. The high output (15 volts) is regulated by an MC1461 IC voltage regulator and the low output (5 volts), by an LM309K regulator. Assuming you install your oscillator on the chassis with the remainder of the time base, provision is made for oscillator power from an external battery/trickle charger to supply standby power during power outages or short term transportation (J8).

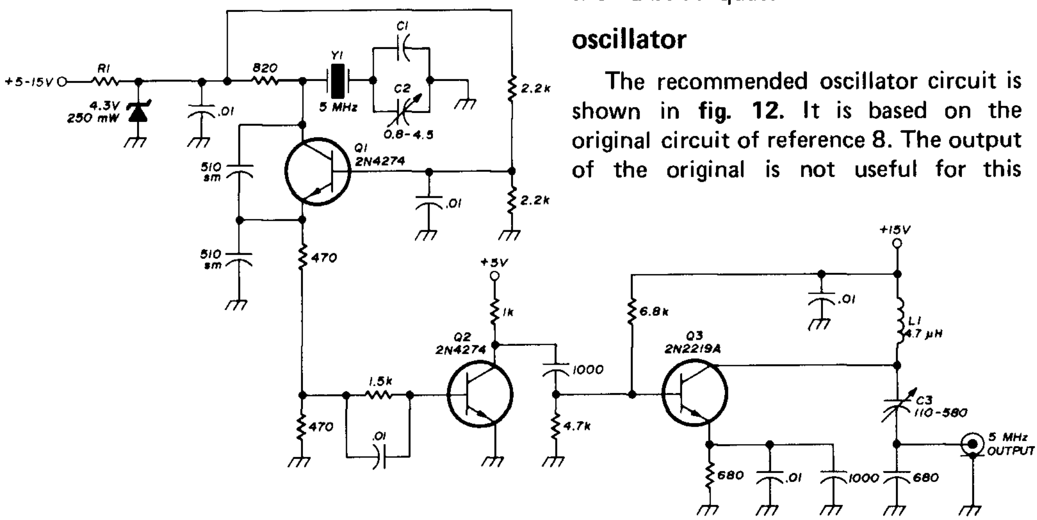
When driving Q1 with an external oscillator (J7), you will need to provide plus 20 dBm (100 mW) of 2.5-MHz signal or plus 15 dBm (30 mW) of 5-MHz signal. No tests have been made to determine how much 10-MHz signal is required to drive Q3/U1 directly but gain data indicate that zero to 10 dBm (1-10 mW) should be adequate.

oscillator

The recommended oscillator circuit is shown in **fig. 12.** It is based on the original circuit of reference 8. The output of the original is not useful for this application. The output of **fig. 12** will drive Q1 of **fig. 11** directly. This oscillator circuit may be built on a separate PC or perf board and included on the chassis with the circuitry of **fig. 11.**

Mount the oscillator on *top* of the chassis, not inside with other heat producing components. Provide as much thermal isolation as possible.

I have not built this oscillator, but the original circuit is well documented and the output stage is similar to the 2.955-3.155 MHz amplifier described earlier. Tuneup procedures for both amplifiers are identical.



C1	Temperature-compensating capacitor
C2	0.8 - 4.5 pF glass or ceramic trimmer
C3	110 - 580 pF trimmer (Elmenco 467)
R1	5-volt source, 36 ohms, 1/4 watt; 15-volt source, 510 ohms, 1/2 watt
Y1	5-MHz crystal (International HA-1)

fig. 12. Recommended time-base oscillator circuit.

It is possible to duplicate the double-balanced mixer shown schematically in **fig. 7** using easy-to-acquire parts. The mechanical details may be seen in the photograph of **fig. 13.** The enclosure is a Bud CU-2100-A Minibox. The connectors are BNC types (UG-625A/U). The circuitry is built on a piece of perf-board with holes on 0.1-inch centers. The board is cut just over the width of the transformers (0.75 inches) and approximately 1-3/4 inches long. Adhesive copper foil is applied to one side of the board; the two transformers are mounted from the other side.

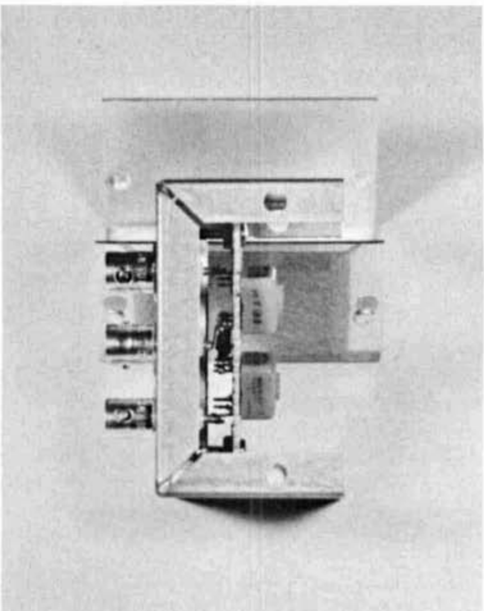


fig. 13. Construction of the double-balanced mixer shown in fig. 7.

The two sets of transformer terminals labeled 2, 3 and 6 are placed adjacent to each other on the board and the four diodes are used to interconnect these two windings. The copper foil can be cut away from non-grounded transformer terminals with a razor blade or X-acto knife. Grounded terminals are soldered to the foil with a small 25-watt iron.

The assembled perf-board is mounted diagonally in the Minibox. This allows each of the transformer pin-1 connections to be soldered directly to the center pin of a UG-625A/U connector. Be sure to drill the connector mounting holes on the correct centers for direct insertion of transformer pin-1 terminals. A short piece of bare tinned wire is required to connect the center tap of transformer T2 to the center coaxial jack.

Construction of filters A through C is very similar. A photograph of filter C is shown in **fig. 14.** The enclosure is a small Minibox; the size depends upon the number of filter sections. The connectors are female BNC chassis-mounting types (UG-625A/U). The filter components are mounted between the connector center pins, small standoff insulators (about 1/2-inch high) and strategically located ground lugs. Use dipped mica capacitors and Nytronics Wee-Ductors or equivalent for the inductors. Keep individual component leads short and the total circuit path as direct as possible. Where possible, arrange the inductors at right angles to one another for minimum mutual coupling.

Filter D is constructed using the same type of enclosure and connectors as the previous filters although the circuit elements are arranged differently. The general layout is shown in the photograph of **fig. 15.** This is the 134-MHz filter. Half-inch standoff insulators are used to support the inductors and are mounted as close to right angles to one another as possible.

The 2.955-3.155 MHz amplifier is also built into a Bud CU-2100-A Minibox. The circuitry is built on a small piece of perf-board with holes on 0.2 inch centers. The components are inserted from one side and the leads soldered together on the other side. The layout is shown in **fig. 16.** The board is mounted on 1/2-inch spacers. The input, output and power connectors are UG-625A/U. The power supply for the amplifier is built in a separate Minibox which is large enough to contain the components. No particular construction technique is required in this unit – just keep it small and neat.

The time base is built on a 5x7x2-inch aluminum chassis. The switches, connectors, pilot lights and voltage regulators are mounted on the front, rear and side panels of the chassis. The oscillator is mounted on the top surface and the circuit boards are mounted inside. **Fig. 17** is a photograph of the unit.

The rf and digital circuitry of **fig. 11** is built on the main board plus a small piggy-back board. The power supply and control wiring is accomplished in a point to point manner within the chassis. The main board contains the analog circuitry (Q1, Q2, Q3 and associated components) lengthwise on one side and the digital circuitry on the other. Q4 and its associated components are on the small board.

Outputs from the boards are via push-in terminals and attached sections of miniature RG-174/U coaxial cable. All board-mounted components are inserted from one side and leads soldered on the opposite side. The large board is mounted on the chassis top surface using two half inch spacers. The small board is mounted on a one-inch spacer screwed to the main board. Keep the coaxial cables carrying the output signals as short as possible. This is particularly true of the 5- and 10-MHz high impedance outputs.

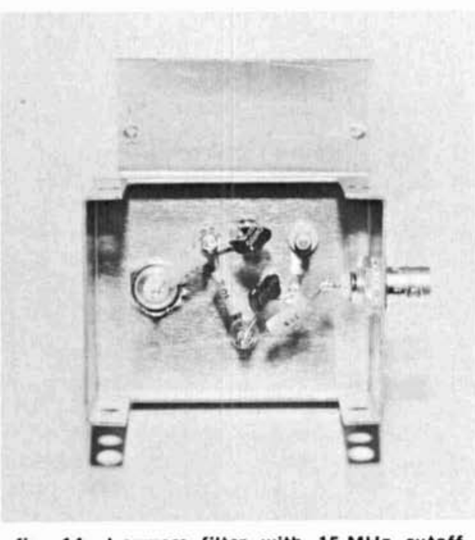


fig. 14. Lowpass filter with 15-MHz cutoff (filter C).

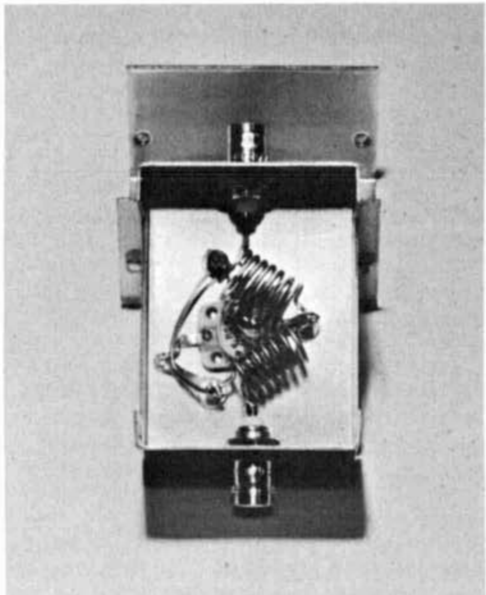


fig. 15. Construction of the 130-MHz bandpass filter (filter D).

If the oscillator is mounted on the chassis top surface, the terminals should protrude into the chassis interior for power and rf output connections. When wiring the power supply portion, use point-to-point wiring. It is important that all component leads be kept to a minimum length. This is particularly true of the MC1461 which has the larger number of external components and contains active devices capable of sustaining high-frequency oscillations. Use the same wiring precautions with both voltage regulators that you would use with any high-frequency rf circut. The parts location and wiring of the remainder of the power supply is not cricical.

adjustment

The i-f vfo and hfo signals must be tapped and brought out to the measurement system. I used the technique suggested in reference 1 but component values were revised slightly as shown in **fig. 18.** It may be necessary to vary the value of the 45-pF capacitor across L1 to maximize the i-f output level. With this

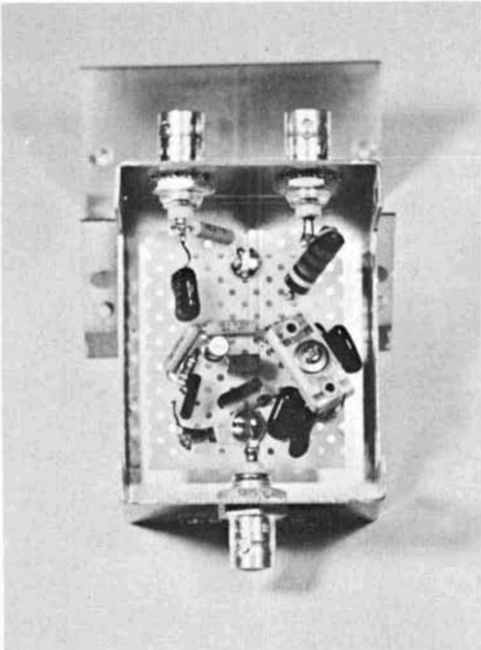


fig. 16. Layout of the 2.955-3.155 MHz amplifier module.

method of obtaining signals for the frequency measurement system I have observed no degradation to normal 75S-3 performance. If you desire further isolation for your receiver, you may use mosfet amplifiers such as used in the Heath SB-650.[5]

The vhf local-oscillator injection voltage is obtained from the 62S-1 by using a homebrew tee adaptor at the input to the 148-MHz trap (J31 of FL2 in the 62S-1). The tee adaptor is made from two standard phono pin connectors. The male portion is a phono plug to miniature phone jack adaptor (Lafayette 99R63455 or Switchcraft 365). To gain access to the center pin terminal, remove and discard the shell. The female portion of the adaptor is a single hole mounting phono jack (Lafayette 99R62341). The center pins of the two connectors are soldered together. The ground lug of the female connector is then connected to the remaining portion of the shell of the male connector.

The coaxial cable running to the frequency measuring system is then soldered (center conductor and shield) to the adaptor at the junction of the two component connectors. The male end is then inserted in J31 of FL2 and the cable which originally plugged into J31 now plugs into the female portion of the adaptor. The RG-174/U coaxial cable from the tee adaptor is then run *through* the center hole of unused phono jack (J13) on the rear apron of the 62S-1. Thus, no permanent change to the transverter is necessary.

Some experimentation with the length of the vhf local-oscillator cable and how it is terminated may be required. My cable is 41 inches long; if I leave the cable unterminated, a fairly low impedance is reflected back to the tee adaptor. Receiver and transmitter LO injection is thus reduced and vhf receiver performance and transmitter output power are affected.

If the cable is terminated in 50 ohms or lengthened about 12 inches and left unterminated, 62S-1 receiver and transmitter performance are normal. This effect could be completely eliminated by using a broadband 36 to 134-MHz amplifier with a high input impedance at the tee adaptor. Such an amplifier would have to be tailor fitted to the 62S-1. However, it is reasonably easy to obtain normal 62S-1 operation without resorting to a buffer amplifier.

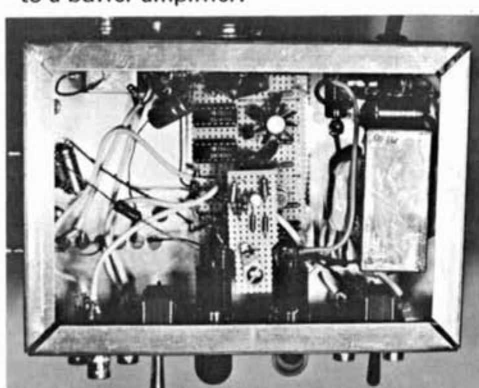


fig. 17. Bottom view of the time-base module. The MC1461 voltage-regulator IC and BNC connectors are mounted on the side of the chassis.

Using the circuit connections described here, you should get the power levels shown in **table 2** at the output end to the respective cables.

Filters A through C should not require any adjustment after construction. However, to insure that there are no faulty components, it would be wise to check the cutoff frequency and ultimate rejection of each filter. This will require a signal generator (with frequency calibration) and a volt-meter (or powermeter) with a frequency response higher than the highest filter cutoff frequency (15 MHz).

Filter D must be tuned-up. This will require a calibrated signal generator and voltmeter or powermeter with adequate frequency range (220 MHz). A swept-frequency generator plus detector would be better. In any event, C1 sets the filter center frequency and the tap points on L1 and L2 determine bandwidth and insertion loss. The taps given in **fig. 9** should be satisfactory. Capacitor C1 should be tuned to the offset frequency region of interest. If you want to measure frequencies more than a few hundred kilohertz apart, you should re-peak C1 for each different region.

table 2. Signal levels available from the Collins 75S-3 receiver and 62S-1 transverter.

signal	condition	level (into 50 ohms)
i-F	S9 signal	−5 dBm
VFO	—	−7 dBm
HFO	80 and 40 meters	+3 dBm
HFO	20 and 15 meters	−7 dBm
HFO	10 meters	−10 dBm
VHF LO	—	+5 dBm

*Although the mixers have no adjust*ment provisions you should insert a known rf and LO signal and check for proper mixing action and suppression of undesired feedthrough products. Feedthrough from rf and LO ports to the i-f port should be more than 20-dB down, referenced to the input levels at the rf and LO ports, respectively. Performance less than this may indicate a bad or mismatched set of diodes.

To properly tuneup the 2.955-3.155 MHz amplifier, a calibrated signal generator and voltmeter or rf powermeter operating to beyond 3 MHz are required. Set the signal generator to 3.055 MHz and adjust C1 for maximum gain (about 25 dB). To optimize the combination of filter A and the 2.955-3.155 MHz amplifier, connect the two in tandem and run response measurements at 2.955, 3.055 and 3.155 MHz. It should be possible to skew the response curve of the amplifier down in frequency to make the overall gain at 2.955 MHz more nearly equal to that at 3.155 MHz (on the low side you have filter *and* amplifier rolloff, while on the high side you have only amplifier rolloff).

Your vhf or broadband amplifier plus filter D should be initially adjusted for 20-25 dB overall gain at the offset frequency to be measured. For example, if you are measuring frequencies near the low end of two meters, the vhf amplifier or filter should be tuned to 134 MHz. Ultimate gain adjustment will depend on counter sensitivity.

The only adjustments necessary to the time-base module will be to the analog section. The tank circuits for Q2 and Q3 may be optimized by starting with slightly lower value capacitors and increasing the value with small padders until the respective outputs peak. Varying the value of the 62-pF capacitor from pin 12 of U3 to ground may result in a cleaner 100-kHz output waveform.

The value of the resistor from pins 8 to 9 of the MC1461 voltage regulator IC determines the high output voltage. A value of 24k ohms results in about 15 *volts out. This value is not sacred;* If you want to change the voltage for your oscillator, it may be done. Lowering the voltage will result in less 10-MHz (low impedance) signal output.

If you increase the value be careful not to exceed transistor breakdown voltage. To increase the voltage substantially, you will need a different power transformer. The maximum current drain for the MC1461 regulator is 500 mA; be sure your oscillator load does not exceed this value. If you use a non-oven oscillator there should be no problem.

operation

Typical offset counter readings for specific received frequencies on each band are shown in **table 3**. It should be evident from this table that on all bands except 80 and 40 meters you simply add

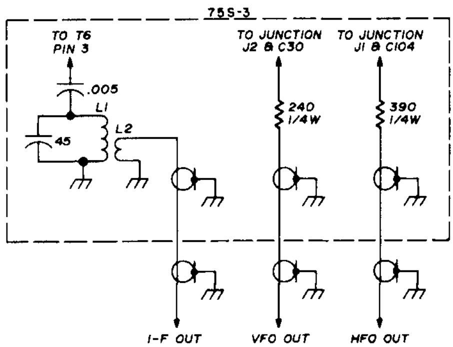


fig. 18. How to connect the Collins 75S-3 receiver into the offset frequency-measuring system. L1 is a 2.5-mH rf choke (miller 6302). L2 is 5-1/2 turns no. 26 wound between two end pi sections of L1.

10.000 MHz to the counted frequency to arrive at the actual received frequency. On 80 and 40 meters you subtract the counted frequency from 10.000 MHz to obtain the received frequency.

80 and 40 meters can be offset in a manner so that to obtain the received frequency you simply subtract 10.000 MHz from the counted frequency. This is accomplished by making the second mixer (**fig. 5**) a difference mixer and the third mixer a summing mixer. However, there are two disadvantages to this scheme. First, you have to change or bandswitch new filters at B1, B2 and C; second, you cannot count 80 and 40 meters on a 10-MHz counter. For these reasons, I did not use this arrangement.

If a spectrum analyzer or frequency-sensitive voltmeter is available when you are initially assembling the modules into the system, it may be enlightening to observe the output spectrum of each mixer/filter combination. The undesired mixer products should all be down approximately 20 dB. If they are not, the individual mixers and/or filters should be checked for proper operation.

When installing the mixers, be aware that there is an rf port (R), LO port (L) and an i-f port (I). I have achieved best conversion loss and isolation performance in this up-conversion application by applying the lowest frequency to be mixed to the i-f port and taking the up-converted output from the rf port.

In **fig. 5** no amplification is shown following filter C for HF operation. If your counter doesn't have a built-in preamplifier, you may need to include one or more stages of gain at this point. The amount of gain and whether it be provided by tuned or wideband stages will be determined by things such as counter sensitivity, amplifiers presently available to you and the state of your junk box and/or pocketbook.

System calibration may be achieved by setting your time-base oscillator against a laboratory standard of known frequency or by using WWV. To use WWV with the 75S-3, tune the receiver to 15 MHz and during a period of no tone modulation, place the receiver in the CW mode (narrowest bandwidth) and center the WWV carrier in the passband. Your counter should now read 5.000 MHz plus/minus 1 count offset frequency. If it does not, adjust your time base oscillator until it does. If the counted frequency is not even remotely close to 5.000 MHz, do not adjust the oscillator, something else is wrong and this must be corrected first.

table 3. Typical offset frequencies for each band (all in MHz).

band	received frequency	offset frequency (counted)
3.5	3.774368	6.225632
7.0	7.024732	2.975268
14.0	14.251051	4.251051
21.0	21.246188	11.246188
28.0	29.499987	19.499987
50.0	50.101687	40.101687
144	146.160752	136.160752
220	222.561357	212.561357

Be sure that you make this adjustment when propagation conditions between WWV and your location are quiet. Typically, this will be during a daylight period at both WWV and your location. If the WWV signal is varying widely in strength or is of marginal strength, postpone the adjustment.

System stability on all bands should be excellent. No whistles, howls or low-frequency motorboating should be heard with the receiver operating normally and the measuring system activated. The ultimate test is to have the normal station antenna connected to the receiver and the counter on and counting. If your system is stable under these conditions on all bands, you are in business. If not, suspect mixer leakage as the primary culprit.

Leakage of 2.955-3.155 MHz energy through the second mixer, filters B1 and B2 and the third mixer is the main sneak path. Another possibility is radiation from the 2.955-3.155 MHz amplifier to the third mixer. If you have one mixer with particularly good rf/i-f port isolation, use it for the second mixer.

summary

Results at my station have been most rewarding. Previous systems based on signal synthesis proved to be oscillatory, very difficult to tune up and always in need of tweaking. The present offset system provides the capability of making frequency measurements on incoming hf

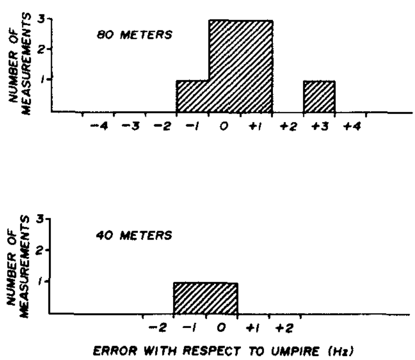


fig. 19. Histograms of offset frequency measuring system error vs umpire readings in recent ARRL Frequency Measuring Tests.

signals at least 20 dB weaker than with my best performing signal-synthesis scheme. On two meters, system sensitivity has been increased by more than 40 dB using the offset technique.

Accuracy in the ARRL Frequency Measuring Tests (relative to umpire measurements) has averaged better than 0.2 parts per million for ten measurements submitted during the last 20 months. Most of these measurements have been on the lower amateur frequencies (80 meters) which results in the average absolute error being less than 1 Hz. A histogram of my system error relative to ARRL FMT umpire measurements is shown in **fig. 19.**

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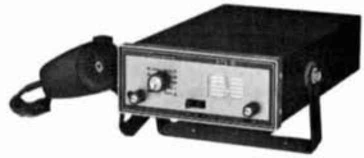

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electronic bandpass tuning in the new Drake R4C

The new Drake R4C amateur receiver uses a long-neglected technique to provide a simple system for electronic bandpass tuning

D.H. Horner, 186 Golfwood Drive, Dayton, Ohio 45449

Many amateur operators believe bandpass tuning is an invaluable adjunct to the main communications receiver in their station. It enables one to adjust the exact bandpass of the receiver in relationship to the bfo frequency so that interference from other stations can be dodged, another operator's voice characteristics can be made more pleasant, or a pleasing pitch can be obtained from CW. Also, the proper bandpass can be set for RTTY reception.

Many years ago the Collins Radio Company incorporated bandpass tuning in their 75A-4 receiver by rotating the PTO assembly at the same time as the bfo capacitor was tuned. This gave the advantage of bandpass tuning along with the flat-topped and steep skirt selectivity afforded by the Collins mechanical filter. This was accomplished mechanically, of course, and worked well because Collins maintained extremely close manufacturing tolerances; zero-beat moved no more than 50 Hz during the adjusting process.

The R.L. Drake Company has featured bandpass tuning in many of its amateur communications receivers by tuning the resonant frequency of a four-pole LC filter operating at 50 kHz. The bfo frequency remains fixed, and the actual i-f is tuned to one side or the other by an amount selected by the operator. This system has proved to be very successful, and is readily attested to by the continued popularity of the Drake receivers.

There is a small drawback, however. With the advent of mechanical and crystal filters, 6-dB to 60-dB skirt ratios in the order of 2:1 or even 1.7:1 are now expected by the amateur fraternity, with

in the BC-312 was turned into an injection oscillator, a mixer stage was added, and the i-f was mixed down to some lower frequency.

Some builders made their own 50 kHz i-f amplifiers; others cannibalized 85-kHz i-f transformers from the ubiquitous BC-453 command set. Then – and this is the crux of the system – the same bfo was used to mix back to 915 kHz where an added oscillator and detector operated. By varying the frequency of the

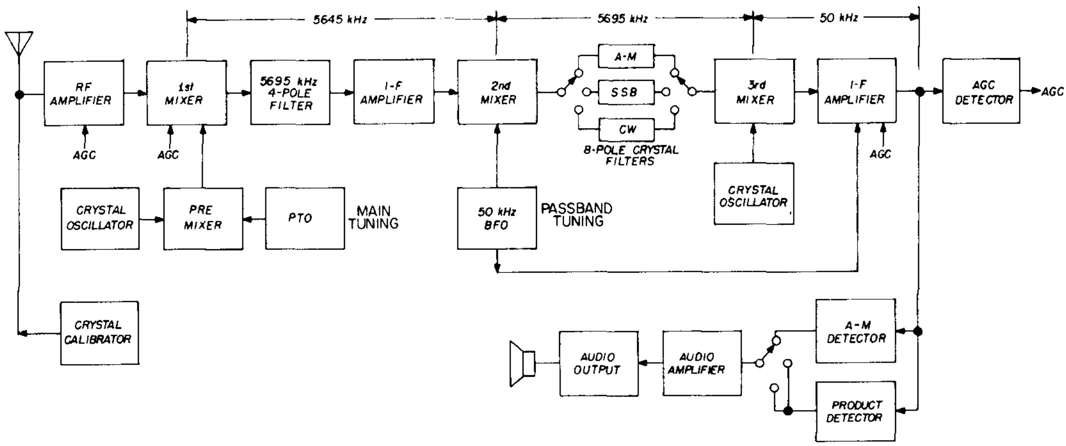

fig. 1. Basic block diagram of the new Drake R-4C communications receiver which features electronic bandpass tuning.

the ultimate stopband rejection in the order of 80 dB or so. The 50-kHz LC filters do not give this magnitude of performance.

Right after World War 2 a large segment of amateurs purchased sophisticated surplus communications gear at a fraction of the original government cost. Among the units available was the venerable BC-312, a military communications receiver using design techniques not followed today. This set used a last i-f of 915 kHz, and was rather broad. When ssb became popular, some sort of sideband selection became necessary. Some clever amateurs devised a system where the bfo

vfo, which now operated at a frequency removed far enough from 915 kHz to mix to the new i-f frequency, the signal could be made to sweep across the low frequency i-f, and bandpass tuning was the result.

To the extent of my knowledge this technique remained in limbo until the R.L. Drake Company announced the new R-4C Amateur Receiver in March, 1973. This receiver uses electronic bandpass tuning coupled with the selectivity afforded by 8-pole crystal bandpass filters. However, the mixing scheme is different from that used with the BC-312 (see **fig. 1**).

The rf signal is first converted to 5645 kHz where a 4-pole crystal-lattice filter does much to ward off the degrading effects of strong stations near the operating frequency. Converting all bands into the high i-f is accomplished by mixing the output of the vfo with various crystal-controlled frequencies so the i-f is the same for all bands. Then, a tunable oscillator at 50 kHz is used to up-convert to 5695 kHz. The standard 2.4-kHz ssb filter as well as the accessory filters for CW and RTTY operate at this frequency. Then, the 5695-kHz signal is mixed with a 5645-kHz crystal oscillator to obtain a 50-kHz output. Two tuned LC circuits at 50-kHz help establish "distributed selectivity."

The same tunable oscillator that is used to mix from 5645 kHz to 5695 kHz also serves as the bfo. Because it both up-converts and down-converts (i-f to audio) at the same time and by a similar amount, zero-beat does not vary during adjustment of the bandpass tuning control.

The Drake R-4C Communications Receiver also has provision for a-m. However, as bandpass tuning is not needed in this mode, the additional conversion process is not used. Filters with 4.0- and 6.0-kHz selectivity are stocked by the Drake Company, but these are at 5645 kHz, rather than at 5695 kHz where the ssb and CW filters operate. In the a-m mode a different crystal frequency (5595 kHz) is used to convert to the 50 kHz i-f.

This dual-conversion electronic bandpass tuning technique is simple and relatively inexpensive for the manufacturer to build, and it makes operating much more pleasurable than with receivers which have fixed bfo/filter frequency relationships. One benefit of this particular bandpass tuning technique is less noise in the audio output — receivers that lump all the selectivity in the filters and then use broadband i-f amplifiers following them tend to have "broadband hiss" or white noise appear at the output. The Drake R-4C is pleasantly free from this effect.

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SPECIFICATIONS

GENERAL: • **Frequency Coverage:** 144 through 148 MHz. 12 Channels, 2 supplied: (1) Receive: 146.52 MHz, Transmit: 146.52 MHz; (2) Receive: 146.94 MHz, Transmit: 146.34 MHz • **Power Requirements:** 13.0 Volts DC±15% • **Current Drain:** Transmit: 450 mA, Receive: 45mA • **Antenna Impedance:** 50 Ohms • **Dimensions:** 5⅜" x 2$\frac{5}{16}$" x 7½" (13.6 x 5.8 x 19.1 cm) • **Weight:** 3.75 lbs (1.7 kg)

RECEIVER: • **Sensitivity:** Typically .5 microvolt for 20 dB quieting • **IF Selectivity:** 20 kHz at 6 dB down; ±30 kHz channel rejection greater than 75 dB down. • **First IF:** 10.7 MHz with 2-pole monolithic crystal filter. • **Second IF:** 455 kHz with ceramic filter. • **Intermodulation Response:** At least 60 dB down. • **Modulation Acceptance:** ±7kHz. • **Audio Output:** At least 1 Watt at less than 10% distortion. • **Audio Output Impedance:** 8 Ohms

TRANSMITTER: • **RF Output Power:** 1 Watt minimum • **Frequency Deviation:** Adjustable to ±10 kHz maximum, factory set to 6.0 kHz. • **Multiplication:** 12 Times

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More Details? CHECK—OFF Page 110

the ham notebook

portable fluorescent light

Have you ever wanted one of those portable fluorescent lights to use during field day or on camping trips? They are quite expensive and are not used too often, but they're handy to have around. If you have a well supplied junkbox and a little ingenuity, you can build your own.

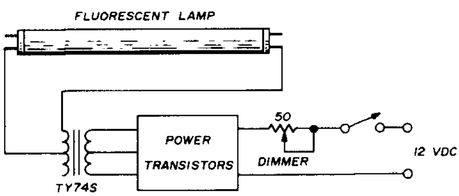

fig. 1. Portable fluorescent light uses transistor power transformer cannibalized from an old CB set. Vibrator power supply can also be used as described in the text.

Most amateurs have some old portable vacuum-tube equipment or CB rigs laying around that can provide most of the parts you need. Try to find a Triad TY74S transistor power transformer or something similar. This will provide the high-voltage supply for your portable fluorescent light.

Remove the power transformer and the power transistors from the old rig, but discard the rectifier diodes and the filter network. You won't need them. Build a metal or wooden framework similar to that found on commercial portable fluorescent lights so you can mount the fluorescent lamp with sockets. The switch, transformer and power transistors can be mounted on the back or in the base.

When you're putting the unit together be sure to use a heatsink. If the light is used continuously there will be a fair amount of heat generated. However, the heat sink doesn't have to be anything fancy, a small aluminum chassis will work.

Wire up the unit as shown in **fig. 1** using the original transistor oscillator circuit. Connect the center tap of the power transformer to one end of the fluorescent lamp and one of the transformer end windings to the other end of the lamp. If you want to include a dimmer control place a 10-ohm, 50-watt rheostat in the primary lead as shown in the schematic. My unit required from 0.5 to 1.6 amps, depending upon the setting of the dimmer control.

If you can't find an old high-voltage transistor power supply, you can accomplish the same thing with an old vibrator power supply. These are plentiful – just get an *old* car radio from your local junk yard. Disconnect the filter network and bring out the center tap from the vibrator transformer to one side of the fluorescent tube. Run the other lead to the other end of the tube as shown in **fig. 1.**

Ken Gray, K8BYO

great circle charts

A number of Great Circle Distance and Azimuth Charts are available from United States Government sources. These may be obtained from numerous local sales agents and distribution centers, or by mail from the U.S. Naval Oceanographic Office, Washington, D.C. 20390.

Such charts are available at $1.50 centered as follows: 5180, Fairbanks, Alaska; 5181, Seattle; 5182, Honolulu; 5183, Guam; 5184, San Francisco; 5185, Washington, D.C.; 5186, Moscow; 5187, Adak, Alaska; 5188, Kodiak, Alaska; 5189, Eniwetok Atoll; 5190, San Diego, Calif.; 5191, Cutler, Maine; 5192, Balboa, Canal Zone; 5193, Yosami, Japan; 5194, Australia-N.W. Cape; and 5195, Keflavik, Iceland.

For those purchasing a computer readout, possibly they can plot the results conveniently on a 5142 Azimuthal Equidistant Projection of a Hemisphere, priced at 50 cents.

Bill Conklin, K6KA

using the HW-101 transceiver with a separate receiver

Modifications to transceivers such as the Heath HW 100/101 or Swan 350 for split frequency operation have been described as has incremental tuning by means of a variable capacitance diode[1] or operation in conjunction with a separate vfo.[2] In both cases some limitations remain; cross-band operation is not possible and the separation between receive and transmit frequency must remain small because the driver and receiver preselector circuits cannot be independently peaked.

Conversion for use with a separate receiver, using the built-in TR relay, is simple and eliminates these limitations. It is described here for the case of the HW-101.

Two rf-quality phono jacks are installed on the rear chassis apron into the compartment containing the antenna connector, located just below the rf cage. The coax lead from the rf driver circuit board (the receiver input) is disconnected from lug 4 of the relay and from the adjacent ground lug. If sufficient in length, this coax is connected directly to the nearer of the two jacks just installed. Otherwise, it is first extended or replaced by a longer section. Another 50-ohm coax lead is connected to the more distant of the two jacks at one end and to lug 4 of the relay (inside conductor) and to the adjacent ground at the other end. The trap coil (L905) is not disturbed.

In the transceive mode an external coax connector (50-ohm with two rf-quality phono plugs) is jumpered between the two new jacks; the transceiver then functions as before the modification. For separate receive operation a 50-ohm coax lead is connected between the receiver antenna input and the jack, which is fed from lug 4 of the relay. The other jack remains unconnected. To spot the transmit frequency on the receiver the mike/CW level control is turned counterclockwise below the point where the relative power meter indicates any measurable output; the transmitter is then keyed. This provides sufficient signal for spotting without radiating an appreciable amount of power. All that is needed to transmit is to advance the mike/CW level control to the point of full rf output.

This simple modification leaves the transceiver circuitry essentially undisturbed and permits semi break-in operation, cross-band or in-band operation with any desired frequency separation between receiver and transmitter, with both tuned for peak performance.

Incidentally, the modification permits the insertion of an rf-preamplifier, tuned or broadband, between the output coming from the TR relay and the internal or external receiver. A low-noise preamplifier improves the signal-to-noise ratio, especially on 10 and 15 meters.

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Max Blumer, WA1MKP

comments

ic speech clipper

Dear HR:

Although K6HTM seems satisfied with the performance of his speech clipper, his conclusion that it's "not much better than a typical agc-type compressor" caused me to take a critical look at the schematic. The main reason for what I would regard as poor performance is fairly obvious. The output impedance of pin 9 of the IC, to which the clipping diodes are connected, is that of an emitter follower and given as 85 ohms on the data sheet.

If author Bird uses 1N75 diodes for the clipper he would obtain virtually no clipping since their ac resistance is likely to be several hundreds of ohms. Even with hot-carrier diodes, clipping will not exceed 4 dB or so. This would definitely make the device perform comparable to alc or agc schemes! A resistor (2000 ohms for hot-carrier diodes, 20k for 1N75s) in series with the 0.01-μF coupling capacitor will improve matters.

I am distressed at the author's lack of concern for carrier rejection. Unless the peak carrier level, at pin 9 of the LM373, is appreciably lower (at least 10 dB), than the voltage at which clipping starts, the advantages of rf clipping will be lost. K6HTM probably forgot to mention the need for precise frequency adjustment of the crystal oscillator, so that the proper low frequency response is achieved, as well as some additional carrier rejection.

Some builders may experience difficulty, and possibly instability, due to undesired coupling via the two oscillator injection points, which are, in fact, connected together. Fortunately, the level at pin 6 of the LM373 is much higher than needed (by a factor of 100), so the desired isolation can be obtained by means of a resistive attenuator.

The second mechanical filter, FL2, is quite superfluous. Filtering after the clipper is needed to prevent rf harmonics from reaching the final mixer. A moderate Q pi network is quite adequate for this purpose and obviously much less costly. As the available post clipper gain is much larger than needed, two 10,000-ohm resistors, one in series with "high" side of the output level control, the other in series with the FL2 input, will give desirable isolation of the filter from the clipping diodes.

If Mr. Bird will modify his device in accordance with this letter, he will obtain results similar to the ones I get with the Comdel unit.

Walter Schreuer, K1YZW
Ipswich, Massachusetts

I am pleased that Walter Schreuer read my article, "IC Speech Clipper" in the February, 1973, issue of ham radio. *What follows is a reevaluation encouraged by his suggestions and some additional information.*

The 1N75 diodes first used in this project were abandoned early in favor of hot-carrier diodes in both the balanced modulator and clipper sections. Carrier balancing with these devices far exceeds the 10-dB figure referred to by Mr. Schreuer, and the total suppression after the first filter, FL1, is further enhanced by placing the carrier frequency 20-dB down the filter slope as recommended by Collins.

The surplus crystal referred to may

not necessarily meet this requirement. Two or three could be tried, but ultimately the builder may wish to warp the crystal frequency, or he may wish to modify the crystal itself. I have found that regardless of the carrier position, greater roll-off of high frequency audio is desirable at the output of the device. Adding a simple series RC network (R = 10k pot, C = 1 µF) from pin 7 of the LM373 to ground allows more latitude in tonal balance.

I do not find that adding any of the suggested resistors really improves matters. There is an overall loss of system gain necessitating compensation with the transmitter gain control. Instability is not a factor in my device, but the suggested isolation may be helpful, especially at high frequencies. Concerning resistors, one which could be left out of the circuit diagram is incorrectly shown at the input to FL1 as going to ground. If retained, it should be connected to the other filter terminal.

The real problem in comparing the clipper to the compressor, it seems to me, has more to do with the limiting action of the transmitter alc than with any defects in the clipper. With or without the 2k ohm resistor suggested by Mr. Schreuer, severe clipping is possible (my data sheet indicates 70 ohms at pin 9). Heavy driving of the device will produce a solid carrier-like bar on an oscilloscope trace at the clipper diodes.

Power gain measured on a wattmeter appears to be about 4 dB with the compressor and 7 dB with the clipper when the input levels to the transmitter are equal. I note that Comdel claims 10 dB for their device, Magnum-Six, 6-dB, and DX Engineering, 5-dB power gain. But it is possible to drive the transmitter harder with the compressor and achieve 5 to 6 dB gain without flattopping. This observation is the basis of my conclusion, which has not changed.

With this condition I made the following table showing how the compressor and clipper treat pure vowel sounds and two voiceless consonants. English vowels are diphthongized and, therefore, unsustainable. So I have used the French vocalic system.

		compressor	clipper
a	a	5.18 dB	6.02 dB
e	e	3.97 dB	6.02 dB
i	i	7.78 dB	10.00 dB
o	o	2.55 dB	3.80 dB
u	y	7.78 dB	9.03 dB
s	s	11.76 dB	11.76 dB
t	t	6.99 dB	6.99 dB
average		6.57 dB	7.66 dB

I have long felt, as does Mr. Schreuer, that the second filter is superfluous and could be replaced by a tuned circuit. The statement in the ARRL Radio Amateur's Handbook, *1970, page 258, ". . .a filter as good or better than the filter used to form the original ssb signal," in my opinion, only serves to confuse the issue.*

I used two filters because I had them. Two identical filters might actually have the disadvantage of reinforcing certain peaks and valleys in the passband.

Regardless of the refinements suggested, this device will add considerable authority to an otherwise unmodified signal.

Charles G. Bird, K6HTM

vertical antenna

Dear HR:

The article by VQ9N in the December, 1972, issue of *ham radio* on a single-element vertical antenna caught my eye immediately as it is similar to the antenna I have been using for about ten years.

When I moved to my present location I was unhappy with the antenna prospects. There is a 12,000-volt Edison line across the back of the lot about 60 feet from the shack with a maple tree in between. Putting up an antenna was a problem. After much head scratching, I finally took three sections of aluminum tv mast (about 30 feet), set it on a soft-drink bottle alongside the house with an insulator extended from under the eaves of my two-story house. A single wire was run through a basement window about three feet to a tuner which is coax fed to the transmitter. I used the hot-water heating system for ground.

While reports are not spectacular, I get into the East Coast with good reports. My operating time is very limited so I have not calculated the engineering parameters of the system, but it is about as simple as you can get. I have used the antenna on 80, 40 and 6 meters but never on 20 where it would be most efficient. There are no guy wires. I would be the first to agree that it is makeshift, but it works. If you need an antenna in a hurry, or have a limited area, try it. My only problem is the tv service men who try to sell me a new tv antenna for the mast!

Paul R. Smith, W8FHB
Toledo, Ohio

Dear HR:

I was very interested to read the article by VQ9N on the use of a half-wave vertical antenna. For many years as VS4RS I used a 14-MHz version of the J-vertical described in the vhf section of the *ARRL Antenna Handbook.*

This antenna is essentially a half-wave vertical situated a quarter-wave above ground. My arrangement was a 50-foot guyed steel pipe with a 17-foot matching section at ground level, fed with 600-ohm open-wire feeder. No ground system is required.

Deducting two S-units for the rarity of the VS4 prefix, I am still convinced it was a very effective DX antenna.

Ron Shelton, 6Y5SR
Kingston, Jamaica

fetrons

Dear HR:

After reading your comments in the August, 1972, issue of *ham radio* regarding Fetrons, plus hearing some scuttlebutt from a ham in Seattle, I decided to give them a try.

I work for Pacific Northwest Bell in the Portland, Oregon, mobile telephone and two-way radio shop. We have some GE Progress line 150-MHz base-station receivers, and I thought we could upgrade them with the Fetrons. To make a long story short, Fetrons are very good attenuators at 150 MHz.

After going around in circles a few times I called Teledyne. Their first question was, "at what frequency are you using them?" When I indicated it was 150 MHz, they told me that was the problem. Contrary to the specification sheets, which call for a 500-MHz upper frequency limit, the upper usable frequency for current production Fetrons is 10 MHz.

The TS6AK5 and TS12AT7 Fetrons are being manufactured for telephone carrier systems, which have a top frequency of about 4.5 MHz. Teledyne indicated that Fetrons *could* be made to work at 150 MHz, but they do not at the present time.

As you pointed out in your editorial, most fetrons are designed for the lower frequencies. Before amateurs spend some of their hard-earned money on the current TS6AK5 and TS12AT7 Fetrons, they should realize that they are not usable much above 10 MHz.

Walter J. Loomis, K7BQE
Portland, Oregon

current limiting

Dear HR:

The December, 1972, issue of *ham radio* carried an article on adding current limiting to existing solid-state power supplies. I have used the circuit in the article, roughly as shown.

I tried other versions of current limiting and for simplicity's sake, I settled on one similar to the one in the article. As shown in the schematic, there is still one hangup that needs to be overcome. There is no protection against short circuits for the *current-limiting* transistor. Several deceased transistors attest to this fact. There are two causes for this. First, the instantaneous output voltage of the power supply appears across *R1,* the current-sensing resistor, at the onset of a short.

This places 15 volts, in this instance, across the base-emitter junction of Q1. This is several times the voltage rating of most transistors. To overcome this problem, placing a 1000-ohm resistor (ballpark figure) in series with the base lead to Q1 effectively gives it some current-limiting protection without unduly adversely affecting performance.

A second, and less likely to happen, transistor failure is due to the fact that Q1 is called upon to instantly discharge the electrolytic capacitor that is usually placed across the reference zener. This involves high peak currents with higher voltages and large value electrolytics. A 10- or 15-ohm resistor in series with the collector lead of Q1 limits this current. It is recommended that a fairly high beta transistor with at least 600 mW of dissipation be used at Q1.

The addition of these two resistors is a simple and inexpensive way to make the supply truly current limiting with very little adverse effect on operation.

Donald G. Cheshier, K5MKO
Garland, Texas

cooling fan error

Dear HR:

In the November, 1972, issue of *ham radio* the *ham notebook* section contains an erroneous concept of the nature of reactance. Author WB8IUF, in an attempt to lower the voltage supplied to a cooling fan, placed a capacitor in series with it. This is legitimate in itself, but he has fooled himself.

A capacitor is *not* equivalent to a resistance value equal to its reactance. The impedance also has a phase angle of 90 degrees. This means that the total impedance of the fan-capacitor combination is not the simple sum of the two impedances, but is the vector sum. This can be found from

$$Z_T{}^2 = R_F{}^2 = X_C{}^2 = 1200^2 + 442^2$$

$$Z_T = 1280 \text{ ohms}$$

where Z_T is the total impedance, R_F is the resistance of the fan and X_C is the reactance of the capacitor. In this case, where the total impedance is 1280 ohms, the current is 93.8 mA and the voltage across the fan, V_F, is 112.5 volts. Thus, V_F is considerably higher than that given by WB8IUF.

To find the correct value capacitor for this application use the following equations:

$$Z_T = \frac{V_T}{I} = \frac{120 \text{ V}}{75 \text{ mA}} = 1600 \text{ ohms}$$

$$X_C{}^2 = Z_T{}^2 - R_F{}^2 = 1600^2 - 1200^2$$

$$X_C = 1060 \text{ ohms}$$

At the line frequency of 60 Hz, 1060 ohms capacitive reactance is provided by a 2.5-μF capacitor. That gives the desired 90 volts across the fan.

Terry Conboy, WB6GRZ
Redwood City, California

yaesu spurious signals

Dear HR:

Reference, "Spurious Signals with the Yaesu," December, 1972, issue of *ham radio,* page 69. Unfortunately, the article by K6KA was incomplete with regard to which units of the FTdx560 can be affected by the spurious radiation problem.

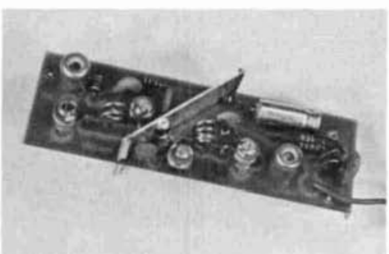
Starting with serial number 30001, the vfo frequency range in the FTdx560 was changed from 8400-8900 kHz to 8700-9200 kHz. The resulting change in local-oscillator frequencies produced a heterodyne with the second harmonic of the 3180 kHz i-f in some units to produce the spurious output. This was eliminated by addition of the 6358.6-kHz crystal.

All FTdx560s manufactured after introduction of the FTdx570, and all FTdx570s, have this circuit modification incorporated during production.

James Young
Spectronics
Signal Hill, California

new products

450-MHz mosfet preamp



Topeka FM Engineering has announced their new line of dual-gate mosfet preamplifiers. Their new model HF450 is designed for use in the range of 406 to 470 MHz, and is available in three models: 406 to 430 MHz, 430 to 450 MHz and 450 to 470 MHz. The rf voltage gain of these amplifiers is typically 15 dB with a 10-to 15-volt dc power supply. Noise figure is typically 4.5 dB. Superior cross-modulation performance and the greater dynamic range of the mosfet greatly reduces spurious responses. Each mosfet gate is protected by back-to-back diodes.

These new preamplifiers feature rf shielding of input and output circuits on both sides of the printed-circuit board and silver-plated G11 epoxy-glass boards for high performance. The preamplifier comes complete with all mounting hardware, rf jumper and detailed instructions. The HF450 is priced at $29.95 shipping prepaid. Also available is the model HF450 MO for Motrac radios. For more information, write to Topeka FM Communications and Electronics, 1313 East 18th Terrace, Topeka, Kansas 66607, or use *check-off* on page 110.

free electronics catalog

One of the biggest problems for the amateur home builder is obtaining the necessary electronic components for his projects. This problem is particularly bad if you don't live in or near a large metropolitan area. The new 1974 Olson Electronics catalog will solve many of these problems. This new, illustrated catalog, which features over 8000 quality-tested items, is free for the asking. In addition to a complete line of electronic components, transistors, ICs, etc., Olson carries the best of name brands in amateur radio equipment, antennas, test equipment, kits, electronic calculators and stereo and four-channel sound systems. For your free copy, write to Olson Electronics, Dept. HH, 260 S. Forge Street, Akron, Ohio 44327, or use *check-off* on page 110.

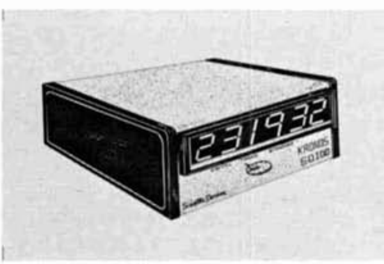
tube substitution handbook

Since 1960, there has been a tremendous increase in the number of American and foreign models of home-entertainment equipment. Consequently, an expansion of more than 600 percent has taken place in the number of available tube substitutions. This new and up-to-date edition of the Howard Sams *Tube Substitution Handbook* has kept pace with this rapid expansion by listing over 12000 direct replacements for all types of receiving tubes.

For convenience, this handy guide is divided into seven informative sections. Section 1 presents a cross-reference of all American receiving tubes, and section 2 lists picture tubes and their recommended substitutes. Section 3 contains a cross-reference of subminiature tubes, while the fourth section consists of industrial substitutes for receiving tubes. The fifth section is a substitute listing for communications and special-purpose tubes. The final two sections feature cross-references of American and foreign tubes.

There are easy-to-follow instructions accompanying each section that help you make proper tube substitutions and that explain how to cross-reference between sections for other substitutes. This guide fills the need of electronic experimenters, radio amateurs and service technicians who desire quick and accurate information for making suitable vacuum-tube substitutions. 96 pages, softbound. $1.75. A companion pocket-size volume which will fit into your pocket or your tube caddy is also available in a twin-pack which includes the regular sized volume, $2.25. Order from Comtec Books, Greenville, New Hampshire 03048.

digital chronometer kit



The Kronos KR100 chronometer kit features an LSI National clock chip, and a 32-page brochure with pictorials and easy-to-understand, step-by-step instructions. The chronometer includes 3 setting controls, 1-hour per second, 1-minute per second, and hold button. Easy-to-change from 12 to 24 hours, 4 to 6 digits, 50/60 Hz operation.

There are 3 models to choose from: 7-segment MAN-3 type LEDs, 6-digit kit, $47.00; 7-segment MAN-1 type LEDs, 6-digit kit, $69.95; and 7-segment Nixie type tube kit for $47.00. Available from Poly Paks, P.O. Box 942H, Lynnfield, Massachusetts 01940. For more information use *check-off* on page 110.

drawing template catalog

Tangent Template, Inc. has published a short-form catalog that gives quick reference to a comprehensive line of user-designed drafting templates for printed-circuit design work. A total of fourteen different templates are described and illustrated, including templates for 1:1, 2:1 and 4:1 reproduction ratios. If you design PC boards, these templates should prove to be very useful. For a copy of the catalog, write to Tangent Template, Inc., Post Box 20704, San Diego, California 92120, or use *check-off* on page 110.

phone patch

New from Radio Shack is the Realistic Phone Patch at a price which the company says brings this useful accessory within the reach of any amateur's budget.

A phone patch provides an interconnection between your station's equipment and the telephone system, making it possible to place or receive telephone calls through a base station and relay them to another station which does not have access to a telephone. Phone patches

More Details? CHECK—OFF Page 110

have been of great use during civil emergencies, in providing communications in disaster areas, and often as a means for servicemen overseas to talk with relatives at home.

The *Realistic Phone Patch* is priced at $19.95 and comes complete with 15-foot telephone leads, three-foot transmitter lead and installation instructions. It features a built-in VU meter, gain control and locking push-to-talk bar. Not for use with transceivers employing electronic switching.

Realistic products are available at more than 1650 Radio Shack and Allied Radio Stores in all 50 states and Canada, and through Radio Shack Authorized Sales Centers, nationwide. For more information, use *check-off* on page 110.

rotary log-periodic antenna

Recently, KLM Electronics began manufacturing and marketing a line of antennas designed by Oliver Swan. This line includes a variety of high-performance vhf antennas that cover the spectrum from 50 to 520 MHz. The most recent addition to this line is a rotary log-periodic antenna that covers from 13 through 30 MHz. This antenna covers not only the 10-, 15- and 20-meter amateur bands, but MARS frequencies, the shortwave broadcast band and CB. Performance is equivalent to a three-element Yagi on any frequency between 13 and 30 MHz.

The KLM 13-30 log periodic uses seven elements and provides 9.2 dB over an isotropic. Front to back ratio is 12 dB. The input impedance is 50 ohms (a 4-kW balun is supplied), and maximum vswr at any point between 13 and 30 MHz is 2.0:1. The antenna weighs 76 pounds and has a boom length of 29.5 feet. The antenna is priced at $289.00, FOB San Jose, California. For more information, write to KLM Electronics, 1600 Decker Avenue, San Martin, California 95046, or use *check-off* on page 110.

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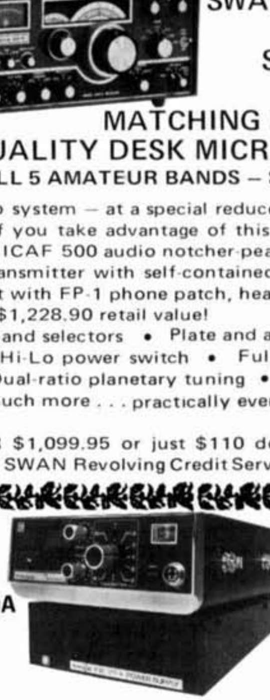

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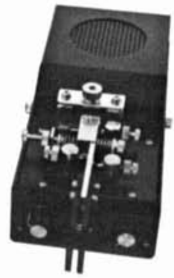

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- Crystal controlled time base (can be field calibrated)
- Convenient 3-position range select switch allows:
  1. Readout always in MHZ.
  2. Eight digit resolution by range selection
  3. Direct reading pre-scalar operation to 999.999 MHZ.
- "Battery save" switch for spot checks
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- Dimension 6" x 3.5" x 2.3"
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- Can be operated on internal or external power, with trickle charge and full charge positions
- Sample control lets operator determine how often the readout is updated. Can "hold" present count without being updated

Mail orders directly to:
Great American Miniatures, Inc.
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Midwest City, Okla. 73110

Model C-65 Freq. counter	$199.00
Battery charger	8.00
Battery charger & eliminator	18.00

OLD RELIABLE *OLD* *OLD* RELIABLE

$425.00 RX 1 ROTATOR $425.00

10 YEARS OF PROVEN SERVICE. THE HEAVY DUTY ROTATOR THAT WILL TURN ANY BEAM ARRAY YOU WANT TO INSTALL. CONTROL BOX TO MATCH S LINE. 115VAC SELSYN IND. 380 DEGREES ROTATION. LIMIT SWITCH UNIT IS 9½" DIA. 28" IN HEIGHT.

Designed Built Backed

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REEDLESS ENCODER

- SUB-AUDIBLE
- -30 TO 65°C OPERATION
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- R F SHIELDED
- OPERATES ON 9 TO 18 VDC
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AVAILABLE FOR ANY EIA CODE 67.0 THRU 192.8 Hz

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8MC. XTALS–8333–9000.
Silk Screened Panel.
18 Watts Output.

Price 49.95

Net to Amateurs
Complete with Tubes
Power Supply $9.95

SIX METER TRANSMITTER

for
- MOBILE
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TUBE COMPLIMENT
6U8 Oscillator Multiplier
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2E26 Final Amplifier
6BQ5 Modulator

ERL

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MANUFACTURERS OF ELECTRONIC DEVICES
224-15 Linden Blvd. Cambria Heights, N. Y. 11411

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two calculating ideas

Texas Instruments TI-2500

Fast-figuring. Adds, subtracts, multiplies, divides . . . does true credit balances, mixed and chain calculations.

8-Digit Display with a floating decimal, no need to ever set it. Constant switch allows shortcut multiplication and division. Everything Included. AC adapter/charger, rechargeable batteries and handy carrying case. Only 12 oz. Measures 5½" x 3" x 1-7/10".

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The electronic slide rule

Made to order for technicians, engineers and scientists, this amazing little machine not only adds, subtracts, multiplies and divides, but also figures reciprocals, squares and square roots. It also incorporates capability for sign change and scientific notation.

Display includes 8 digits plus 2 digit exponent and 2 signs for 12 total characters.

Compact size 6.3" x 3.1" x 1.5". Includes AC adapter/charger, nicad batteries, carrying case and owners manual.

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Here are the wonder products of the 70's — two exciting Texas Instruments' handheld portable calculators all ready to go to work for you at any time and any place. Two models to fit virtually every need, both made in the USA by Texas Instruments and backed with their warranty and service organization.

Order today. Send either a check or money order or all of the information from your MasterCharge or BankAmericard and we'll send your new calculator by return mail.

comtec

Greenville, New Hampshire 03048

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YAESU

YOUR ASSURANCE OF PERFORMANCE & QUALITY

Amateur Price Net

F-LINE MODEL 400D/400SD

Includes 6 & 2 Meters

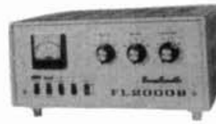
FR DX 400 $299.00
FR DX 400 SD $399.00

SP-400P $59.00
SP-400S $19.00

FLDX 400
Exciter $339.00

FL 2000B
1200 Watt Linear Amp.
$339.00



FR-DX-400 RECEIVER

Mode of Operation:	SSB, CW and AM (FM)
Frequency Range:	1.7-2.3Mc, 3.5-4.1Mc, 6.9-7.4Mc, 13.9-14.4Mc, 20.9-21.4Mc, (26.9-27.5Mc), 27.9-28.5Mc, 28.5-29.1Mc, 28.9-29.5Mc, (29.5-30.1Mc), (9.9-10.5Mc), (50-52Mc or 52-54Mc), (144-146Mc or 146-148Mc).
Sensitivity:	SSB/CW; 0.5uV at S/S+N 10 db. AM; 1uV at S/S+N 10 db.
Selectivity:	SSB/CW/AM; 2.4kc/6 db 4 kc/25 db. AM; 4 kc/6 db 7.5kc/60db. (CW; 600 cps/6db 1.5kc/60db). (FM; 24kc/6db).
Spurious Response:	Better than -60 db at 14Mc
Frequency Stability:	After warm up less than 100 cps. per any 15 min. or 10% line voltage, fluctuation.
T-notch Attenuation:	-50db
Antenna Impedance:	50-75 ohms
Audio Output Impedance:	4 or 600 ohms
Output:	1 watts @ 5% distortion
Power Requirement:	100/110/117/200/220, or 234 volts AC, 50 or 60 cps., approx. 50 watts
Dial Calibration:	50 kc main dial division, 1 kc reading
Calibration:	100kc or 25kc
Dimensions:	14 1/2" W, 6 1/4" H, 11 1/5" D.
Weight:	Approx. 24 lbs.

( ) option

SP400-P

SP400-P: Hand Liner Phone Patch, speaker is designed for the FT dx 400 and 401 series, single side band trans ceivers. Front panel: Patch switch, off and on switch, (meter level to phone line). TX and RX gain controls. Rear apron: Receiver 8/4 ohm jack, 600 ohm receiv er jack, monitor null switch, balance control, line jack trans-mitter 600 ohm jack trans mitter Hz jack.

FL-DX-400 TRANSMITTER

Frequency Coverage:	3.5-4.1Mc, 6.9-7.5Mc, 13.9-14.5Mc, 20.9-21.5Mc, (27.9-28.5Mc), 28.5-29.1Mc, (28.9-29.5Mc).
Modes of operation:	SSB; Upper and lower side-band on all bands. CW; Grid block keying, VOX circuit keying. AM; Either side-band with carrier.
Dial Calibration:	Main dial calibrated 0 to 500kc and 500 to 1000kc. Vernia dial calibrated 0 to 50kc and 50 to 100kc in 1kc division.
Stability:	Less than 100 cycles within any 15 minutes after warm-up, less than 100 cycles with 10% change in line voltage.
Sideband Suppression:	50 db at 1000 cps.
Carrier Suppression:	Better than 50 db.
Distortion Products:	In excess of 30 db down.
Frequency Response:	300 to 2700 cps.
Input Power:	SSB and CW-240 Watts PEP AM-100 Watts.
Output Impedance:	Nominal 52 ohms adjustable with pi network.
Microphone:	High impedance dynamic or crystal.
Power Requirements:	100/110/117/200/220 or 234V, 50/60 cps AC.
Dimensions:	14 1/2" W, 6 1/4" H, 11 1/5" D.
Weight:	Approx. 25 lbs.

FL-2000B LINEAR AMP

Circuit: Grounded Grid
Frequency: 80 to 10 meters
Max. Input: 1000 watts DC
Plate Voltage: 2400 volt DC
Power :115/230 volt AC,
Requirement 50/60 cps.
Input Impedance: Approx. 60 ohms
Output Impedance: 50 to 100 ohms
Cooling: Forced air cooling
Tubes: 572B; 2 in parallel
Dimensions: 14 1/2" W, 6 1/4" H, 11 1/2" D.
Weight: Approx. 40 lbs.

Here is the opportunity you have been waiting for:

- *Split frequency operation for working that choice DX and Net Operation.*
- *This is the ultimate for complete coverage with control from either the transmitter or the receiver.*
- *2 Mechanical Filters for excellent selectivity plus T notch rejection tuning.*
- *Solid State VFO for excellent stability.*
- *Clarifier for Net operation.*
- *Built in Monitor for listening to transmitted signal.*
- *100KC and 25KC calibrator -- 3 step AQC.*
- *Pick up a brochure at your Dealers and ask for Demonstration.*

FACTORY SERVICE IS AVAILABLE WHERE WARRANTY HAS EXPIRED. MODIFICATION KIT MK-160 (160M) $16.00 WITH XTAL PLUS LABOR. MODERNIZATION KIT MIR-1 FOR OLDER MODEL FT101 $40.00 PLUS LABOR. BROCHURES WITH SPECIFICATIONS ARE AVAILABLE AT YOUR DEALERS LISTED BELOW:

HENRY RADIO STORES/213-477-6701
Los Angeles, Anaheim, CAL.; Butler, Mo.
HAM RADIO OUTLET/415-342-5757
Burlingame, Cal.
RACOM ELECTRONICS/206-255-6656
Renton, Wash.

WILSON ELECTRONICS/702-457-3596
Pittman, Nev.
ED JUGE ELECTRONICS/817-926-5221
Fort Worth, Tex.
AMATEUR ELECTRONICS SUPPLY/414-442-4200
Milwaukee, Wis.; Cleveland, Ohio

FRECK RADIO & SUPPLY/704-254-9551
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HARRISON RADIO/516-293-7990
Farmingdale, L.I. Valley Stream, L.I. New York City, N.Y.

YAESU MUSEN USA INC.

7625 E. ROSECRANS AVE. UNIT 29 PARAMOUNT, CAL. 90723

TWX 910-346-7624
TEL 213-633-4007

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ELECTRONICS COMPANY

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New 2-Meter, 12-Channel VHF/FM Mobile Transceiver

Midland, for years one of the top names in communications equipment, proudly introduces this sensational new 15-watt Amateur mobile. More power than most sets, it drives linear amplifiers to full output. And there's a low-power position switch for short-range 1-watt output. Instantaneous final protection circuit prevents damage from excessive VSWR. Receiver has multiple F.E.T. front end with high "Q" resonator filter, ceramic filters in I.F.—superb selectivity, sensitivity, and bandpass characteristics. King-size illuminated S/RF meter, channel selector. Variable squelch, volume controls. ADL circuit maintains deviation level without distortion. Crystals supplied for .16-.76, .34-.94, *and .94 simplex.* Each receive-and-transmit crystal has individual frequency trimmers. Equipped with connector for attaching tone burst and discriminator meter accessories. Includes dynamic mike and mounting hardware.
Model 13-500

For information on Midland Amateur radio, write for our full-line brochure: P.O. Box 19032, Kansas City, Missouri 64141

MIDLAND®
ELECTRONICS COMPANY
"Coming on Strong in Amateur Radio"

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What has feathers, a lot of class, and loves hams?

Why, the FLAMINGO of course.

SAROC

The Fun Convention

hosted by **SOUTHERN NEVADA AMATEUR RADIO CLUB, INC.** at

FLAMINGO HOTEL CONVENTION CENTER, LAS VEGAS, NEV. 89109 — JANUARY 3-6, 1974

ADVANCE REGISTRATION — $10.50 per person includes:

1. Special room rate of $15.00 plus room tax per night single or double occupancy, effective January 3 through 10, 1974 while 500 rooms last at the Flamingo Hotel.
2. Advance Registration drawing ticket for Saturday.
3. Regular Registration drawing ticket for Saturday.
4. Complimentary cocktail at the Flamingo Hotel.
5. Complimentary KENO ticket at the Flamingo Hotel.
6. Admission to technical seminars, meetings and exhibit area, Friday and Saturday.
7. Ticket for admission to cocktail party hosted by SAROC and HAM RADIO MAGAZINE, Friday.
8. Ladies who register will receive admission ticket for their program on Saturday.
9. Ticket for admission to cocktail party hosted by SAROC and SWAN ELECTRONICS, Saturday.
10. Ticket for Flamingo Hotel Buffet Hunt Breakfast with Champagne, Sunday.
11. Tax and Gratuity on all items 1 through 10 except hotel accommodations.

ADVANCE REGISTRATION with midnight show — $17.50 per person:

Includes all items 1 through 11, plus Flamingo Hotel Midnight Show and two drinks, Sandler and Young are scheduled in the Flamingo Hotel Main Show Room.

ADVANCE REGISTRATION with dinner show — $21.50 per person:

Includes all items 1 through 11, plus Flamingo Hotel Dinner Show (Entree: Brisket of Beef) no cocktails, Sandler and Young are scheduled in the main show room.

Advance registration must be received in SAROC, P. O. Box 73, Boulder City, NV 89005 on or before December 15, 1973. Refunds will be made if request in writing received in P. O. Box 73 on or before January 3, 1974.

SEVENTH NATIONAL FM conference Friday and Saturday, FM Hospitality Room 16/76, 28/88 and 34/94 repeaters. WCARS-7255 and WPSS 3952 special events stations to assist mobile operators.

Mail accommodations request to Flamingo Hotel, Las Vega, Nevada 89109 — Do it now!

Mail Advance Registration fee to, SAROC, P. O. Box 73, Boulder City, NV 89005 — Before December 15

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COMPUTER KEYBOARD $7.00 (as is)

Several styles on hand in poor condition, broken key/keys, broken case or no case, etc. Still a good value at $7.00 for parts, switches, and each has encoder board in base.

2N2152	45 volt	170 watt	PNP-G	$1.00
*2N3713	80	150	NPN-S	1.00
*2N3773	160	150	NPN-S	.75
*2N3789	60	150	PNP-S	.75
2N5301	40	200	NPN-S	1.25
*2N5301	40	200	NPN-S	1.00

*Removed from used equipment

TRANSFORMERS

BRAND NEW, 115 volt AC input. OP - AMP XFMR, out puts: 16 VCT ½ amp, 17 VCT ½ amp. $3.50

FILAMENT or BTRY CHARGER XFMR

output of 18 volts at 4.5 amp $3.50

313,344 CORE MEMORY $125.00

From SPECTRA computer, visually OK. 64 x 68 x 4 x 18 core stack. Figures out to 35K Byte.

LED 7 SEGMENT READOUT

Similar to MAN-1. Factory seconds but functionally OK. Fit 14 pin DIP socket.

7 segment w/left decimal #LED-A-L $3.00
7 segment w/right decimal #LED-A-R 3.25
7 segment no decimal #LED-A 2.75
Socket for above, gold plated leads 3/1.00

IC SALE YOUR CHOICE 3 for $1.00

μl 900 BUFFER TO-5
μl 914 DUAL 2 INPUT GATE TO-5
μl 923 JK FLIP FLOP TO-5
μl 926 Hi speed JK FLIP FLOP TO-5
μl 931 JK/RS FLIP FLOP (DIP)
10 pin socket for TO-5 IC 3/1.00

GIANT NIXIE B7971

Used $1.00 Brand New $2.00
With schematic for GIANT clock.

COMPUTER TAPE DECK $75.00

Takes ½ inch tape, made by Computer Entry Systems. Visually ok, with electronics, no data available.

LASER DIODES, new listing just arrived, send SAE.

Postage Extra on all Items

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Racom has the following Yaesu models in stock for immediate delivery.

FT-101, FTDX-570 & 401, External VFO's, Speaker-Patches Frequency Counters.

Also 2 Meter FM Mobile Transceivers FT-2FB and FT-2 Auto

Write to Woody, W7RC

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COMPLETE REPAIR SERVICE

WE PAY HIGHEST PRICES FOR ELECTRON TUBES AND SEMICONDUCTORS

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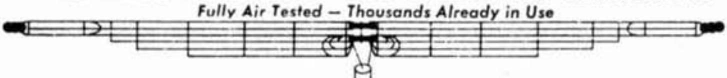
NOW IS BACK

EXCLUSIVE 66 FOOT 75 THRU 10 METER DIPOLE

NO TRAPS — NO COILS — NO STUBS — NO CAPACITORS

Fully Air Tested — Thousands Already in Use



#16 40% Copper Weld wire annealed so it handles like soft Copper wire—Rated for better than full legal power AM/CW or SSB-Coaxial or Balanced 50 to 75 ohm feed line—VSWR under 1.5 to 1 at most heights—Stainless Steel hardware—Drop Proof Insulators—Terrific Performance—No coils or traps to break down or change under weather conditions—Completely Assembled ready to put up—Guaranteed 1 year—ONE DESIGN DOES IT ALL; 75-10HD—ONLY $12.00 A BAND!

Model 75-10HD . . . $60.00 . . . 66 Ft . . . 75 Thru 10 Meters
Model 75-20HD . . . $50.00 . . . 66 Ft . . . 75 Thru 20 Meters
Model 75-40HD . . . $40.00 . . . 66 Ft . . . 75 Thru 40 Meters
Model 40-20HD . . . $33.00 . . . 35 Ft . . . 40 Thru 20 Meters
Model 80-40HD . . . $42.00 . . . 69 Ft . 80-40-15 Meter (CW)

ORDER DIRECT OR WRITE FOR FULL INFORMATION

MORGAIN 300H Shawnee Leavenworth, Kansas 66048

OR THRU YOUR FAVORITE DISTRIBUTOR

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If you like 2 METER . . .

YOU'LL LOVE OUR

15 OR 1 WATT POWER OUT/SWITCH SELECTABLE/
FULL 12 CHANNEL TRANSMIT AND RECEIVE CAPABILITY

You'll like the crystal clear transmit and receive performance of this compact, 2 meter unit and so will those listening. The 12 transmit channels are provided with individual trimmer capacitors for the optimum in point-to-point and repeater applications. A HI/LO power switch provides 1 watt output or full rated output. The receiver has an audio output of 3 watts at excellent sensitivity. Solid state, American made quality at a low price.

$229.00 AMATEUR NET

includes plug-in ceramic mike, mounting bracket and transmit and receive crystals for 146.94 MHz.

Regency
the first name in solid state

THE FM LEADER IN 2 METER AND 6 METER . . . AND NOW 220 MHz

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MISC.

Scanamatic HighBand 8 channel scanner Write
Kenwood Twins R-599 & T-599. Receiver and transmitter. Regular net $818 for both. Excellent, guaranteed good as new $618.00
Henry 2K-4 Floor Model, ... mint, Write
U.S. Callbooks .... $ 8.95
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Gonset 903A 2 meter amplifier, 5 watts in, 500 watts out all modes .... new condx., $375.00
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Millen magnetic shields for 3" C.R. scope tubes with brackets .... Brand new $ 6.95
TP-9 Similar to EE-8 field telephone with built-in amplifier. Up to 50 mi. Less easily obtained batteries .... Brand New $ 75.00
115 Volt AC Power Supply for BC-221, slips into battery compartment. Also can be used with LM freq. meters .... $ 29.50
ALUMINUM DIE CAST BOXES in many different sizes. Dosy boxes. Details in New Green Sheet No. 23. Write for latest prices.
2 METER VHF DUMMY LOAD/WATTMETER Good up to 15 watts — w/SO-239 CONNECTOR and 3" METER .... $ 15.95

Jackson G80 brass 50:1 anti-back-lash dial drive ¼" to ¼". $19.00 value .... $ 5.50

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2 Meter Transceiver

10 WATTS OUTPUT PLUS
TREMENDOUS RECEIVER
22 CHANNELS

Ready to go with crystals for 5 frequencies at no extra charge

$289.00

Extra Crystals $6.95 each

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$89.00 when purchased with IC-22

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Little Dipper, GDO .... $ 94.50
Dummy Load - Wattmeters - 52Ω
333 DC-300 MHz, 5, 50, 125 or 250 watts int. .... $ 79.95
334A DC-300 MHz, 1000 watts $139.95
374 DC-300 MHz, 1500 watts int. $169.95
850A, 852 Inductors .... $ 59.95
851 Inductor .... $ 29.95
425 Low Pass Filter, 10-80 meters .... $ 24.95
210 Audio Osc., ideal for lab & broadcast $329.95
410 Distortion Meter, ideal for lab & broadcast $369.95
AM-141 Amplifier, 2000 watts RF output, continuous 2-18 MHz, complete with coils and 833-A's. Built-in 115 VAC Supply, unused. $950.00

STANDARD

SRC-146A 2 meter handheld transceiver $289.00

SBE

SBE-450 TRC, use with 10 watt, 2 meter transceiver to operate on 450 .... $195.00

LITTLE LULU

6 Meter AM Transmitter with VFO
12 VDC/115 VAC Power Supply
Available factory wired or as parts
Write for details

INVERTER/CONVERTER:

INVERTER, 12 volt DC input, 115 volt AC out, Model 12-115 solid state power supply, 200 watts continuous .... new, $ 59.95

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610-2 Deluxe Economy 2000 ohm headset with cushions .... $ 9.95
EN-5 Stereo Headphones .... $ 9.95

Tube Headquarters. Diversified Stock. Heavy inventory of Eimac tubes, chimneys, sockets, etc.
572B .... $ 17.50
Barry Now Stocks Bogen, Electrovoice & University. Call or Write.

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C.D. HAM "M" ROTATORS, new complete $99.95
CD Ham-M for 220 VAC in stock $175.00
C.D. TR-44 ROTATORS, new (complete) $ 63.95
CABLE for Ham-M & TR-44 .... @ 14¢/ft.
CD AR-22R complete rotator for small beams $ 33.95
BN86 Balun by HyGain .... $ 14.95
RG-8A/U 100 ft. rolls. VHF connector PL-259 one end Type "N" (UG-21E/U) other end $ 12.50
RG8A/U — 65 feet with PL-259 connectors on each end .... $ 9.50
Times Wire & Cable, T-4-50 (FM-8) 50 ohm lowest loss type RG-8 cable .... 25¢/ft.
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BNC to RG-17 adapter UG-167C/U .... $ 7.95
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Authorized factory dealers for Antenna Specialists, CushCraft, Gam, Heights Towers, Hy Gain, Mor-Gain Antenna, Mosley, Newtronics, Tri-Ex, Rohn, E-Z Way, Times Wire, Telrex.
Cush-Craft Trick Stick, universal dipole, 10-2 meters, 1.5 dB gain at 146 MHz .... $ 8.95
English deluxe balun, low power .... $ 9.95
RINGO AR-2 3.75 dB gain, 135-175 MHz $ 12.50
BBLT-144 Trunk Lip, 3.75 dB gain .... $ 34.95
Newtronics CGT-144 mobile 5.2 dB gain $37.95
Quick Disconnect by Newtronics for CGT, etc. $ 10.95
CG-1 Gutter Clip by Newtronics .... $ 1.25
2M MAGNETIC MOUNT w/RG58 & PL259 with 10 ft. RG 58 ready to go .... $ 9.95
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18AVT/WB VERTICAL .... $ 69.95
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Send 50¢ postage & handling (refund 1st order).

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See Page 112 for MORE from Barry

flea market

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COPY No special layout or arrangements available. Material should be typewritten or clearly printed and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** can not check out each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue. Deadline is 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

TELETYPEWRITERS — Kleinschmidt — portable, fixed, sets, punches, parts, reconditioned, reasonable. Mark/Space Systems, 3563 Conquista, Long Beach, Calif. 90808. 213-429-5821.

THE NORTHWEST GEORGIA Amateur Radio Club and Repeater Association, is having its annual hamfest on Sunday, October 7, 1973 at the Coosa Valley Fair Grounds in Rome, Georgia. Everyone is invited to attend. Gates open at 9:00 a.m.

ACTIVE HAMS — Monthly mailer of reconditioned and new equipment specials. Sell — Buy — Trade. Write: Associated Radio, 8012 Conser, Overland Park, Kansas 66204.

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

STILL HAVE A FEW LEFT! We have a few of the Brand New deluxe equipment enclosures left. See our ad on Page 89 of the July 1973 Ham Radio.

DX'ers — New Logarithmic Speech Processor. Nominal 8 dB increase in average power. Less than 5% distortion @ 1kHz. L/C filter. HI-Z Meter. $49.95. Also, low noise dual gate MOSFET receiver preamplifier. Nominal 20 dB gain. 10-30 MHz. $39.95. With cabinets. Dynacomm, 1183 Wall Road, Webster, N. Y. 14580.

WE BUY ELECTRON TUBES, diodes, transistors, integrated circuits, Semiconductors. Astral Electronics, 150 Miller Street, Elizabeth, New Jersey 07207, (201) 354-2420.

WESTERN UNION DESK-FAX Telefax Transceivers: Several extra machines (checked out), $14 each, shipping collect. Bill Johnston, 1808 Pomona Drive, Las Cruces, New Mexico 88001.

WORLD QSL — See ad page 105.

EMBROIDERED EMBLEMS AND PATCHES. Custom made from your design. 10 to 1000's. Write Russell, 1109 Turner St., Augusta, Maine 04210.

MULTI ELMAC PKG. (HF68, PMR8A, PS1070), $130.00; Heath 6M SSB pkg. (SB110A, PS23A, PS13A, SB600), $325.00; Heath Spectrum Analyzer SB620, $100.00; Hammarlund HQ180AX, $300.00; Hy Gain 6M 6 el. beam, $30.00. All equipment even the beam, absolutely mint, with manuals and cables. You pay shipping or will deliver within 150 miles. Prices not firm, make offer. Dan Roe, W8JSR, 24644 Madison Ct., Apt. #227, Farmington, MI., 48024. Phone 1-313-477-4846.

WANTED, RD 92/UX; fax recorder. Need several mechanical drum subassemblies. Don W1GBO, Box 803, North Falmouth, Mass. 02556.

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

USED MYLAR TAPES — 1800 foot. Ten for $8.50 postpaid. Fremerman, 4041 Central, Kansas City, Mo. 64111.

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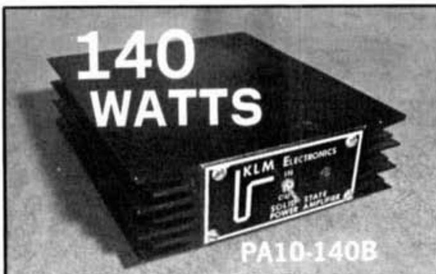

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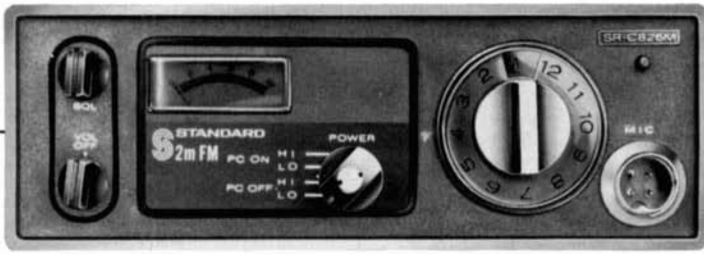

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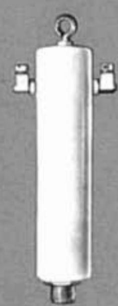

The famous sealed helium filled Balun . . . employed with the DGA Series Antenna Systems. Solderless center insulator and easily handles more than full legal power while reducing unwanted coax radiation. Equipped with a special SO-239 type coax connector and available either 1:1 or 4:1.

MODEL DGA-2000-B . . . $12.95
Postpaid in U.S.A.

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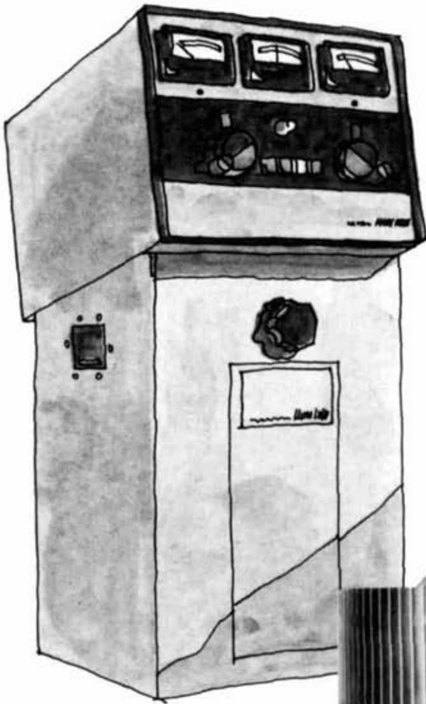
EIMAC's rugged 8877 powers Henry Radio's reliable 4K Ultra linear commercial amplifier.



When Henry Radio set out to design a sophisticated linear amplifier for high reliability communications in the high frequency range, they chose EIMAC's high-mu 8877 power triode.

The result is the new 4K Ultra, which operates with 4000 watts input in heavy-duty commercial service. The 4K employs continuous variable capacitive and inductive elements tunable over the 3.0 to 30 MHz range. Optimum input and load conditions are provided for a wide variation in antenna systems. EIMAC's 8877 enables the 4K to deliver over 2500 watts of continuous SSB or CW output with only 50 to 75 watts of drive. For RTTY the 4K will provide about 2000 watts of continuous duty output.

The 8877 is a ceramic-metal triode that delivers a lot of power and linearity in a package only three and one-half inches high. At 30 MHz, typical power gain is 15 dB. This impressive gain is achieved with 3rd order intermodulation products −38 dB below one tone of a two equal-tone drive signal.

With Henry Radio, you know quality counts. And they know you can't do better than EIMAC. For full specifications on the 8877, write to EIMAC Division of Varian, 301 Industrial Way, San Carlos, California 94070. Or contact one of the more than 30 Varian/EIMAC Electron Tube and Device Group Sales Offices throughout the world.

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