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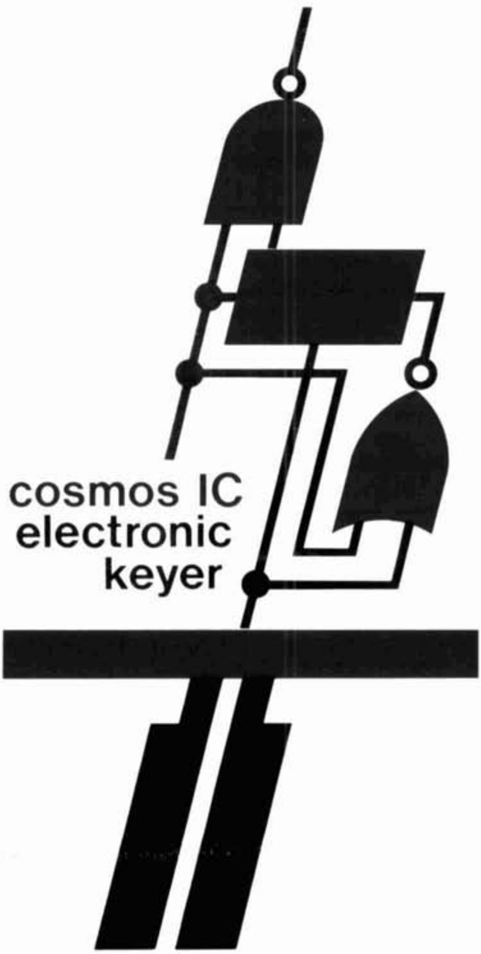

cosmos IC electronic keyer

this month

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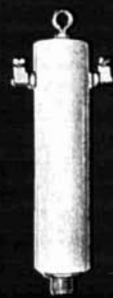

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a second look

by jim fisk

Solid-state imaging devices which may eventually find their way into high-quality television cameras are now becoming available for lower-resolution applications such as slow-scan optical character recognition and basic-recognition security systems. Some of these devices, which consist of large arrays of charge-coupled photo-sensitive semiconductors that are scanned with digital circuitry, are so sensitive they can detect objects illuminated only by candle-light. The big problem has been to obtain the 525-line video resolution required for television cameras. However, Bell Labs has built a solid-state vidicon capable of full *Picturephone* resolution (250 by 225 lines) and Fairchild has a 100-by-100 element imager that is suitable for many industrial applications.

Actually there are two different types of solid-state imaging devices: the charge-coupled device or CCD and the charge-injected device or CID. Although the image signal is developed and stored in both devices in much the same way, the signal is retrieved differently. In the CCD array the image is scanned sequentially while in the CID array the image is scanned in an x-y fashion. This permits the CID array to be scanned at any speed or individual points to be scanned randomly. Another advantage of the CID design is that single-image defects cause the loss of only one image point. In a CCD array a single-image defect blanks out the rest of that line.

ATV test pattern from W8DMR is generated by solid-state charge-coupled image sensor.

Bill Parker, W8DMR, is believed to be the first amateur to use a solid-state image sensor to transmit an amateur television test pattern. Bill, who has been on amateur television since 1949 and is still active, uses the 10,000-element Fairchild CCD-201 solid-state imager mentioned earlier. This sensor is mounted in a 24-lead dual-in-line package with an optical glass window. The 1.2- by 0.8-mil (0.03- by 0.02-mm) image-sensing elements are located on 1.2-mil (0.03-mm) vertical centers and 1.6-mil (0.04) horizontal centers. This provides an image aspect ratio of 4 by 3. In addition to the image-sensing chips the CCD-201 includes 100 columns of two-phase shift registers interdigitated among the photo-sensitive elements, a 102-unit two-phase analog output shift register, an output preamplifier and a compensation amplifier. A photograph of the transmitted test pattern is shown to the left. When 512-by-320 element arrays become commercially available the resolution wedges on the test pattern will be much more distinct. Who will be the first to apply this new technology to real-time slow-scan TV?

Jim Fisk, W1DTY
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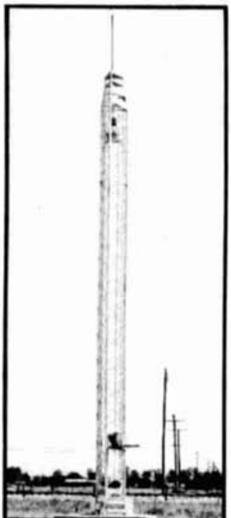

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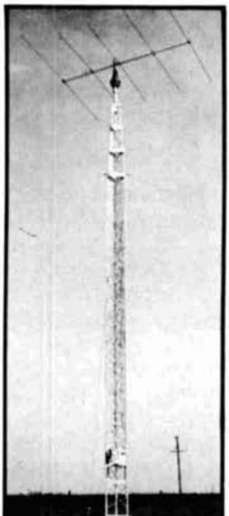

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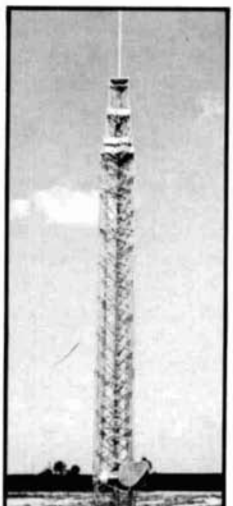

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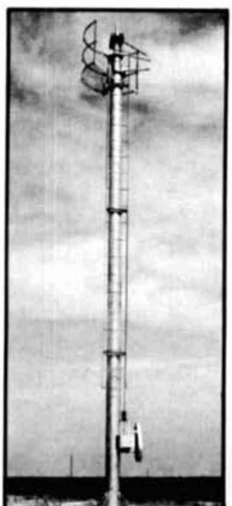

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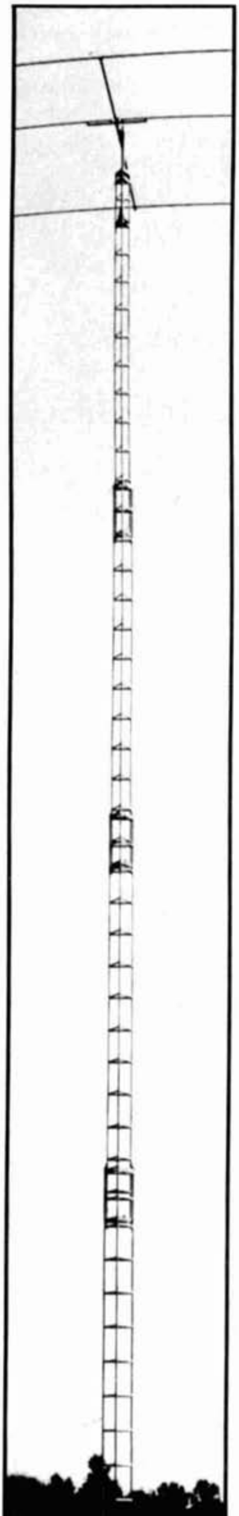

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cosmos IC electronic keyer

Applying cosmos technology to a versatile, compact, low-drain keyer design

James W. Pollock, WB2DFA, 6 Terrace Avenue, New Egypt, New Jersey

Since the advent of ICs on the amateur radio scene, a number of articles on keyers have appeared in the amateur magazines. Now that c-mos or cosmos ICs have become available to the amateur from various sources at modest prices,* it's time that these "state of the art" building blocks found an application in a homebrew project.

The cosmos keyer described here draws only 0.4 mA on standby, with an average key-down current drain under 2 mA at a supply voltage of 10 volts. The keyer can work properly with supply voltages from 4 to 15 volts. When operated at 5 volts, the keyer actually consumes less than 100 microamps, less power than a set of headphones! Its low power requirement makes it ideal for the QRP or field day enthusiast. If TTL logic was substituted for the cosmos ICs in this keyer, the current drain would be in excess of 200 mA.

Some of the more important features of an electronic keyer using cosmos ICs is low power consumption, simple construction and modest construction cost (about $10.00 for parts). The keyer described in this article features self-completing dots, dashes and spaces, a built-in transmitter keying circuit and sidetone generator and small size – the circuit board can be mounted inside your favorite rig.

*Poly Paks, Inc., Box 942, South Lynnfield, Massachusetts 01940.

The basic clock gating circuit used in the keyer is illustrated in **fig. 1**. The clock gate allows only full clock pulses to pass through it, regardless of the timing inaccuracies of the *enable* signal. No partial pulses or "slivers" can be tolerated if a keyer is to send perfect code.

In **fig. 1**, U1 is a type-D flip-flop. The logic level present at the *data* or D input will be transfered to output Q at the next positive going edge at the *clock* input (pin C). Notice that the clock feeds pin 1 of U2, pin 1 of U3 and pin C of U1. U2 is the output gate and U3 is the reset gate.

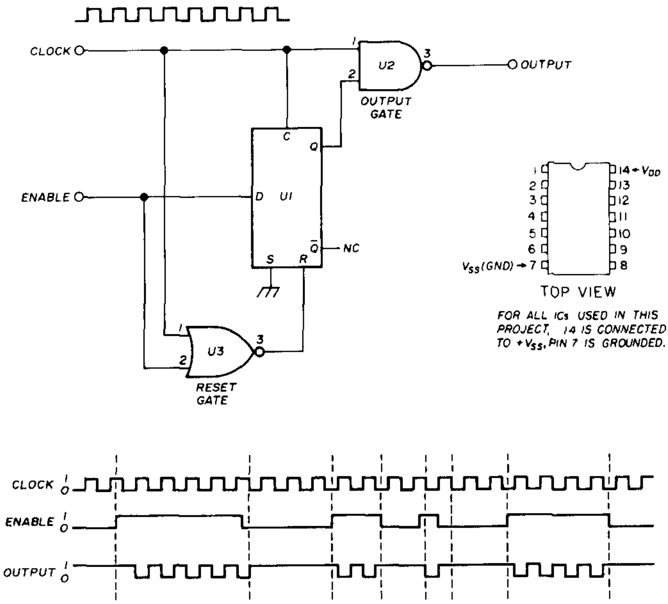


fig. 1. Basic clock gating circuit for the cosmos keyer.

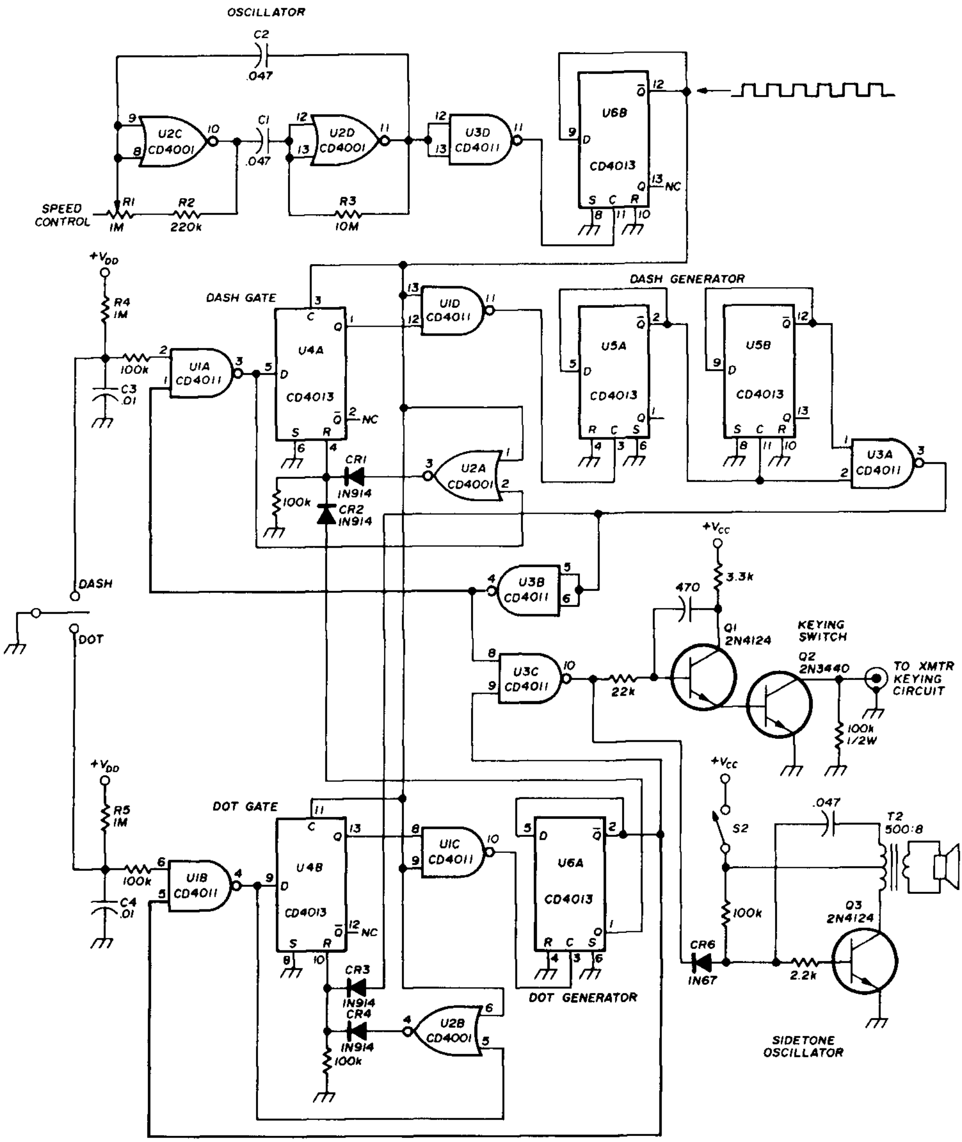
When the *enable* signal is low (zero), pin Q of U1 is assumed to be zero if pin C is being clocked. As long as pin Q of U1 is zero, U2 will not pass the clock pulses. Also, as pin D of U1 is zero, the clock pulses at pin 1 of U3 will pulse the reset pin (pin R) of U1. A 1 on the reset pin of a cosmos flip-flop will force Q to zero and will override all other input signals. The *set* pin of U1 is grounded for this reason. A 1 at the *set* pin will force Q to a 1 and will override all other input signals.

When the *enable* signal goes to a 1, pin Q will go to a 1 at the next positive-going clock pulse leading edge. Now that pins 1 and 2 of U2 are 1, pin 3 goes to zero and will go to a 1 when the negative-going clock transition occurs. Thus the output of U2 is the inverted version of the clock pulses.

When the *enable* goes to zero, flip-flop U1 will be reset by the NOR gate U3 at the time when the clock is zero. If the NOR gate were not used to reset U1, then the next positive-going edge of the clock pulse would appear at the pin 1 of U1 simultaneously with the negative transition at pin 2 of U2 due to the data transfer action of flip-flop U1. Thus you would have logic levels changing simultaneously in opposite directions at pins 1 and 2 of U2. This would result in a "sliver pulse" at pin 3 of U2 when the input signals cross the threshold level of U2. *To prevent this from* happening, the reset gate U3 is used to reset the flip-flop when both the *enable* and *clock* inputs are at zero.

final circuit

The circuit I finally settled on for my cosmos keyer is shown in **fig. 2**. The keyer generates dashes and dots at a fixed time ratio of 3 to 1. A space has the same duration as a dot. The time base of the keyer is generated by two NOR gates, U2C and U2D, connected in a class-A multivibrator configuration. Resistor R3 causes U2D to self bias into a class-A condition, with the output reaching a dc level equal to the threshold of the gate itself (about 45 percent of the supply voltage). Resistors R1 and R2 have the same effect on U2C. The time constant (R3-C1) is much greater than (R1 + R2)C2 so the frequency of the oscillator

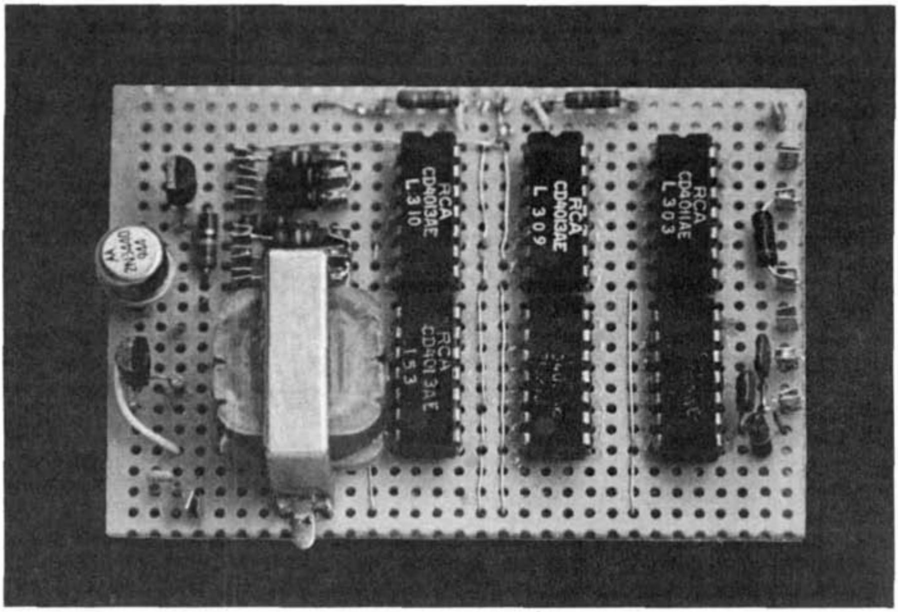


R1	1 megohm linear taper potentiometer	U2	cosmos NOR gate (CD4001 or equivalent
T2	500 ohm CT to 8 ohm transformer (Radio Shack 273-1381)		
U1, U3	cosmos NAND gate (CD4011 or equivalent)	U4, U5 U6	cosmos dual D flip-flop (CD4013 or equivalent)

fig. 2. Complete schematic diagram for the cosmos keyer. The bold lines are the feedback paths for the dot and dash generators, which allow them to end their timing in sync with the clock after paddle release. The sidetone generator, lower right, is optional.

is inversely linear with the setting of R1. Inverter U3D buffers the oscillator and squares up its output. Flip-flop U6B divides the oscillator frequency by two, but more importantly, provides a clock source with a perfect 50 percent duty cycle.

Note that the dot and dash generators each have their own clock gates, and are connected in such a manner that which-



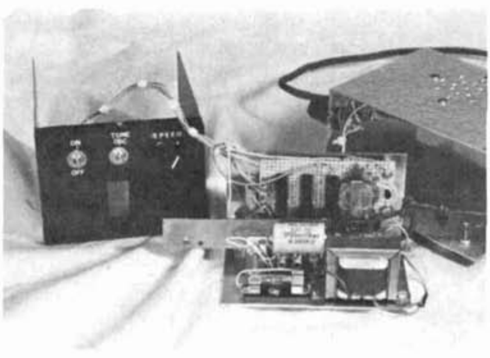
Logic circuitry for the cosmos keyer is wired on small section of perforated circuit board. Keying transistor and sidetone circuit are to the left.

ever of the two gates is enabled first overrides the other until its timing cycle is completed. If the dot gate is enabled first, the dash gate will be held in reset via diode CR2 until the timing cycle for the dot is completed, even if the dash paddle was depressed. Also, a complete space period would elapse before a dash could be sent, and vice versa. Diode CR3 allows the dash gate to reset and override the dot gate while dashes are being sent. The RC networks R4-C3 and R5-C4 provide pull-up bias for gates U1A and U1B and also eliminate the effect of key contact bounce.

The dash generator consists of a four-state binary counter (U5A and U5B) and a gate (U3A) to decode the four count states into dashes that are three clock periods long, separated by spaces one clock period long. U3B inverts the output of U3A to provide the proper logic levels to U3C by cancelling the inverting effect of U3A.

The bold lines are the feedback paths for the dot and dash generators. These feedback lines allow the dot and dash generators to end their timing in synchronization with the clock after their respective paddles have been released. This is what adds the self-completing feature to the keyer.

The output gate U3C drives the Darlington pair Q1 and Q2. Transistor Q2 is a high-voltage device, the Motorola 2N3440, which has a 250-volt B_{vcb} rating. If you plan to use the keyer solely



Construction of the cosmos keyer. Circuit board, keying mechanism and power supply are on center chassis. Speed control and switches are mounted on front panel, left.

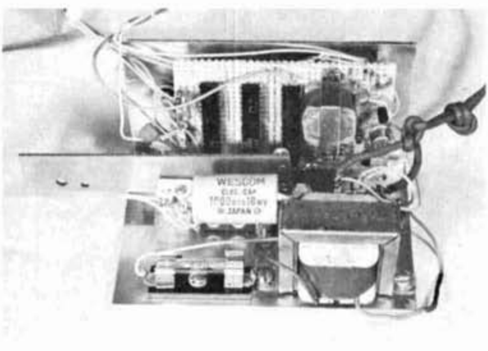
for your solid-state QRP rig, then just about any 5-watt npn transistor will do the job.

The largest portion of the standby current drain can be attributed to the oscillator. When the threshold region of a cosmos device is entered at a slow rate (such as by RC decay), the device draws relatively large amounts of current. The increase in current during key-down conditions is caused by the conduction of Q1 in driving Q2.

important

When working with cosmos ICs it's important to remember to *never* leave unused input pins floating. *Always* make sure that any unused input pin is tied to either ground or to V_{dd}, whichever is logically appropriate. For example, the *set* and *reset* pins of U5 and U6 have been tied to ground. If this precaution is not observed, these high impedance inputs are wide open for electrostatic charge pick up. Also, since the input capacitance of a cosmos device is typically 4 pF and the gate impedance is on the order of 10^{12} ohms, the result is a parasitic RC network with a time constant of 4 seconds. Any electrostatic charge can be stored for several seconds, injecting a false logic level into the cosmos device. That could raise havoc with your logic.

If you follow the schematic faithfully, all of the cosmos gates and flip-flops will be used up in fabrication of the keyer,



Main chassis layout. Cosmos circuitry is installed on perforated board, rear; power supply is in foreground.

with no surplus devices to cause problems. *Check* and *doublecheck* your wiring!

It's also advisable to provide overvoltage protection and regulation if operation from an unstable supply is anticipated. *Do not* exceed 15 volts, or

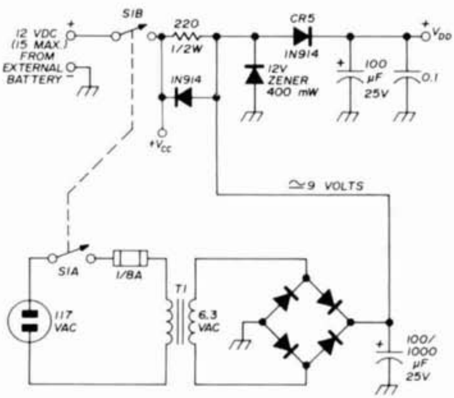


fig. 3. Suggested power supply for cosmos keyer, with inputs for either 117 Vac or low voltage dc. Transformer T1 is a small 6.3 Vac filament transformer.

the cosmos ICs will zener and draw excessive current which may destroy them. I recommend a maximum supply voltage of 12 volts. This should give you plenty of margin for error. Also, be sure to provide a means of protection from accidental polarity reversal of the power supply. Diode CR5 in my supply, **fig. 3**, provides this protection.

When soldering cosmos ICs into a circuit, use an iron with a grounded tip. Finally, make sure that your paddle key is clean. Any leakage path to circuit ground that is less than one megohm will falsely trigger the input gates.

The sidetone generator is optional, but can be included if your present rig doesn't have one. The sidetone adds about 3 mA to the key-down current drain.

reference

1. *COSMOS Digital Integrated Cricuits*, RCA Data Book Series, SSD-203A, RCA Semiconductor, 1973.

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some design ideas for specialized communications receivers

If you have ever considered building your own communications receiver, here is a collection of design ideas you may want to use

Ray Moore, 116 Walnut Street, Walpole, Massachusetts

When I read the editorial in the June 1972, issue of *ham radio* I felt like th boy who, after Shoeless Joe Jacksor confessed his involvement in the 191 Black Sox scandal, said, "Say it ain't so Joe." In that editorial Jim Fisk declared " . . . we at *ham radio* note that the day of the amateur-built receiver may hav passed in favor of the vastly superior anc less expensive commercial version."

Say it ain't so, Jim. Yet, in one sense, a least, the statement is true. The individua amateur can no more hope to beat th professional receiver designer at his owr game than he can hope to build a bette and cheaper family sedan in his back yard.

The catch is that the professiona receiver designer's "game" is *not* to de sign the ultimate performance set but tc design a mass-market, multiband, multi mode, decorator-styled receiver that car sell at a popular price. Many of th homebuilt receivers appearing in the ham magazines are imitations of these commercial designs and, while their builders have the statisfaction of gaining valuable experience and operating homebrew equipment, the result is likely to be an inferior receiver at a higher cost.

Just as there are cars which will outperform the best that Detroit offers, so there are receivers which will outperform any amateur receiver on the market. Free the designer from the price restriction and you get cars like the Jaguar and Rolls-Royce and receivers like the National HRO-600 and the Collins 651-S1. If the product can bypass the mass-market criteria, if it can be designed for a specific, limited purpose, then, as W8YFB has said,[1] any one of 100,000 teenagers can build a performance car that will outperform Detroit's creations on the drag strip. And many an amateur can build a better receiver than he can buy.

Goodman made the point back in 1951, " . . . the fellows who build their entire receivers from scratch . . . don't do this because they can't afford a commercial job – often they . . . could afford several such receivers – they do it because they know exactly what they want and this is the only way to get it."[2]

Let's look at some of the restrictions imposed on the professional designer of amateur communications receivers. First, the company's management and marketing people place a selling price of, for instance, $400 on the receiver. Unfortunately, advertising, sales costs, shipping, warranty costs, taxes, engineering, overhead and manufacturer and distributor profits eat up $250 of the selling price. This leaves less than $100 for the cabinet and all the parts which go into it. Finally, only $50 is left for all direct labor on the receiver, assembly, alignment, testing and troubleshooting.

The receiver must cover all modes and the six bands from 160 thru 10 meters with single knob bandswitching. It must fit a low-profile, decorator-styled cabinet that matches the other equipment in the company's line. All components must be readily available in quantity and performance must be competitive with other receivers in the same price class.

The homebuilder, on the other hand, can design his receiver for specific objectives. He can design a "performance" or "competition" receiver for a particular band and activity. How about an all-out receiver for 160-meter CW DX or 20-meter ssb DX or 20-meter sstv or 40-meter RTTY or for 20-meter operation in downtown Los Angeles? Net operators could have a switch-tuned, maximum convenience job. The traveling man can design an outstanding, ultraportable receiver.

The limited versatility of such receivers is justified because the amateur may spend 80% or more of his spare time enjoying his particular operating specialty. His chrome-plated commercial job then becomes a second receiver for the occasional excursions to other bands and modes. Sometimes the homebuilt "performance" receiver makes an excellent tunable i-f for other bands.

When designing your receiver at home you need consider only your own interests and radio environment. You can use one-of-a-kind, high quality surplus parts that commercial people can't afford. You can add extra parts and extra stages – if a commercial designer eliminates a $5 stage he saves his company $50,000 over a 10,000 unit production run, and employers look with favor on people who put $50,000 back into the company's pockets. Furthermore, you can use circuits that require considerable diddling and alignment, and you can use the latest components and techniques without worrying about obsoleting a large inventory of parts.

One of the delights of home building is that you can over-design. Instead of using the flimsiest possible chassis material you can start out with a 1/8-inch aluminum plate. You automatically insert parasitic suppressors in all amplifier and oscillator input and output elements. Where the professional runs endless tests to eliminate a capacitor here, a resistor there or a shield somewhere else, the home-builder can double decouple every stage and put in extensive shielding. Where the professional must run two i-f stages on the verge of oscillation, you can run three that loaf quietly along. Space, band-switching and ganging considerations need not limit the number and Q of the tuned circuits used in the front-end.

The receiver shown in this article, the DBR-1, illustrates many of the advantages the homebuilder enjoys. Its primary purpose is to provide superior reception of foreign a-m stations on the standard broadcast band in the presence of interference from the many domestic broadcasters. Secondarily, it is an excellent tunable i-f for shortwave broadcast and amateur ssb reception. While few amateurs are interested in broadcast-band DX, this receiver includes many techniques and ideas that are applicable to homebuilt receivers in general.

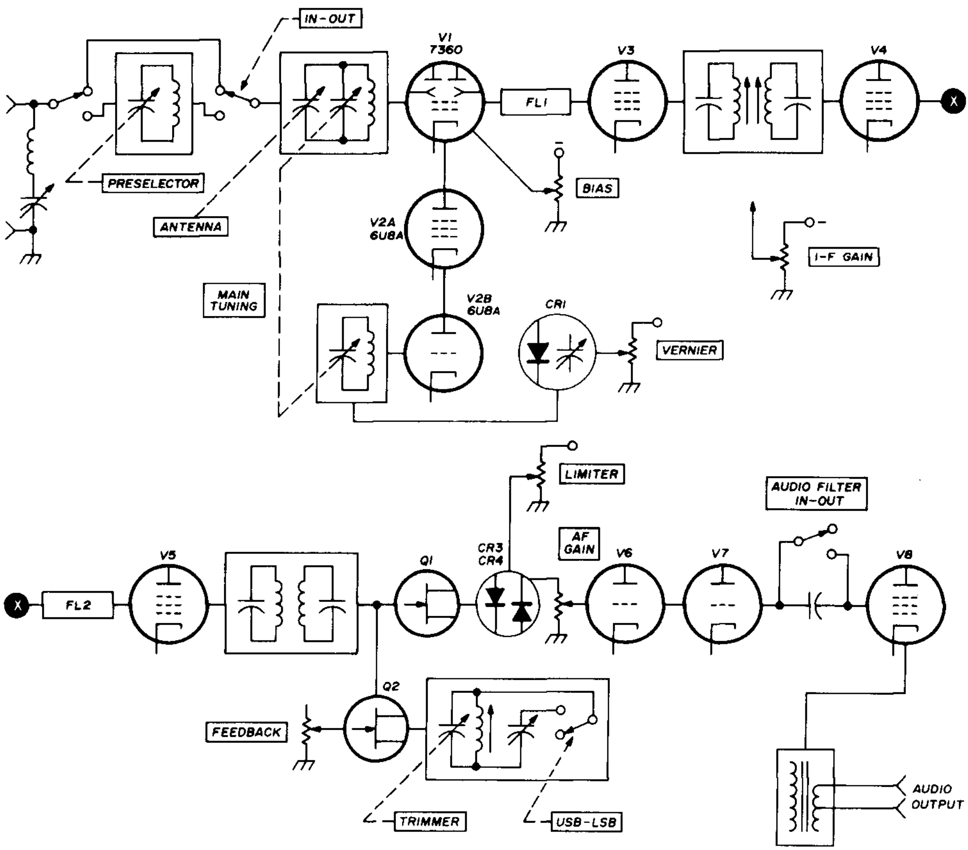


fig. 1. Block diagram of the DBR-1, a specialized DX receiver.

homebrew receiver

Vacuum tubes predominate in the DBR-1 because it was started back in 1967 when tubes had performance advantages in many circuits. Referring to **fig. 1,** the first circuit is a series-tuned i-f trap to eliminate feedback problems encountered at high i-f gain levels with no rf stage.[3] Next is an additional preselector circuit which can be switched in to eliminate images when using a long-wire antenna. A tuned loop antenna is normally used with the receiver and then images are not a problem.

The first tube, V1, is a 7360 mixer. The vfo and buffer amplifier is a 6U8A. A varactor diode, CR1, provides vernier tuning of the oscillator to facilitate precise positioning of the carrier for exalted-carrier a-m reception. Bias for the 7360 mixer is adjustable from the front panel for experimental purposes.

A 455-kHz i-f is used to take advantage of the sharp and inexpensive Kokusai mechanical filters. In accordance with good strong-signal practice, the first filter is placed as close to the antenna as possible, at the output of the mixer.[4] A second mechanical filter, between the second and third i-f stages, combines with the first to give a stop band over 120-dB down and a shape factor approaching 1:1. Three i-f stages (V3, V4 and V5) are necessary to provide sufficient gain to make up for filter losses and the lack of an rf amplifier.

Transistor Q1 is an infinite impedance detector; Q2 is a Q-multiplier to provide the peak in the i-f passband which exalts the a-m carrier. A switch changes the peak from one side of the passband to the other for receiving either USB or LSB. The feedback control adjusts the height of the peak and thus the amount of carrier exaltation. Diodes CR3 and CR4 are a full-wave audio limiter.

The gain of the receiver is kept as low as possible through the detector, and three stages of audio (V6, V7 and V8) are required to bring the overall gain up to an acceptable level. A small coupling capacitor to the output stage can be switched in to provide low frequency roll-off. A high-impedance output for a tape recorder and a low-impedance output for stereo phones or a speaker are also provided.

An OB2 voltage regulator, V9, regulates the oscillator plate voltage and Q-multiplier drain voltage. A zener diode, CR2, also works from the +105-volt regulated line to provide additional regulation for the varactor tuning diode. The power supply is in a separate cabinet. It uses five silicon rectifiers and supplies 6.3 Vac for the heaters, a 250-volt B+ supply controlled by an auto-transformer and a 0-150-volt bias supply.

mechanical details

Here is where most receivers are inadequate. They are flimsy, using the lightest possible material for the chassis and cabinet. The DBR-1 has a 1/8-inch aluminum panel bolted to a 1/8-inch chassis plate which is then bolted into a rugged, welded steel cabinet from an old Meissner Signal Shifter. When completely buttoned-up the whole thing is almost a solid cube. I use relay rack panels for the 1/8-inch material and work it with a hacksaw, electric drill, file and ordinary chassis punches.

Another mechanical weakness in many commercial and homebuilt receivers is the tuning mechanism. The homebuilder can use one of the rugged military surplus mechanisms or pick up an old National PW dial and drive as I did for the DBR-1. A tuning mechanism with a solid, smooth feel gives the impression of quality much as the "thunk" of a well-fitted door does to an automobile.

The plug-in coil assemblies are from an old National HRO receiver. They are excellent for experimenting with different coils. You can try the effect of different coil form materials on oscillator drift, for instance, or determine the effect of different antenna couplings on front-end selectivity. The coil drawers were sawed and chiseled from an old HRO chassis.

mixer circuit

The ideal mixer should be able to *simultaneously* handle the strongest and weakest signal you will encounter while contributing little noise itself and providing sufficient conversion gain to reduce the noise of the i-f strip to insignificance. Not many mixers meet these requirements. The 7360 does to a greater degree than most and is used in the DBR-1.

The overall noise factor of the mixer plus i-f strip is

$$F = F1 + \frac{F2 - 1}{G1} \quad (1)$$

where F is the overall noise factor of the complete receiver, F1 is the noise factor of the mixer, F2 is the noise factor of the i-f strip and G1 is the power gain of the mixer.

Assume the 7360 mixer with its preselector circuits has an 8-dB NF (6.3X) and the i-f strip has a 13-dB NF (20X), including the 6-dB insertion loss of the input filter, and the mixer power gain is 10 dB (10X).

$$F = 6.3 + \frac{20 - 1}{10} = 8.2X \text{ or } 9.2 \text{ dB}$$

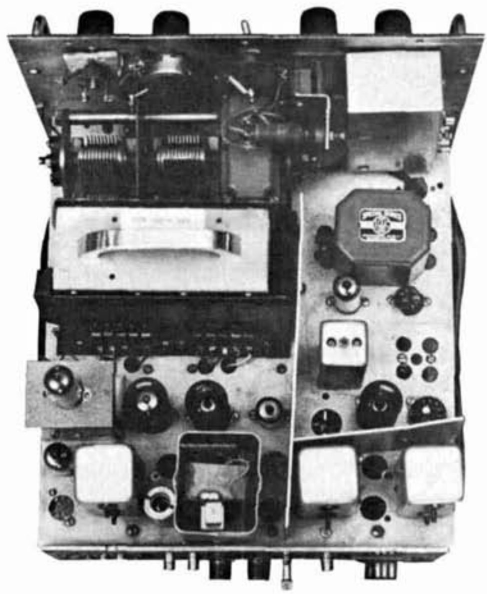
Suppose you contemplated using a diode balanced mixer instead of the 7360 into the same i-f strip. Suppose also that the diode mixer has the same 8-dB NF as the 7360 but it has no gain (G = 1). Therefore,

$$F = 6.3 + \frac{20 - 1}{1} = 25.3X \text{ or } 13.9 \text{ dB}$$

A screwdriver adjustable pot on the front panel is for experimenting with the effect of bias on the sensitivity and overload characteristics of the 7360. In my location two 50-kW locals, WBZ on 1030 kHz and WHDH on 850 kHz, produce a strong third-order IM product on 1210 kHz which is convenient for evaluating the linearity of the mixer. At one bias point, which is quite sharp and may vary from one 7360 to another, the IM product nulls out. On the tubes with which I have experimented the point of maximum immunity to IM is also very close to the point for maximum sensitivity.

vfo circuit

The vfo and buffer amplifier use two sections of a 6U8A. A 20-pF varactor diode connected across the oscillator tank circuit provides a vernier tuning control to precisely set the carrier. The control voltage comes from the regulated +105-volt line and is further regulated with a 12-volt zener diode.



Top view of the DBR-1 shows the 1/8" chassis plate which is bolted to the front panel and the rear apron. The i-f strip runs along the back with filter FL2 in the open shield can in the center. From left to right below the coil drawer are V8, V1, V2, V7, V6 and Q1. Below the output transformer are the voltage regulator, V9, and CR3 and CR4.

The National tuning capacitors which are an integral part of the PW gear drive have straight-line frequency characteristics. When the DBR-1 is used as a tunable i-f with the 3.5- to 4.0-MHz coils, the calibration is 1kHz per division of the PW dial. The 0.9- to 2.0-MHz coils give 2 kHz per division across most of the dial.

i-f strip and filters

The i-f strip provides the required adjacent channel selectivity with enough gain to drive the detector on the weakest signal and should have a low enough noise figure to prevent degrading the NF of the mixer. The noise factor (ratio) of a properly designed i-f strip is

$$F_{i\text{-}f} = F1 + F2 \qquad (2)$$

where $F_{i\text{-}f}$ is the overall noise factor of the i-f strip, F1 is the insertion loss of the input filter and F2 is the noise factor of the first i-f tube.

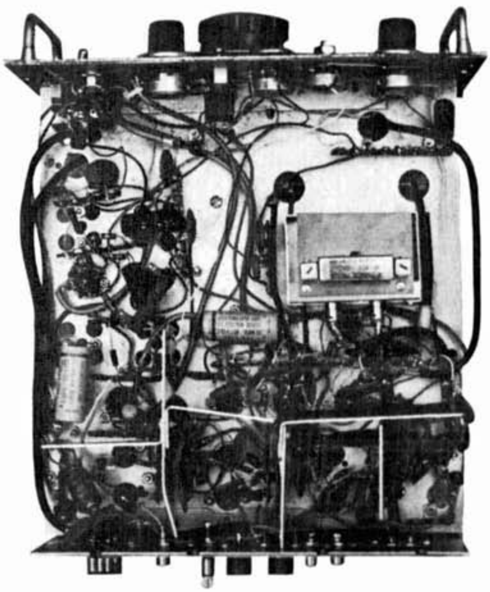
Cascading mechanical or crystal lattice filters allows you to achieve near perfect shape factors and stop bands down more than 120 dB. The National Radio Club has done considerable work on the cascading of mechanical filters[5] and the club periodically purchases batches of Kokusai filters which it matches and sells to members in sets of two or three. Each Kokusai filter comes with a complete set of characteristics and the current units are extremely good.

Kokusai filters are used in the DBR-1. Filter FL1 has a center frequency of 454.98 kHz, a bandwidth of 2.58 kHz at the 6-dB points and a −60 dB bandwidth of 4.50 kHz, resulting in a 6/60-dB shape factor of 1.74:1. Filter FL2, at the same points, is 455.00 kHz, 2.48 kHz, 4.26 kHz and 1.72:1.

The combined characteristics of the two filters in the DBR-1, checked both by the point-by-point method and by the sweep method to be described later, are 2.1 kHz at −6 dB, 2.45 kHz at −10 dB and 2.8 kHz at −60 dB, resulting in a 6/60 dB-shape factor of 1.33:1 and a 10/60 dB shape factor of 1.14:1. The 10/60-dB shape factor is given because it seems to describe more accurately the *useful* nose bandwidth and gives a better idea of the *steepness* of the skirts than does the usual 6/60-dB figure.

For several reasons it is best to distribute the filters in the i-f strip rather than connecting them back-to-back.[6,7] One of the filters should be at the input to the i-f strip to satisfy the requirement of putting the adjacent channel selectivity as close to the antenna as possible. Additional filters should be placed one or two stages further along.

If all the filters were placed at the input to the i-f strip the following amplifiers could generate considerable wideband noise outside the bandpass of the filters. Also, two or more cascaded filters are capable of stop bands more than 120 dB down and if they are lumped together it requires a lot of isolation between the



Bottom view of the DBR-1 receiver. Filter FL1 is in right center. Note the shielding around each i-f stage at the bottom. Rf chokes at bottom left filter the incoming B+ and bias voltages.

input and output. If they are separated the isolation requirements can be reduced to 60 dB at a step. Another consideration is that two or three filters at the input to the i-f strip could degrade the NF of the strip enough to degrade the NF of the entire receiver.

A good bet for a filter for a high frequency i-f is the Swan SS-16. This 16-pole crystal-lattice filter is centered on 5.5 MHz. The 6-dB bandwidth is 2.7 kHz and it has a 6/60-dB shape factor of 1:28:1 and a 6/120-dB shape factor of 1/8:1. The stop band is 140-dB down.

Stop bands of −120 and −140 dB sound great and they are not too difficult to achieve. In practice, however, anything over 100 or 120 dB is of little practical use because there are mighty few front-ends which will simultaneously handle two signals 120-dB apart (in strength). For instance, 140 dB above 1 μV is 10 volts and most receivers will block completely long before the signal gets up to 10 volts. Thus, the fact that the filter could separate the 1-μV signal from the 10-volt signal is only of academic interest because it will never get the chance.

The filters must be complemented by a meticulous job of shielding, isolating, bypassing and filtering to prevent degradation of the filter characteristics. In the DBR-1 each stage is individually shielded and double decoupled and the heater leads are bypassed. A copper shield is soldered between the grid and plate pins of each i-f tube socket. These measurer insure a quiet and electrically stable i-f strip that takes full advantage of the filter characteristics.

checking the i-f passband

The most accurate and simple method of evaluating an i-f strip is to plot its passband and skirts point-by-point with a signal generator, counter and attenuator. This is fine for getting a picture of the final results, but it isn't something you want to repeat continuously during the alignment process.

The requirements for sweeping a steep-sided, narrow passband are much more stringent than those for sweeping a tv i-f strip. You must sweep at a very slow rate, 1 Hz or less, to show the steep sides and sharp corners. Commercial firms use special sets of equipment made by people like Rhode & Schwartz that sell for $10,000 or more.

Fortunately for home-builders there is a $130 piece of equipment that, with a simple variable frequency signal source, will do almost the same job. This is the Heath SB-620 Spectrum Analyzer. My experience is limited to the SB-620 connected for a 455-kHz i-f, but it should give similar results at higher frequencies.

Connect the end of the i-f strip to the i-f *input* jack on the rear apron of the SB-620 through a 5-pF, or less, capacitor. Place the function switch on the rear apron in the *ham scan* position. Set the *variable sweep rate* and the *variable sweep width* controls at the maximum counter-clockwise positions and the *sweep width* switch in the 50-kHz position. At these settings the sweep rate is about 2 Hz and the horizontal calibration of the screen is about 0.7 kHz per division.

Connect a variable frequency signal source to the input of the i-f strip and manually sweep it *slowly* (say 5 seconds per sweep) back and forth across the passband. The peak of the pip will trace out the passband of the i-f passband on the face of the CRT (see **fig. 2**).

Another method which is useful for examining the skirts of the passband is to feed the output of a noise generator into the i-f strip with the SB-620 connected and adjusted as before except that the *sweep width* switch should be in the 10-kHz position. The result is a trace of the passband as shown in **fig. 3.** The top of the trace is jagged due to the random amplitude of the noise pulses but you get a good picture of the skirts which can be extended down to −60 dB by using the *−20 dB log* position of the *amplitude scale* switch.

table 1. Comparison between some of the performance characteristics of a typical amateur-band receiver and the better military and professional class receivers.

	i-f feedthru	image ratio	stability	IM	i-f stop band	front-end dynamic range	frequency read-out
commercial amateur receiver	50 dB	60 dB	100 Hz per hour	60 dB	60 dB	60 dB	1 kHz
best professional & military receivers	100 dB	100 dB	1 Hz per month	100 dB	100 dB	130 dB	1 Hz

Optimum reception of weak a-m signals in the presence of noise, QRN and QRM demands the exalted-carrier technique. In addition, the exalted carrier must be phase locked to the original carrier so that the technique of chopping off the a-m carrier with the filter in a ssb receiver and then using the bfo to replace it is ineffective. You must either lock a locally generated carrier to the original carrier or filter, process and amplify the *original carrier*.

In the DBR-1 a 15- or 20-dB peak is superimposed on one edge of the passband with a Q-multiplier.[8] The carrier of the desired station is placed on the peak, exalting it in relation to the sidebands. The series resistor between the Q-multiplier and the i-f transformer which is advocated in the referenced article is not used because the resistor restricts the height of the peak to only about 10 dB which is not sufficient.

When used for carrier exaltation the Q-multiplier is best placed at the very end of the i-f strip. Here it is protected from off-frequency stations by the filters and it "sees" all inband signals at about the same strength since the gain of the receiver is largely controlled by the i-f stages. The Q-multiplier is double decoupled and completely shielded and has parasitic supression resistors in both the gate and drain leads. Voltage is taken through a divider from the +105 volt line. The result is a remarkably stable Q-multiplier which can be set on the verge of oscillation, if desired, and remain there indefinitely.

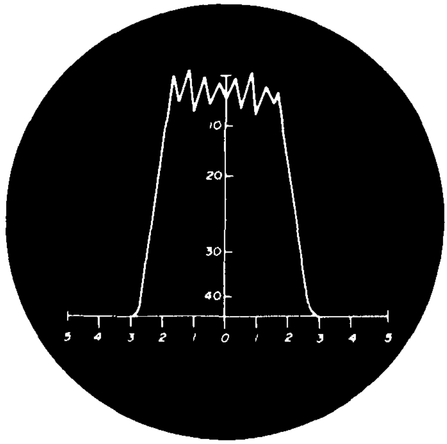


fig. 3. I-f passband of DBR-1 as displayed on SB-620 with white noise fed into i-f strip input. This gives good picture of the skirts which can be extended down to -60 dB with the built-in 20 dB attenuator. Horizontal calibration is about 0.5 kHz per division.

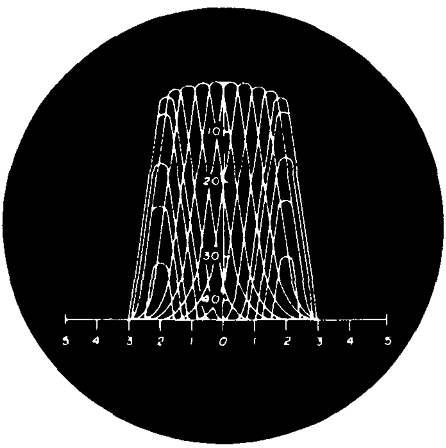


fig. 2. I-f passband of DBR-1 as traced on *SB-620 by slowly sweeping a CW signal across* the i-f. Note that the sides of the passband are defined by the ascending and descending row of peaks at each side, not by the skirts of the pips. In practice only one pip is seen at a time, first climbing up one side, then across the top, and finally down the other side. Horizontal calibration is about 0.7 kHz per division, vertical in dB.

The Q-multiplier is fix tuned to either the upper or lower edge of the passband and can be switched from one to the other for sideband selection. A small trimmer is mounted on the front panel for those occasions when the receiver is used as a tunable i-f. The Q-multiplier is then made to oscillate to serve as a bfo for ssb reception and the trimmer is used to position the carrier down the side of the passband.

audio amplification

This section of the receiver is often taken for granted, yet proper attention to detail here can make a significant improvement in the readability of signals. It has been shown that if the high speech frequencies are cut it is necessary to also cut some of the lows to maintain optimum readability.[9,10] In a properly designed ssb transmitter the lows are cut in the audio and filter stages. A-m broadcasts, on the other hand, carry the full audio range out to at least 5000 Hz. If you cut the highs at 2500 Hz and leave the lows, the speech will sound boomy and muffled. Roll off the lows below 300 or 400 Hz and the speech becomes crisp and readable.

Various exotic audio filters were tried in the DBR-1 with expensive high-Q inductors but the simple trick of using a small coupling capacitor to the audio output stage worked as well as any in practice. The switch marked *audio filter* shorts out the capacitor if the full response of the amplifier is desired.

The audio noise limiter shown in the diagram is a solid-state copy of the limiter used in the Collins R390. Like all other audio limiters and clippers tried in the DBR-1 it is ineffective because of the filters and Q-multiplier ahead of it. Experiments with i-f limiting just before the second filter have been much more favorable. The idea is to limit all signal and interference components in the passband to the level of the desired sidebands up to the point of carrier exaltation.

conclusion

The DBR-1 is a tailor-made receiver for my interests and location. It is superior to any receiver available commercially for the purpose for which it was designed and in the environment in which it is used. It is superior only because the market for such a receiver is so limited that no manufacturer could afford to design and build a receiver for that purpose. The future of homebuilt receivers is in this area of "performance" or "competition" — receivers which are designed for ultimate operating characteristics for a specific, limited task.

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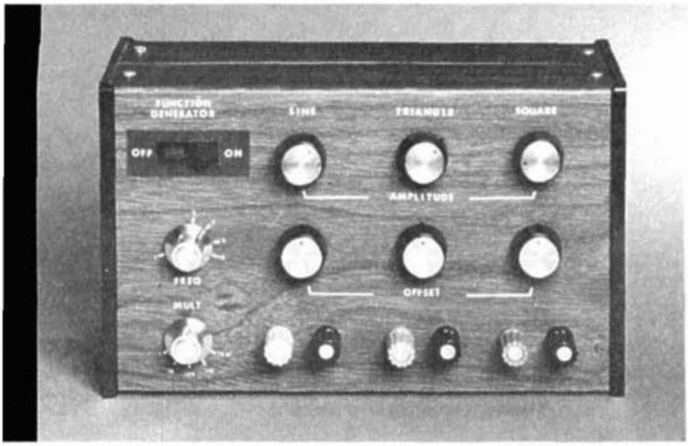


integrated-circuit function generator

Construction of a variable function generator that provides sine, square and triangular waveforms over the frequency range from 1 Hz to 100 kHz

Ray Megirian, K4DHC, Box 580, Deerfield Beach, Florida 33441

I recently had need for a second audio generator when I became involved with some circuits requiring two different and simultaneous audio frequency inputs. After considering the alternatives, I decided this would be as good a time as any to put to use a new waveform generator IC that I purchased some time ago. I expected that the IC would minimize the time required to go from drawing to finished generator. This proved to be the case. I highly recommend the function generator described here to anyone who needs a compact audio signal generator.

general description

The heart of this instrument is a waveform generator IC manufactured by Intersil Inc. and known as the 8038.* The IC will produce sine, square and triangle waveforms over the frequency range from 0.001 Hz to 1.0 MHz. Frequency may be continuously varied with a pot, programmed with fixed resistors or swept with a voltage ramp. The function generator described here covers the range from 1.0

*The version of the 8038 IC used here is the *8038CC which is an economy copy selling* for under $4.00 in singles. Since Intersil products are not widely distributed through outlets available to most amateurs, drop me a line and let me know if you wish to purchase one or more of these ICs. If the demand is sufficient I will order enough to take care of requests. An sase would be appreciated.

Hz to 100 kHz in 5 decade bands. A pot is used as the vernier element and a 5-position switch as the band selector.

Since the outputs from the 8038 IC are not all of uniform amplitude nor at useful power levels, an amplifier was added to each output along with trimmers for adjusting the output to a standard level. The amplifiers each consist of an op-amp and booster which is capable of driving low-impedance loads.

One final feature which was included is adjustable offset for each of the outputs. This function may or may not be useful to you, but I find it helpful on occasion and decided to include it since it only required the addition of 3 pots and 3 resistors.

8038 IC

This IC contains quite a bit of circuitry with over 50 transistors, some diodes and dozens of resistors squeezed onto the chip to accomplish the desired functions. **Fig. 1** is a block diagram of the IC showing the relationships between the major sections of the waveform generator.

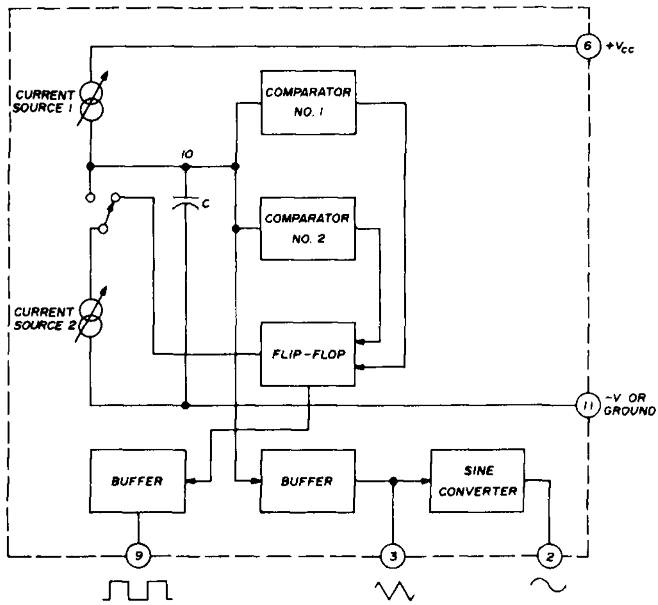


fig. 1. Block diagram of the 8038 waveform generator.

Initially, current source 2 is disconnected and an external timing capacitor, C, is charged through current source 1. When the voltage across C reaches the trigger threshold of comparator 1, an output is generated which causes the flip-flop to change state and current source 2 becomes active.

As the capacitor discharges towards a negative peak, it reaches the firing threshold of comparator 2 whose output resets the flip-flop, disconnecting current source 2 and permitting the cycle to repeat. The current level supplied by the two current sources is controlled by either a series resistor to the power supply or an external bias voltage.

The bias voltage controls both sources simultaneously and is normally used for fm or swept operation of the oscillator. Since the power supply connections for the current sources terminate at separate pins, current levels may be controlled independently with separate resistors. This latter method is used when unsymmetrical waveforms such as a pulse or sawtooth is desired.

By varying the ratio or charge to discharge time of the timing capacitor, the square-wave duty cycle may be varied from 2% to 98% and the triangular wave adjusted for either a positive or negative sawtooth or ramp.

The waveform appearing across the capacitor is internally fed to a buffer amplifier and then to both the sine converter and the output (pin 3). With a triangle input to the sine converter, total harmonic distortion of the resulting sine wave is typically less than 1%. With proper adjustment distortion levels as low

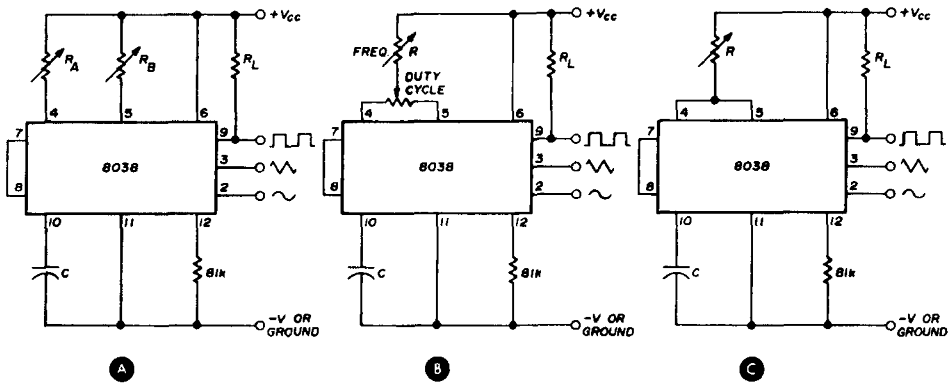


fig. 2. Three possible methods for connecting external timing resistors.

as 0.5% are possible, according to the data sheet.

The square wave is taken from one side of the flip-flop and is fed to a buffer amplifier having an uncommitted collector. Supply voltage for this stage is thus independent of the rest of the circuitry and allows use of a separate 5-volt supply for TTL compatibility, if desired.

The IC will operate from either a single supply of from 10 to 30 volts or dual supplies of ±5 to ±15 volts. With dual supplies all outputs will be symmetrical about ground.

external connections

Fig. 2 shows three possible connections for the external timing resistors at pins 4 and 5. Potentiometer R_A controls the rising portion of the sine and triangle and the zero state of the square wave. The method used in **fig. 2A** is most desirable for large changes in duty cycle while that shown in **2B** is used mainly to control frequency with minor changes in duty cycle. A 50% duty cycle results when $R_A = R_B$ and if no adjustment is contemplated, the connection shown in **fig. 2C** may be used. Resistance values for R_A and R_B may vary from a minimum of 500 ohms to a maximum of 1.0 megohm.

Both frequency modulation and sweeping may be applied to the waveform generator. **Fig. 3** illustrates how this is accomplished. For small deviations the modulating signal is applied to pin 8 through a coupling capacitor. Pin 7 is an internally-set bias voltage which is normally applied to the current sources by a direct connection between pins 7 and 8.

The input impedance at pin 8 is around 8000 ohms and may be raised by inserting a resistor between the two pins as shown. For large deviations, or sweep-

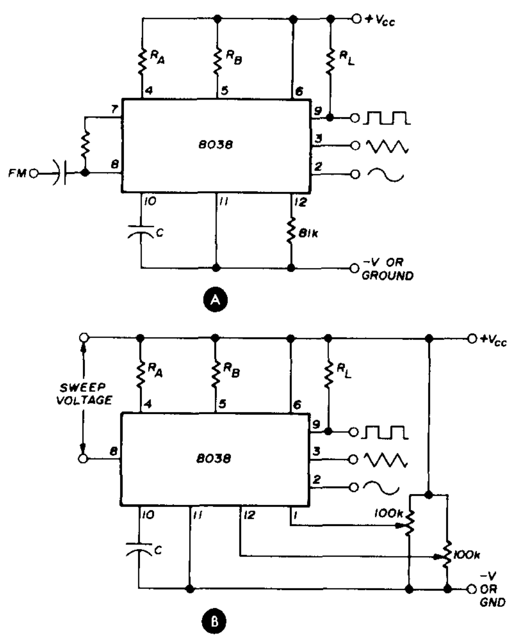


fig. 3. Connections for fm (A) and sweep operation (B) of the 8038 IC.

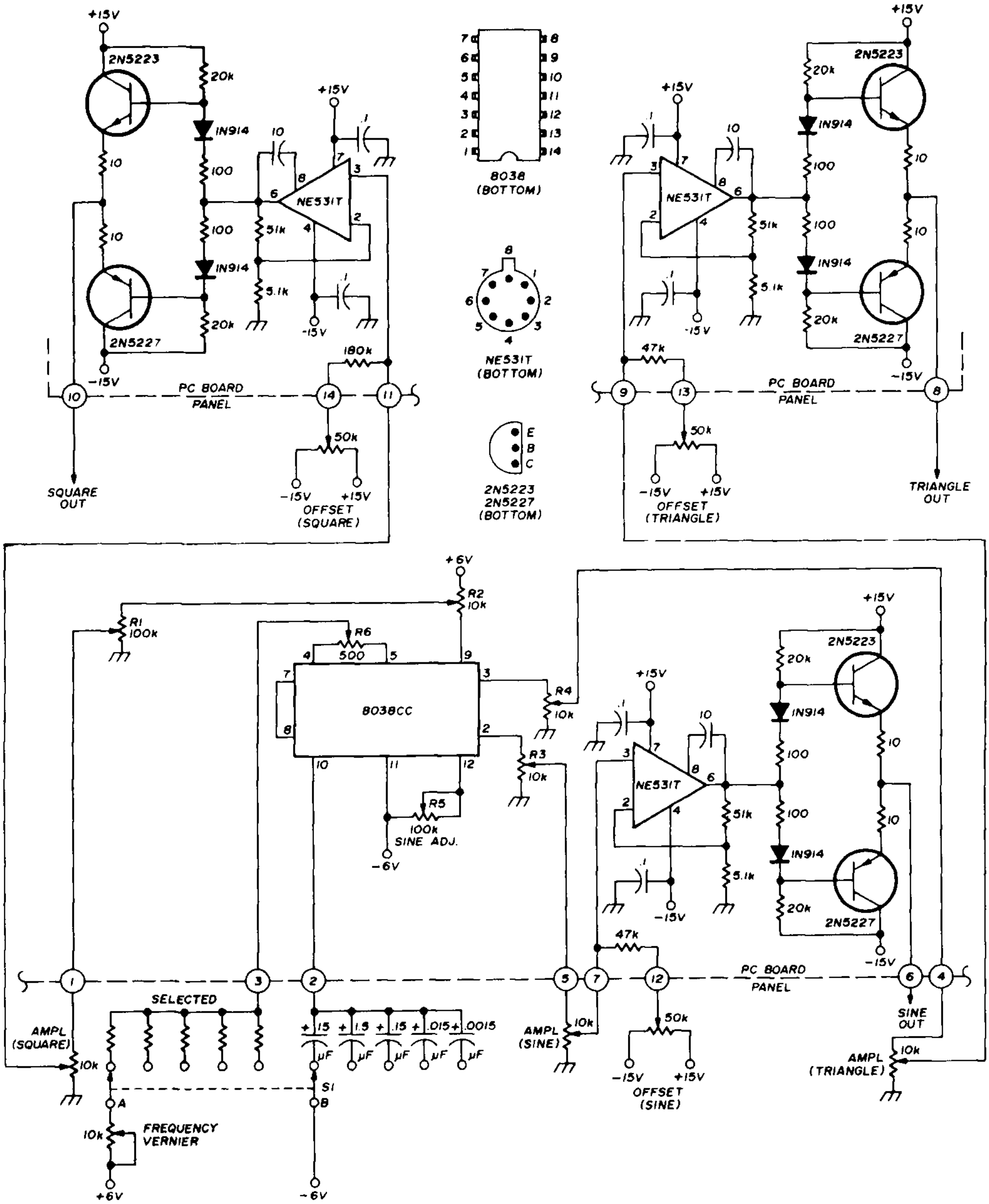


fig. 4. Circuit diagram for the function generator. Numbered leads correspond to holes provided in the PC board. All resistors are ¼ watt, 5%.

ing the oscillator, pin 7 is disconnected and the sweep voltage is applied between the positive supply and pin 8 as in **fig. 3B.** The sweep voltage must be confined within the limits V_{cc} and 2/3 V_{cc}.

It may be noted that in most of the diagrams discussed so far, an 81k resistor is shown connected from pin 12 to the negative supply. Pin 12 is a sine wave adjustment point and 81k is an approximate value for minimizing distortion. If a 100k pot is substituted for this resistor, a more accurate adjustment is possible. Additional improvement in correcting sine wave distortion is possible by applying correction at both pin 1 and pin 12 as illustrated in **fig. 3B.**

in the output afford some protection but it is not recommended that the output be short-circuited on a regular basis.

Neither fm nor sweeping operation were incorporated in this instrument since I had no need for these functions. My other audio generator has wide range

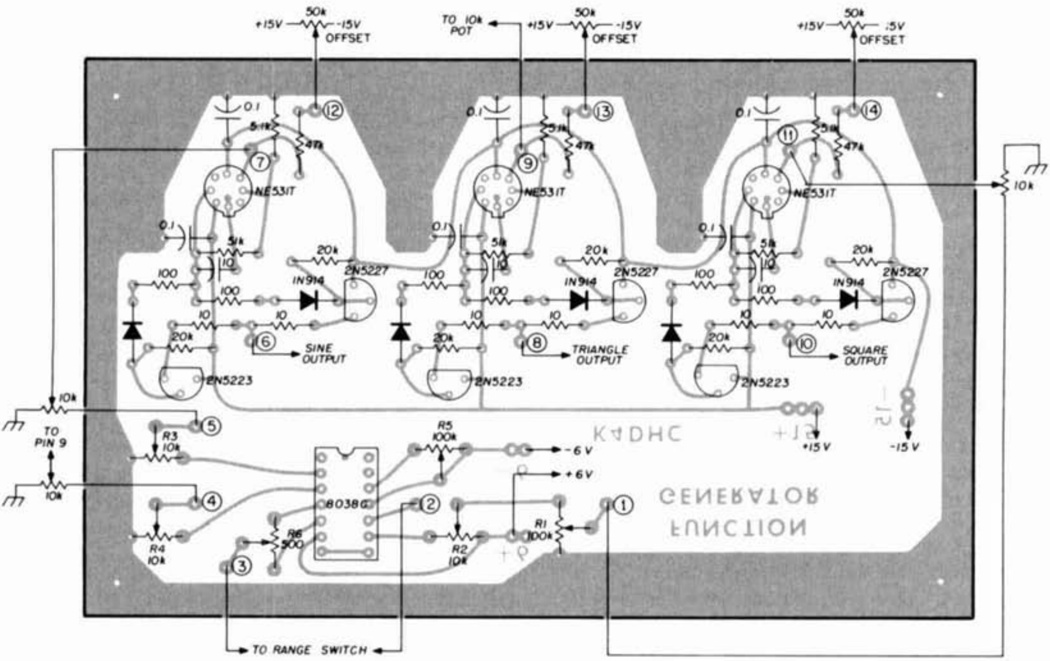


fig. 6. Component layout for the function generator board. Numbered holes correspond to numbers on the schematic for external connections. All diodes are 1N914 or equivalent. Printed-circuit is shown in fig. 8.

function generator

The circuit for the function generator is shown in **fig. 4.** To achieve the desired bandwidth of 100 kHz, a high performance op-amp was required for the three output stages. The Signetics NE531T was chosen since it is available from several sources at a reasonable price. This amplifier requires only a single external capacitor for compensation and has a full-power bandwidth well beyond 100 kHz. It also shows no tendency towards instability.

The current boosters are a pair of complementary transistors with diode/resistor networks in the input to prevent crossover distortion. The 10-ohm resistors

sweep capability as well as CW operation so the function generator was kept to CW only. Of course, there is no reason why these functions could not be included if required and the preceding discussion should be helpful towards that end.

Power supplies need not be regulated since frequency is not dependent on voltage as far as the IC is concerned. If external bias is applied to pin 8 for either sweeping or setting frequency, however, stability will be affected by any fluctuations in this control voltage. It should be regulated.

Frequency drift of the vco with temperature is 50 ppm/°C typical. This is quite good and applies over the range from 0°C to 70°C. In order to maintain low dissipation and allow little heating of the 8038, the IC is run at the relatively low voltage of ±6 volts. The remaining circuitry operates at the full ±15 volts.

The circuit for the power supply is shown in **fig. 5.** A PC-type power transformer was used since it was available, but a suitable substitute could be mounted in the same space if not too large. The power supply circuit board layout and parts placement are shown in **fig. 7.**

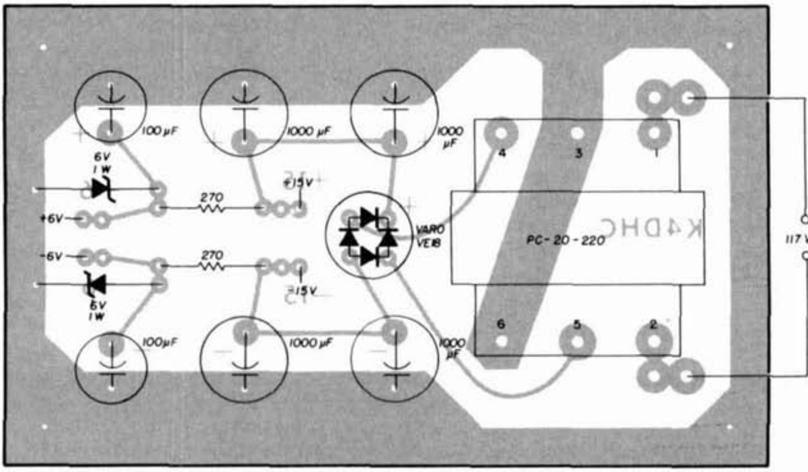


fig. 7. Full-size component layout for the power supply. Printed circuit is shown in fig. 8.

construction

All leads from the front panel of the instrument are routed directly to holes provided on the printed-circuit board. The five timing capacitors and five calibrating resistors are mounted right on the rotary switch assembly. The resistors are not installed until calibration has been completed, at which time the proper values will have been determined. The printed-circuit board is 5.6-inches (14.2 cm) long by 3.5-inches (8.9 cm) wide. A full-size layout is shown in **fig. 6.**

calibration and adjustment

When you are ready to fire up the function generator, preset all the trimmers on the PC assembly to *mid-position* and temporarily connect a 1k pot in series with the arm of S1A and lead no. 3 from the PC board. Set the pot for maximum resistance and turn the band switch to position 3 (100 to 1000 Hz). Turn the frequency control pot to the high end and the offset controls, if used, to mid-position *(zero offset).*

The triangle will be checked first since it will most clearly indicate proper symmetry. Connect your scope to the triangle output and apply power to the

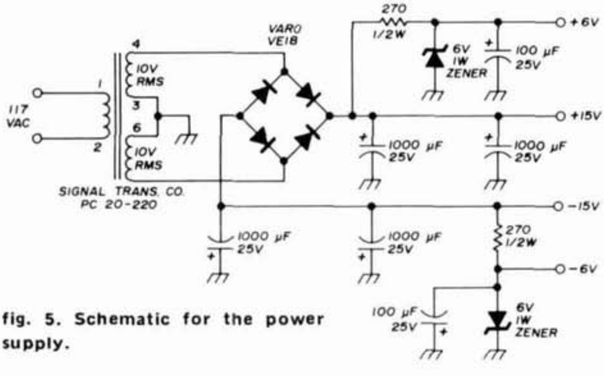
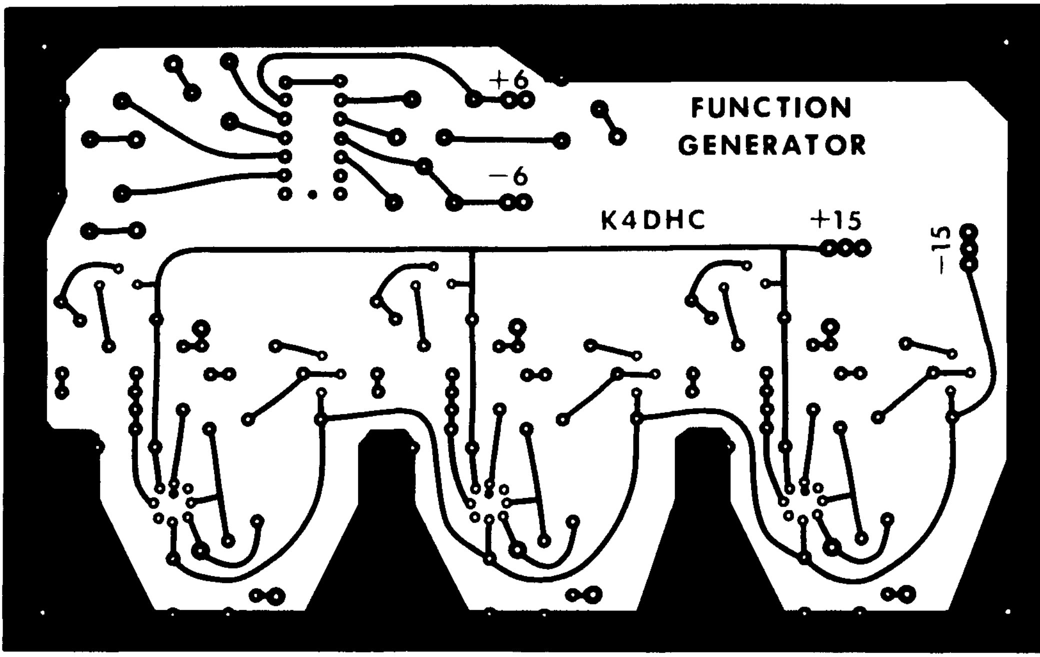


fig. 5. Schematic for the power supply.



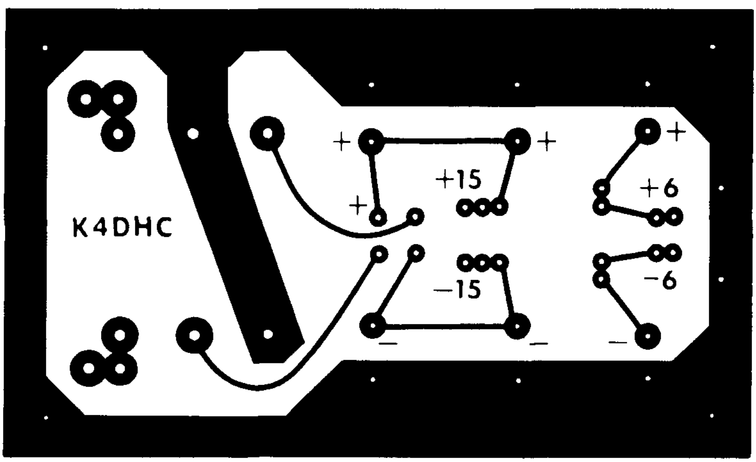


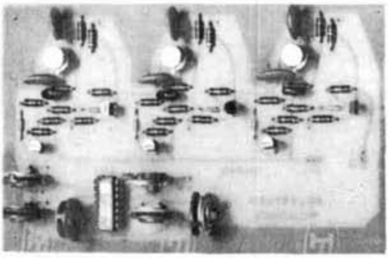
fig. 8. Full size printed-circuit layouts for the function generator (above) and power supply (below).

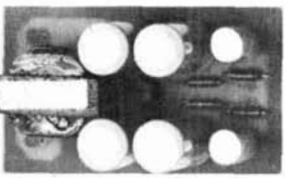
generator. With the amplitude control at maximum, adjust trimmer R4 for a 10-volt peak-to-peak output. Adjust R6 for proper symmetry and then the temporary 1k pot for a frequency of 1.0 kHz. If a slight overlap is desired, adjust for a slightly higher frequency. Oscillator frequency with the vernier control at minimum should be 100 Hz or less.

The sine wave should be adjusted next, so transfer the scope probe to the sine wave output. With the amplitude at maximum, adjust R3 for a 10-volt peak-to-peak output. Adjust trimmer R5 for minimum distortion of the sine wave.

Next look at the square wave output. Both R2 and R1 must be adjusted for the desired 10-volt peak-to-peak output signal. Trimmer R2 will cancel out the offset voltage present at the square wave *output and R1 will control the amplitude.* There is interaction between these two controls, but juggling back and forth a few times will accomplish the desired results.

Measure the resistance of the temporary pot and make a note of the required value for this band. Go to each of the remaining bands in turn and





The completed function generator and power supply circuit boards.

determine *a value of resistance to calibrate* each one. Remove the 1k pot and install fixed calibrating resistors at the proper points on the bandswitch.

The 1.5- and 15-μF timing capacitors for the two lowest frequency bands should preferably be tantalum types. The remaining capacitors may be mylar or polystyrene. If any band gives calibration trouble, it may be due to a capacitor with a wide variation from the marked capacitance value and a replacement may be necessary. The trimmer resistors are the blue plastic vertical PC mounting type made by CTS and available from many distributors or at Radio Shack stores. The 0.1-μF bypass capacitors are 50-volt discs while the 10-pF units are small dipped micas.

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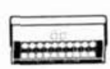

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the need for coherent frequency-shift keying in RTTY

A discussion of modern communications theory, and how it may be used to improve the performance of RTTY communications

Stephen A. Maas, K3WJQ, 1195B Bear Mt. Drive, Boulder, Colorado 80303

In recent years many modern designs have been presented for high performance RTTY demodulators.[1,2,3] These units have used a-m demodulation techniques, acknowledged to be superior to fm, and incorporate excellent autostart features. Unfortunately, many RTTY enthusiasts consider these demodulators to be optimum for demodulating signals in noisy environments not realizing that, with slightly more elaborate signal-processing techniques, it is possible to achieve considerably better performance, especially when the signal-to-noise ratio (snr) is low. One such technique, as yet little explored in amateur radio communications, is the use of coherent demodulation.

snr improvement

It is universally recognized that there are severe limits in improving the snr at the receiver input — the most common method is to increase transmitter power. Those amateurs who admire the trend toward low power and solid-state transmitters are not at all pleased with this approach; it is expensive, limited to one kilowatt, and worst of all, does nothing to improve reception at your own station.

With an rf preamplifier using modern transistors, the noise figure of virtually any amateur communications receiver can be reduced. However, in all but the most mediocre receivers electronic noise is insignificant as compared to atmospheric noise and radiation from the sky and earth. Also, it would be possible to improve RTTY performance by increasing the duration of the mark and space pulses because the probability of a receiver error (mistaking a mark pulse for a space pulse or vice versa) decreases with increase in pulse energy. Because of the standardization of mark and space pulses, however, this approach is impossible. This leaves signal processing as a technique for improving RTTY performance. The use ever, fm-based systems exhibit a threshold effect, and below a certain critical snr the signal becomes completely lost in noise. For this reason a-m systems, which at worst exhibit only a minute threshold effect, are preferred when difficult noise and interference conditions exist (the usual conditions on most amateur bands).

The block diagram of a high-performance a-m FSK demodulator is shown in **fig. 1**. The first stage is a filter, used to minimize interference, which admits both mark and space signals. It is followed by separate mark and space filters and envelope detectors; these are

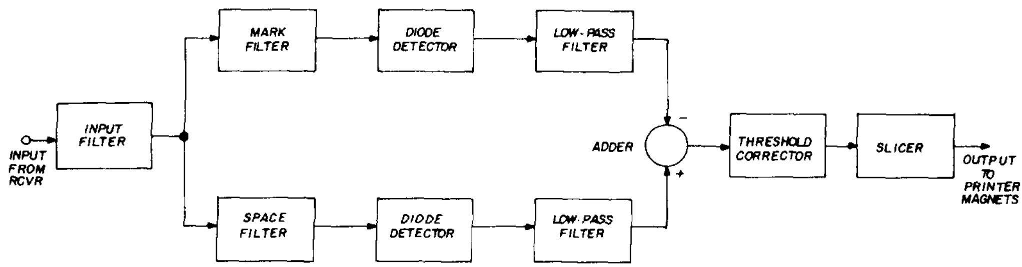

fig. 1. Block diagram of an RTTY demodulator based on a-m techniques.

of coherent techniques over envelope demodulation of the RTTY mark and space pulses offers the best practical improvement.

RTTY demodulation

Frequency shift keying (FSK) or radio-teletype (RTTY) is a type of modulation where the transmitted signal alternates between two distinct frequencies. Hence, it may be treated as a digital form of fm and demodulated by passing the signal through a limiter and discriminator. Alternatively, FSK may be considered as two on-off keyed signals and can be demodulated with a-m techniques; that is, using no limiter and an envelope or product detector.

Because of their limiter, fm systems are capable of practically eliminating noise when the input snr is good. However,

diode detectors followed by simple low-pass filters, and their output is the envelope of the input pulse (see **fig. 2**). The output of one channel is negative with respect to ground, however, and the other positive. These outputs are then added and presented to the threshold corrector and slicer, which together form a sort of "decision" apparatus.

The purpose of the decision stages is to decide which type of pulse, mark or space, was received. When the snr is good, there is little problem. However, if the snr is low, there is great uncertainty as to whether a mark or space pulse was received. One of the larger random fluctuations in the noise could cause a mark pulse to be mistaken for a space pulse, or vice versa, and the printer would print an improper letter.

In making this decision, you are faced

with a problem. Unlike the signal, the noise voltage cannot be given a deterministic mathematical representation; however, its probability distribution *is* known (the probability that the noise voltage will not exceed any given value at a given time). During the interval in which you expect to receive a pulse, you observe a certain voltage at the output of the adder. Knowing the probability distribution of the adder output voltage (which can be calculated, although with difficulty), you can determine whether the voltage is more probably caused by a mark or space, and decide accordingly.

If the voltage is above a certain value, a space was probably sent; if below, a mark. This middle value, where the probabilities are equal, is the threshold voltage* and is determined by the threshold corrector.

For those readers with a mathematical turn of mind, the detector output voltage, when a signal is applied, has a Ricean probability density. Noise alone has a Rayleigh density. At any time, one channel has signal plus noise, the other has noise only, so the output of the adder will have the convolution of the Rice and Rayleigh densities as its probability density function.

If no independent fading of the mark and space signals occurred, both channels were of equal gain, and both signals of equal amplitude, the proper threshold voltage would be zero; that is, choose *mark* if the adder output is negative, and choose *space* if positive. There would be no need for the threshold corrector. However, the mark and space channels do undergo independent fading, so a nonzero threshold value is computed. The usual value of threshold voltage is chosen to be halfway between the average value of mark channel output voltage and the average space channel output voltage:

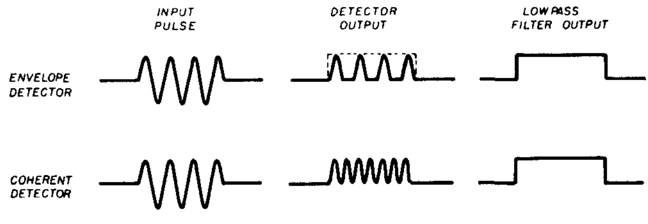

fig. 2. Operating characteristics of envelope detectors and coherent detectors.

table 1. Percent correct copy vs input signal-to-noise ratio.

	probability of error		percent correct copy	
input snr	coherent FSK	non-coherent FSK	coherent FSK	non-coherent FSK
0 dB	0.15	0.30	32%	8%
2 dB	0.10	0.25	48%	13%
3 dB	0.05	0.15	70%	32%
6 dB	0.01	0.10	93%	48%
9 dB	10^{-5}	0.01	99.9%	93%

$$V_{thresh} = \frac{\overline{V_{mark}} + \overline{V_{space}}}{2}$$

Remember, V_{mark} is negative and V_{space} is positive. The line above V_{mark} and V_{space} indicates an average value of voltage. This threshold voltage is at best a close approximation of the optimum, which is a complicated relation, practically impossible to implement electronically.

coherent demodulation

An FSK demodulator using coherent

*Not to be confused with threshold effects mentioned in connection with fm; these are entirely different phenomena.

detection is shown in **fig. 3**. The post-detection filters are identical to those used in the non-coherent demodulator. Instead of using an envelope detector, the received signal is multiplied in a product detector by a sinusoid of the same frequency *and* phase. Thus, the phase information in the signal, which is ignored by the envelope detector, is used in the coherent detector to improve the signal-to-noise ratio at the detector output.

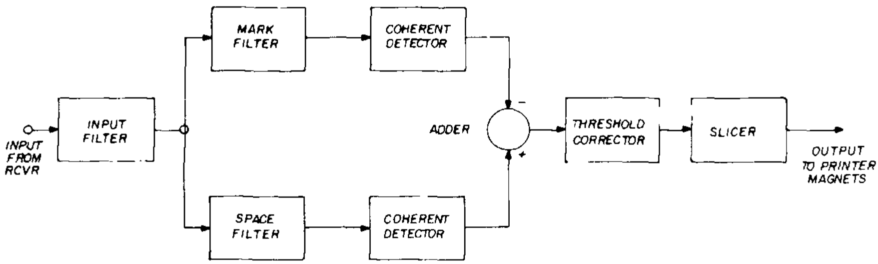

fig. 3. Block diagram of a coherent a-m RTTY demodulator.

The improvement over non-coherent FSK varies from 4 dB at low signal-to-noise ratios to 2 dB at high (greater than 6 dB) snr. This means that using coherent detection instead of noncoherent detection results in an improvement equivalent to an extra 4 dB of antenna gain or transmitter power. In terms of percent correct copy there is a noticeable improvement (see **table 1**). It is clear that coherent modulation results in readable copy in situations where the best non-coherent FSK systems would print gibberish.

An added advantage, not considered in calculating **table 1**, is that

$$V_{thresh} = \frac{V_{mark} + V_{space}}{2}$$

is precisely the optimum threshold for a coherent system. This arises from the fact that the probability density for the adder output voltage is a Gaussian function (*bell-shaped curve*), unlike that of the non-coherent system.

The disadvantage of coherent demodulation is that a signal identical in phase to the received signal must be generated. This task is best accomplished by a phase-locked loop. A detailed discussion of phase-locked loop operation is beyond the scope of this article (and beyond the scope of most communications theory textbooks, for that matter) but a simple description may be presented. You may consult the references for further information.

The phase-locked loop consists of a multiplier (product detector), lowpass filter, and voltage-controlled oscillator (vco)

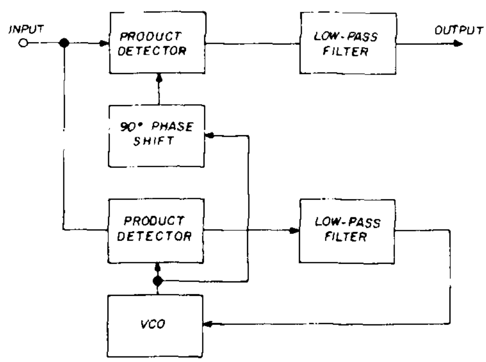

fig. 4. Coherent detector using a phase-locked loop.

(see **fig. 5**). The vco output and the input signal are applied to the product detector; its output is a beat frequency or, if the vco frequency and signal frequency are the same but the phases differ by other than 90°, a dc voltage. This voltage is applied to the vco which changes frequency by an amount proportional to the applied voltage until the signal and vco frequencies are the same and their phases are close to 90° apart. The result is that the vco output follows phase changes of the applied signal. The vco output, shifted in phase by 90°, is used for the

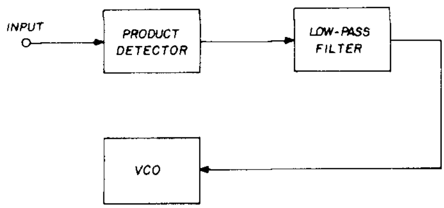

fig. 5. Basic phase-locked loop consists of a vco, product detector and lowpass filter.

heterodyning signal in the coherent detector.

The coherent detector requires complicated circuitry. However, high quality phase-locked loops ICs are available commercially, some with capabilities for a-m as well as fm demodulation. Hence, the entire detector and phase-locked loop can be had with a single integrated circuit. One such IC, available on the surplus market for about $3.60, is the Signetics NE561. Thus, for a total investment of about $7.00 over the cost of a conventional a-m demodulator, you can obtain the best possible RTTY demodulation.

transmitter modifications

In general, no transmitter modifications should be necessary. It is possible to build a phase-locked loop demodulator which will lock onto the mark or space pulse well within its 22-millisecond duration. However, during the time the loop is not locked, signal-to-noise ratio at the detector output is not optimum. This results in lowered performance.

To minimize locking time, the phase of the mark and space pulses should be as constant as possible. Therefore, instead of one oscillator switched between two tuned circuits, the transmitter should use two free-running oscillators, one to generate mark signals, the other for space signals. Keying is accomplished by switching between the outputs of the two oscillators. In this way the phase-locked loops will not have to re-lock onto each successive pulse. It should be emphasized, however, that the coherent system described here will work with conventional FSK transmitters with superior performance as compared to that of non-coherent reception.

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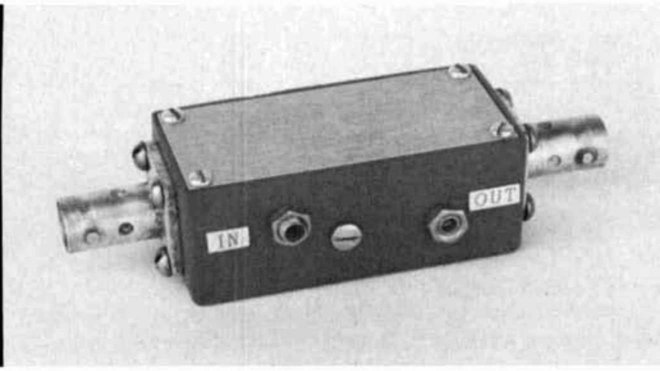

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Construction of
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that provides
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Ronald E. Guentzler, W8BBB, Route 1, Box 30, Ada, Ohio 45810

Recently I had need for a small, mechanically-stable, well-shielded, low-noise, two-meter preamplifier. The first three requirements were met by using a Pomona 2397 "Black Box." It is a cast-aluminium box measuring 2.2 x 2.8 x 5.7 cm (7/8 x 1-1/8 x 2-1/4 inches). Because of the small size, direct wiring rather than a printed circuit was used; this might partially account for the good noise figure obtained.

There are three important contributors to the noise figure (NF) of a preamp.

1. **The gain of the preamp and the NF of the stage following it.**[1] This preamp has 15-dB gain which is adequate so long as the unit it is feeding is reasonably good; too much gain between the antenna and the first mixer is bad.[2]

2. **The transistor.** The RCA 3N159 has a respectable NF rating.[3] It is selected by the manufacturer to have a maximum NF of 3.5 dB and a typical NF of 2.5 dB at 200 MHz; the performance will be better at 147 MHz. The 3N200 performs as well as the 3N159.

3. **The input circuit.** The best transistor is worthless unless a low-loss (and thus low-noise) input circuit is used. The best input circuit coupling arrangement is the "tapped capacitor" and the worst is link coupling.[2] A good coil must be used. This circuit uses the JFD LC374 which is a combination piston capacitor and an *air supported, silver plated* coil.

the circuit

The preamp circuit is shown in **fig. 1** and partial parts layouts are given in **fig. 2**. Capacitors C2 and C3 provide the input coupling and present the proper "turns ratio" for impedance transformation as well as providing most of the capacitance for resonating the input circuit; both should be silver micas. L1 and C1 are the JFD LC374. C1 is used mainly to physically support L1 and one transistor lead, but it also provides a means for adjusting the resonant frequency of the input circuit. If all is proper, the input circuit should resonate with C1 near to its minimum capacitance.

A dual-gate mosfet was used because it does not require neutralization (so long as adequate shielding, etc., are provided); this eliminated one coil and a lot of trouble. The 3N159 is the lowest noise

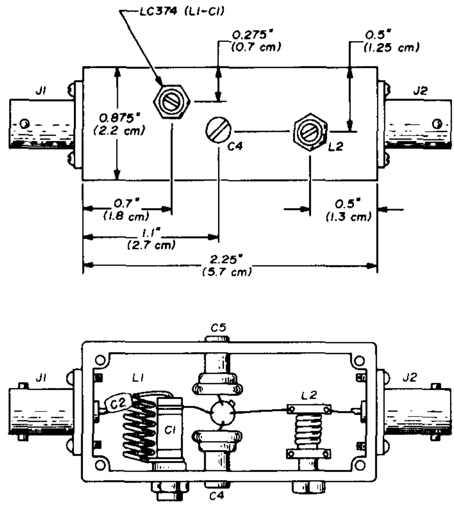
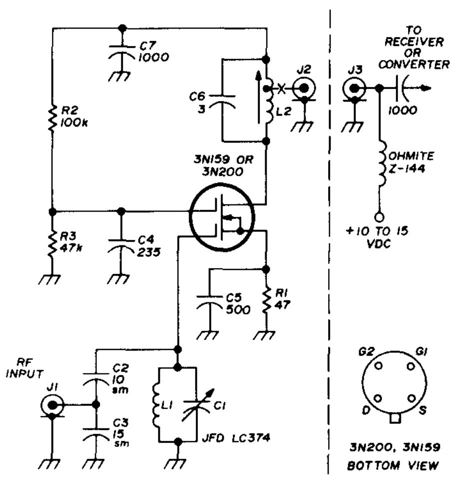


fig. 2. Permanently mounted component location. As viewed from above, the coil on the JFD LC374 is almost beneath its capacitor, C1. The transistor is located between the mica button capacitors. Make sure that the transistor will fit between the capacitors before mounting them.



C1 part of JFD LC374 tank circuit (see text)
C2 10-pF silver-mica with 3N159, 8-pF silver mica with 3N200
C3 15-pF silver-mica with 3N159, 12-pF silver mica with 3N200
C4 235-pF mica button
C5 500-pF mica button
L1 JFD LC374 tank circuit (contains C1)
L2 6 turns no. 22 enamelled on a 5-mm (0.2") diameter slug-tuned form, tap at 1 turn

fig. 1. The two meter preamp. Power is fed into the preamp through the rf output connector. On the right side is shown a power feed arrangement to be located inside the unit (receiver, converter, etc.) with which the preamp will be used.

RCA unprotected dual-gate mosfet but since it does *not* have internal gate protection, caution should be observed in handling the device. Of the protected RCA dual-gate mosfets, the 3N200 has the best NF specification. However, the 3N200 has higher input and output capacitances than the 3N159. Therefore, if a 3N200 is used in this circuit, use an 8-pF silver-mica for C2 and a 12-pF silver-mica for C3.*

construction

As can be seen in **fig. 2**, C4 and C5 are button-mica capacitors. They support two of the transistor leads. Their values may seem small, but these values were chosen to provide series resonance with the transistor leads.

The value of R1 was optimized to give minimum NF with 10-volts dc applied to the preamp. However, if other considera-

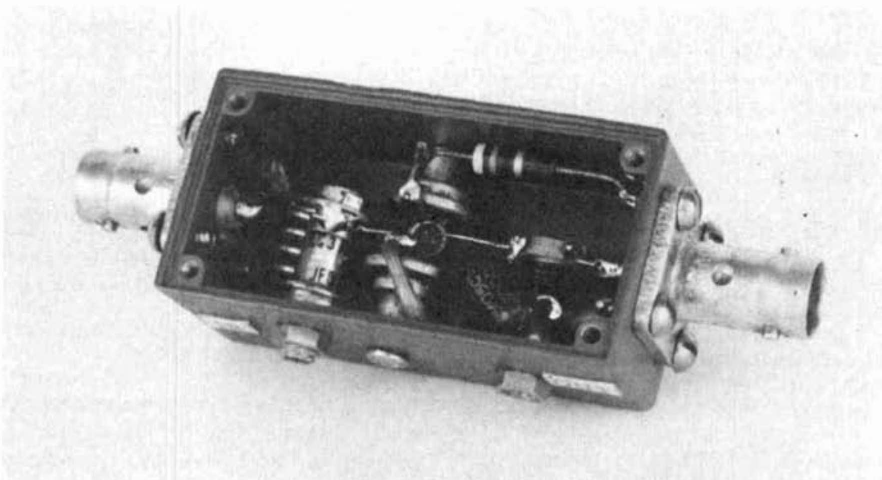
*The Texas Instruments 3N204 and 3N211 appear to be good alternates for the RCA 3N159 and 3N200. Although the 3N204 has a lower gain than the 3N211, it has a better noise figure. The author has not yet tried the TI transistors in the preamp. **Editor.**

tions are important, it can be increased in value.[3]

The output tank is a conventional slug-tuned coil, 5-mm (0.2-inch) diameter, with 6 turns number-22 enamelled wire, tapped 1 turn from the cold end. Depending upon the coil form and slug characteristics, C6 might have to be changed to obtain the proper resonant frequency.

alignment

Alignment is quite simple. Adjust C1 and L2 for maximum signal. Then adjust C1 for minimum noise figure. (Note that tuning can be accomplished from the outside with the cover on.) If noise measuring equipment is not available, the approximate minimum noise figure can be obtained by tuning C1 on the low



Construction of the low-noise two-meter preamp showing the component layout. Case is a small cast-aluminum box made by Pomona (see fig. 2).

frequency side of the resonance (run the piston into C1); tune either for approximately a 10% decrease in output signal or turn the piston into C1 by about one-half turn.

It was considered undesirable to provide a separate input for the dc feed because of the small physical size of the preamp. Therefore, dc power is fed through the output coax (this is desirable for antenna-mounted preamps). An arrangement for feeding the dc into the coax is shown on the right side of **fig. 1**. If direct dc feed is desired, C7 can be replaced by a feedthrough capacitor such as a Centralab FT-2300; then, a 1000-pF ceramic should be inserted in series with the output connector at the point indicated by an X.

To assure good contact between the coax connectors and the box, the threads in the connector flanges were drilled out, and the box was drilled and tapped for 4-40 machine screws. The screws projecting through the box provided a good means for obtaining ground points.

The good results were obtained by using the Pomona "Black Box" (actually, it is blue), the JFD tank, the dual-gate mosfet, and the tapped-capacitor input circuit. Don't cheat yourself—buy the proper components.

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optimum height for horizontal antennas

How to choose antenna height to put your signal where you want it to go

Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University

A recent magazine article[1] showed that the higher the better holds true for the gain of a horizontal dipole, and presented a graph showing improvements of over 20 dB with increased height. Since 20 dB is a pretty healthy gain, I decided a further consideration of this approach to increasing antenna gain was called for.

Any antenna book will show that the distant field strength from a horizontal dipole is proportional to sin (h sinα), where h is the electrical length and α is the radiation angle. This equation is, however, too simplified for many practical antenna comparisons. As is often the case, I fall back to Kraus' Antennas[2] and his equation (11-87) on page 305 is pertinent here. This equation gives the gain for a half-wave horizontal antenna above ground referenced to a half-wave antenna in free space. Thus it allows comparison to be made to a standard reference, as well as comparison of one real antenna to another.

The equation is:

$$G = \sqrt{\frac{R_{11} + R_{1L}}{R_{11} + R_{1L} - R_m}}\ |2 \sin (h_r \sin \alpha)|$$

R_{11} = $\lambda/2$ self resistance

R_{1L} = $\lambda/2$ loss resistance (assumed zero here and in Kraus' *Antennas*)

R_m = mutual resistance between $\lambda/2$ and its image 2h away

h_r = $(2\pi/\lambda)h$

Where R_{11} is 73 ohms at an infinite height, and $(R_{11} - R_m)$ is the antenna resistance at any given height, as shown in fig. 2-45 of the ARRL Antenna Handbook,[3] or Kraus, page 305.

determining factors

This equation shows that the difference in gain between two antennas of different height occurs at certain combinations of height and radiation angle, that the antenna resistance is an important term when the *usual* assumption of constant power is used, that electrical height should be used, and that the maximum gain of a real antenna over one in free space is 2 to 1.

How, then, can you get more than 6 dB gain when the *maximum gain over a free-space antenna is only 2 to 1*? It is a case of comparing apples to oranges. The 2 to 1 improvement figure refers to the comparison of the field strength of a certain height real-life horizontal dipole at the radiation angle giving maximum field strength to a similar one in free-space. The over-6 dB case involves comparing the field of one real antenna *to the field of another real antenna*, at a certain radiation angle. We will see, however, that a 20-dB improvement does not always result, depending upon what combination of α, $(R_{11} - R_m)$, and h_r is used.

table 1. Vertical-radiation from various height horizontal antennas, as compared to a half-wave dipole in free space (see fig. 1). The last column shows angle at which maximum radiation occurs.

antenna height (radians) h_r	antenna height (wavelength) $\frac{h}{\lambda}$	vertical radiation angle α 2°	α 10°	α 20°	α 30°	α 60°	α 90°	α max
36°	0.1	0.08	0.41	0.79	1.15	1.93	2.19	90°
90°	0.25	0.10	0.49	0.94	1.29	1.79	1.83	90°
108°	0.30	0.11	0.56	1.05	1.41	1.74	1.65	56.4°
126°	0.35	0.13	0.64	1.17	1.53	1.63	1.39	45.6°
180°	0.50	0.22	1.06	1.79	2.04	0.83	0	30.0°
270°	0.75	0.32	1.45	1.98	1.40	1.60	1.98	19.5°
360°	1.00	0.44	1.77	1.67	0	1.49	0.0	14.5°

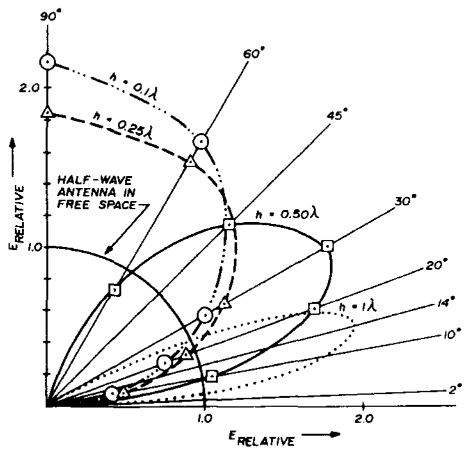

fig. 1. Relative field strength for horizontal half-wave antennas vs h in wavelengths as a function of radiation angle for constant power.

The *equation shows that at any height* between 0.1 and 0.25 wavelength the maximum field of a horizontal is always at a radiation angle of 90° (straight up!), although the low-angle radiation does increase at the 0.25 wave height. Above a quarter wave each height, h, has an angle where the maximum radiation occurs, as shown below:

h/λ	α
0.25	90°
0.30	56.4°
0.35	45.6°
0.50	30.0°
0.75	19.5°
1.00	14.5°

This shows why it is so hard to get good low-angle radiation with a horizontal dipole on 80 meters. This is also what got me interested in using verticals for

80-meter DX. Antenna height in wavelengths should be used with the formula, or at least the height in feet should be converted to wavelengths, as the results will then apply to any amateur band. A 10-foot (3-meter) high dipole works a lot differently at 80 meters than one at that height for the 20-meter band.

The equation, or better yet the principles upon which it is based, shows zero radiation at a radiation angle of zero degrees. This is because the antenna and image have currents equal in strength, but opposite in phase, so the field at a distance cancels at $\alpha = 0^\circ$. There is, however, radiation at angles for $\alpha > 0^\circ$. This radiation is plotted in **fig. 1**.

examples

To illustrate these points, consider some examples using an h of 0.1 wavelength as compared to one of 1 wavelength, at several radiation angles. The field from either is zero at $\alpha = 0^\circ$. Next use $\alpha = 3^\circ$.

The gain of an antenna 0.1 wavelength high compared to one in free-space at $\alpha = 3^\circ$ is:

$$G = \sqrt{\frac{73}{21}}\ |2 \sin (36^\circ \times \sin 3^\circ)|$$
$$= 1.86 \times 2 \times 0.0329 = 0.1226$$

For the 1-wavelength high antenna at $\alpha = 3^\circ$,

$$G = \sqrt{\frac{73}{73}}\ |2 \sin (360 \times \sin 3^\circ)|$$
$$= 1 \times 2 \times 0.3229 = 0.6459$$

The gain of the 1-wavelength high antenna over the 0.1-wavelength height is (0.6459/0.1226) = 5.27, or 14.4 dB. This is about the best improvement that I calculated at this radiation angle. Other 10 to 1 height ratios gave the following gains:

0.5/0.05	10 dB
0.35/0.035	3 dB
0.6/0.06	11 dB

It is possible that there could be 20-dB improvement at smaller radiation angles, but the actual field-strength values would be quite small at such low angles. Let's try $\alpha = 14.5^\circ$, where the 1-wavelength antenna height has radiation maximum.

$$G \text{ for } h = 1\lambda \text{ and } \alpha = 14.5^\circ \text{ is } 2.0$$
$$G \text{ for } h = 0.1\lambda \text{ and } \alpha = 14.5^\circ \text{ is } 0.5828$$
$$G\ (1\lambda/0.1\lambda) = 10.7 \text{ dB}$$

At 30°, the 1 wavelength high antenna has no radiation, so the 0.1 wavelength height would be a better choice and 0.5 wavelength would be even better since its maximum radiation occurs at $\alpha = 30^\circ$. Any number of examples could be worked out, but all of these appear in **fig. 1**. Using it or the data in **table 1**, you should be able to select the optimum antenna height for the radiation angle of greatest interest to you.

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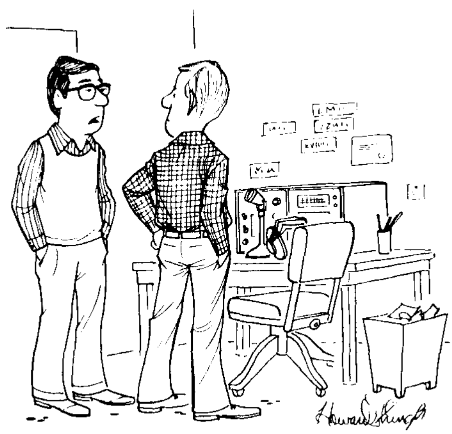

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local-oscillator waveform

effects on spurious mixer responses

Tests with doubly-balanced IC mixers and dual-gate mosfet mixers indicate no detrimental affects with square-wave injection signals

Max Robinson, K4ODS, John W. Smith, Route 8, Box 105, Bowling Green, Kentucky 42101

Up until now receiver local oscillators have been crystal or LC controlled and the waveform was of no particular significance. The output of these oscillators was a fairly good sine wave. Everyone accepted, without question, that the local oscillator output should be *clean*. Current receiver designs require that the first local oscillator be crystal controlled at 500 kHz intervals throughout the high-frequency spectrum. The frequency synthesizer seems to be a natural for this but it can introduce some problems.

The output of a synthesizer is often a square wave, and in the usual tunable i-f receiver, the use of a tuned circuit to clean up the waveform could become very complicated. The question has arisen in my mind, can I get away with using a square-wave local oscillator?

I feel, as do most hams, that one good test circuit is worth a thousand words of theoretical discussion. I decided to try two different mixer circuits. The first was an integrated circuit doubly-balanced mixer, the Signetics N5596, which is equivalent to the Motorola MC1596.[1] The other circuit I tried was a dual-gate mosfet, the MPF122. The reason for using these circuits was that I planned to

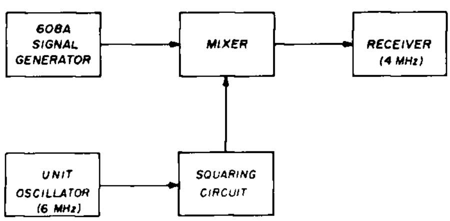

fig. 1. Test setup for comparing performance of N5596 IC mixer and dual-gate mosfet mixer with square-wave local oscillators.

use them in some projects which are now in the planning stage.

test procedure

The source of sine-wave local oscillator was a homebrew signal generator built around a General Radio Unit Oscillator 10 MHz to obtain a beat with the fundamental of the local oscillator, and the signal level was adjusted for a reading of 20 dB over S9 at the receiver. Next, the 608A signal generator was set to obtain a beat with the harmonics of the local oscillator injection frequency and

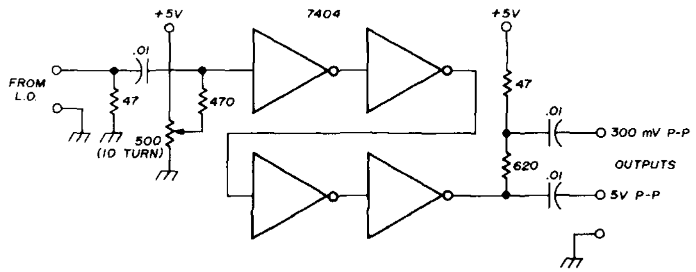

fig. 2. Simple IC circuit for generating a square-wave output from the sine-wave generator.

(**fig. 1**). The square-wave source was derived from the sine-wave source using the circuit shown in **fig. 2**. The 10-turn pot is a symmetry adjustment and in all cases was adjusted for minimum second harmonic response of the mixer circuit under test. The signal source was a Hewlett-Packard 608A signal generator which has a minimum frequency of 10 MHz. The receiver used was a Drake 2B.

In all tests, the local oscillator was set to 6 MHz and the receiver was set to 4 MHz. The signal generator was first set at the 608A. The attenuator was reset to maintain the former S-meter reading at the receiver.

For example, the third-harmonic spurious response of the mixer was checked by setting the 608A at 14 MHz to beat with the third harmonic of the injection frequency. Then the signal level was increased until the S-meter once again read 20 dB over S9. If the signal level required was at or above the overload level, the data point was thrown out.

The N5596 IC was connected as

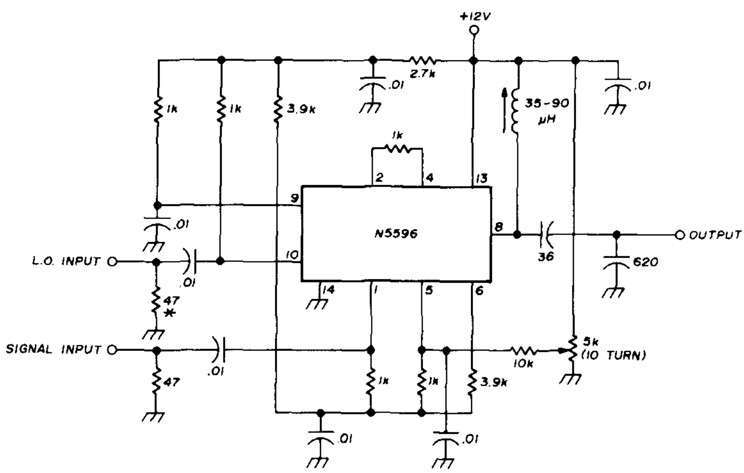

fig. 3. High-frequency mixer circuit using the Signetics N5596. Resistor marked with an asterisk is removed when using a square-wave local-oscillator.

shown in **fig. 3.** The balance control was adjusted by setting the local oscillator to 4 MHz and adjusting for minimum carrier feedthrough. **Table 1** shows the results of the tests. The spurious responses are given in dB relative to the fundamental response at 10 MHz. The injection voltages given in the column headings are according to data supplied by Signetics.[2]

In the table, the notation *OL* means that an overload signal level was required to produce a 20 dB over S9 meter reading. The note *NM* means that no measurement was taken at this frequency.

The dual-gate mosfet mixer is shown in **fig. 4.** I did not have very much data on the MPF122 so I guessed that the output impedance would be about the same as the N5596. The value of the source resistor comes from a page of the information supplied by Circuit Specialists Company. Gate two was placed at dc ground because this gave maximum conversion gain.

The chosen injection voltage of 1-volt rms seemed to be optimum. The 5-volt p-p square wave was chosen to maintain the same sine-wave voltage to square-wave voltage ratio as that used for the N5596 IC.

As a final check, the square wave was viewed on an oscilloscope and it was confirmed that the minimum second harmonic response really did occur when the square-wave was symmetrical. The results of testing the MPF122 mixer are given in **table 2.**

conclusions

One of the more obvious conclusions is that the local oscillator should be operated above the incoming signal. For the frequencies used here, the received frequency would be 2 MHz. The closest spurious signal is at 8 MHz which could be attenuated by a preselector of good design. On the other hand, if the 10-MHz signal were used, the 8-MHz spurious could be rather troublesome.

In both mixer circuits, using a square-wave injection actually reduces the second harmonic product. If the preselector has very good ultimate attenuation, the use of square-wave could be advantageous.

The harmonics of very high order are much stronger with a square wave than with a sine wave and although this may not cause any trouble in a receiver, the presence of vhf spurious signals in a transceiver or transmitter could make TVI reduction more difficult. If a premixer and fixed first i-f are used, the first mixer injection waveform can be easily cleaned up by simply adding another gang to the preselector.

A type-D flip-flop might be a useful premixer because its output is only the

table 1. Harmonic response of N5596 IC balanced mixer in dB below fundamental response.

harmonic number	frequency (MHz)	dB with 60 mV rms sine-wave injection	dB with 300 mV p-p square-wave injection
1	2	NM	NM
1	10	0	0
2	8	NM	NM
2	16	-40	-44
3	14	-19	- 9
3	22	-17	- 9
4	20	OL	-41
4	28	OL	-48
5	26	-34	-12
5	34	-33	-10
7	38	OL	-14
9	50	OL	-14
11	62	OL	-13
13	74	OL	-12
15	86	OL	- 7

table 2. Harmonic response of dual-gate MPF122 mosfet mixer in dB below fundamental response.

harmonic number	frequency (MHz)	dB with 1-V rms sine-wave injection in	dB with 5-V p-p square-wave injection
1	2	NM	NM
1	10	0	0
2	8	NM	NM
2	16	-15	-47
3	14	-12	-10
3	22	-13	- 9
4	20	-23	-41
4	28	-22	-43
5	26	-23	-13
5	34	-22	-12
7	38	NM	-18
9	50	NM	-15
11	62	NM	-21
13	74	NM	-15
15	86	NM	-10

difference frequency (plus odd harmonics) with no sum or original frequencies coming through.

afterthoughts

The article by Moore[3] arrived too late to be factored into the preparation of this article. He points out that one manufacturer of professional receivers uses square-wave injection voltage to the mixer. This adds relevance to the work done in this article.

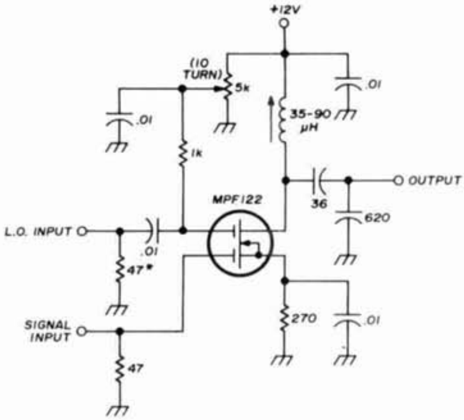

fig. 4. High-frequency mixer circuit using an MPF122 dual-gate mosfet. Resistor marked with an asterisk is removed from the circuit when using a square-wave local oscillator.

I did notice that the input voltage required to saturate the mixer was slightly higher when a square wave was being used as the local oscillator. This was true for both circuits. In general it seems to be possible to replace a double handful of crystals with a frequency synthesizer if some care is taken with the overall receiver design.

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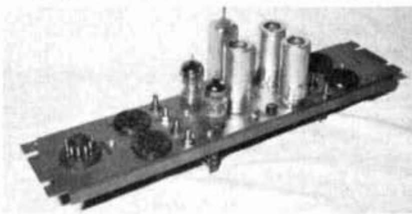

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understanding spectrum analyzers

Spectrum analyzers are very useful in radio communications — here's how they work and how to use them

Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75240

Spectrum analyzers are instruments which display signal amplitude versus frequency on a cathode-ray tube (CRT). **Fig. 1** shows how a typical spectrum analyzer CRT display might appear. The three vertical responses are produced by three separate input frequencies, and the jagged nature of the base line is caused by the system noise. CW signals appear as vertical lines, and modulated signals will show sidebands.

The Heathkit SB-620 Scanalyzer is a special-purpose spectrum analyzer which is used in conjunction with a communications receiver. It displays all signals whose frequencies are within a few hundred kHz of the frequency to which the receiver is tuned. Thus the operator may visually observe the activity on the band and see where the strong signals and clear frequencies are located without having to tune his receiver.

operation

Fig. 2 is a simplified block diagram of a spectrum analyzer. It is nothing more than a receiver whose frequency is swept across a certain band, and whose output causes a vertical deflection on the CRT for each input signal encountered.

The input network may vary according to system requirements. It may contain a filter to limit the input signals to the desired frequency range, and thus prevent image responses. An amplifier may also be included to improve sensitivity.

The mixer produces an i-f output whose frequency is the difference between the input frequency and the local oscillator frequency. This output is amplified by the i-f amplifier, detected, and fed to the vertical deflection plates of the CRT.

As is the case with an ordinary receiver, the frequency of the spectrum analyzer is controlled by varying the frequency of the local oscillator (LO). Since the spectrum analyzer must automatically and continuously sweep across a band of frequencies, the LO must be voltage controlled so that its frequency may be varied by changing its control voltage.

A sawtooth waveform generator provides the control voltage for the LO and the sweep signal for the horizontal deflection plates of the CRT. This arrangement causes the electron beam in the CRT to move horizontally across the face of the CRT as the frequency of the LO changes. Thus the horizontal position of a vertical response on the CRT is directly related to frequency. For good accuracy it is important that the sawtooth waveform be linear, and that the relationship of control voltage to LO frequency also be linear.

As an example of how the spectrum analyzer performs, assume the input frequency range is 3 to 8 MHz, and the center frequency of the i-f amplifier is 9 MHz. The LO must sweep from 12 to 17 MHz to produce a difference frequency of 9 MHz between the LO and input. If an input signal exists at 4 MHz, an i-f signal will be produced when the LO frequency sweeps through 13 MHz.

i-f bandwidth

It is desirable that a CW signal appear as a straight vertical line on the CRT. To approach this condition, however, the i-f bandwidth must be very small compared to the frequency range swept by the spectrum analyzer. In the example above, the horizontal scale of the CRT would cover 3 to 8 MHz, a sweep width of 5 MHz. In this case, an i-f bandwidth of 5 kHz would give good resolution because it represents only 0.001 of the total horizontal scale. The CRT response to a CW signal would be quite narrow, being only 0.001 of the total horizontal scale at the 3-dB points.

If, however, the spectrum analyzer is adjusted so that the total frequency range swept is from 3.00 MHz to 3.05 MHz (a sweep width of 50 kHz), then a 5 kHz i-f

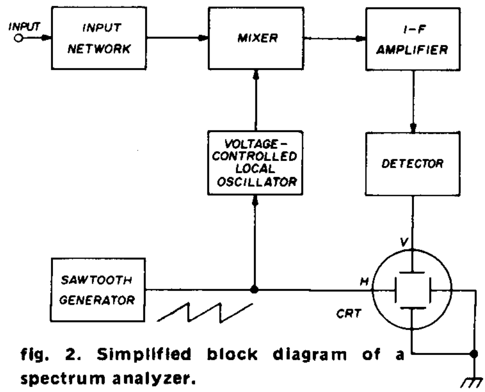


fig. 2. Simplified block diagram of a spectrum analyzer.

bandwidth would be only one-tenth of the horizontal scale. This would produce a poor display such as that shown in **fig. 3,** and the display would show the frequency response of the i-f passband rather than the desired vertical line. This illustrates why the ratio of sweep width to i-f bandwidth should have a value of at least 100, and preferably higher.

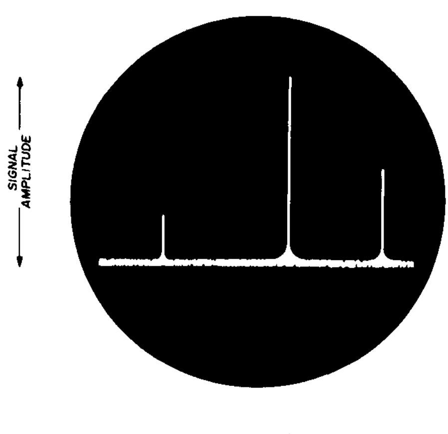


fig. 1. Typical spectrum analyzer display (CW signals).

sweep speed

As the i-f bandwidth is decreased, it lowers the limit on how fast the frequency range may be swept. This is because the rise time of the i-f amplifier increases as its bandwidth is decreased. If the spectrum analyzer sweeps by a signal too rapidly, the i-f amplifier won't have time to respond to it.

Another problem encountered when the i-f amplifier frequency response has steep skirts is that ringing will occur if the sweep speed is too fast. This will produce a distorted display such as that shown in **fig. 4.** It can be shown mathematically that, if the i-f passband is determined by a single tuned circuit, the minimum i-f bandwidth should be

$$\text{i-f } BW_{min} \geqslant 1.18 \sqrt{\frac{\text{sweep width}}{\text{sweep time}}} \quad (1)$$

This is seldom if ever the case, because a single tuned circuit would not provide sufficient selectivity.

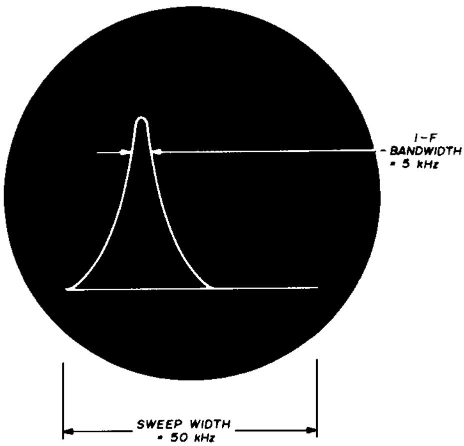


fig. 3. Poor resolution of the spectrum analyzer display occurs when the i-f bandwidth of the instrument is too large a fraction of the sweep width.

It has been determined empirically that if the i-f passband has a nearly ideal rectangular shape, the minimum i-f bandwidth should be

$$\text{i-f } BW_{min} \geqslant 10 \sqrt{\frac{\text{sweep width}}{\text{sweep time}}} \quad (2)$$

As an example, assume the horizontal scale of a spectrum analyzer CRT is to be 500 kHz wide, and the time of one sweep is to be 0.01 second (this corresponds to a sawtooth frequency of 100 Hz). The minimum i-f bandwidth is found from **eq. 2** to be

$$\text{i-f BW} = 10 \sqrt{\frac{500{,}000}{0.01}} = 70.7 \text{ kHz} \quad (3)$$

Obviously the i-f bandwidth is too wide to give a satisfactory display, because it is 14 percent of the total horizontal scale. To obtain a reasonable display, the sweep speed must be decreased to accommodate a suitable i-f bandwidth. Assuming that a 5-kHz i-f bandwidth will be narrow enough, **eq. 2** may be rearranged and solved for sweep time.

$$\text{sweep time} = \frac{100 \text{ (sweep width)}}{(\text{i-f BW})^2} \quad (4)$$

$$= \frac{100(500{,}000)}{(5000)^2}$$

$$= 2 \text{ seconds}$$

This corresponds to a sawtooth frequency of 0.5 Hz. With such a slow sweep speed, a CRT having long persistence phosphor becomes desirable. It should be remembered that somewhat faster sweep speeds may be used as the i-f passband skirts become less steep.

conclusion

This brief article only barely introduces you to the subject of spectrum analyzers and some of the more important design constraints and criteria, but it is hoped that it will assist those of you who are interested in designing, building or modifying these versatile devices. Spectrum analyzers exist in the form of complete instruments and as attachments to be used with oscilloscopes. In the latter case, the oscilloscope should have a direct-coupled (dc frequency response) horizontal channel to preclude distortion of a low-frequency sawtooth waveform before it reaches the horizontal deflection plates of the CRT.

ham radio

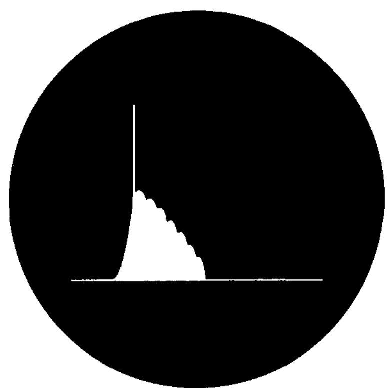


fig. 4. Spectrum analyzer ringing, shown here, is due to excessive sweep speed.

adding private-line to the Heathkit HW-202

The Heathkit HW-202 fm transceiver presents special problems if you want to install private-line — here's how to solve them

Arthur Reis, WA8AWJ, 603 N. Court Street, Howell, Michigan 48843

When you live in an area where low-frequency private-line tone is the key to the local fm repeaters, a new transceiver is a challenge. It's not always easy to find the right place to insert the low-audio tone in the existing circuit without creating additional problems. However, amateurs who live in "private-line" areas know that it must be done. And, as more and more repeaters go to private-line, as some have predicted, the challenge will become more widespread.

Recently, Heath introduced its HW-202 two-meter fm transceiver. It's an excellent rig, but it presents some unusual problems to private-line installations. This article shows how to solve these problems.

the circuit

In most fm rigs, the normal practice is to put the private-line circuit (usually a reed vibrator or twin-T circuit) in the modulation chain, *after* the clipping, limiting and deviation adjustment circuits. This is often accomplished by running the output of the private-line board through an appropriate dropping resistor, to the output arm of the deviation control, which is usually only one or two components removed from the modulation diode or transistor. This keeps private-line level independent of any other modulation.

Unfortunately, the Heath HW-202 design doesn't allow for installation of the private-line output at that point. Heath follows up the output of their modulator (Q202-Q204) with an RC filter chain of high series resistance (R248, R252-R255, C253-C257) (see **fig. 1**). A 4700-ohm resistor (R256) connects this chain to the modulation diode, D207. Unhappily, any bypassing on the output side of the private-line board (mine is a Motorola CTS) represents a comparative short to ground for the modulator output. Deviation drops to zero. It's obvious that another installation point had to be located.

The point finally chosen was more convenient than you might think. Heath inserts the audio to the cathode end of the modulation diode. However, since the diode is not directly grounded, it is possible to modulate either or both sides of the diode at once. Since it was impossible to modulate the cathode in this case, the private-line was inserted on the anode side. **Fig. 1** shows the exact point of insertion (junction of L201, R219, R218, and C236). I used a Motorola CTS Board in my rig, which provides 0.5-volt rms output and requires no series resistance of its own for approximately 500-hertz deviation.

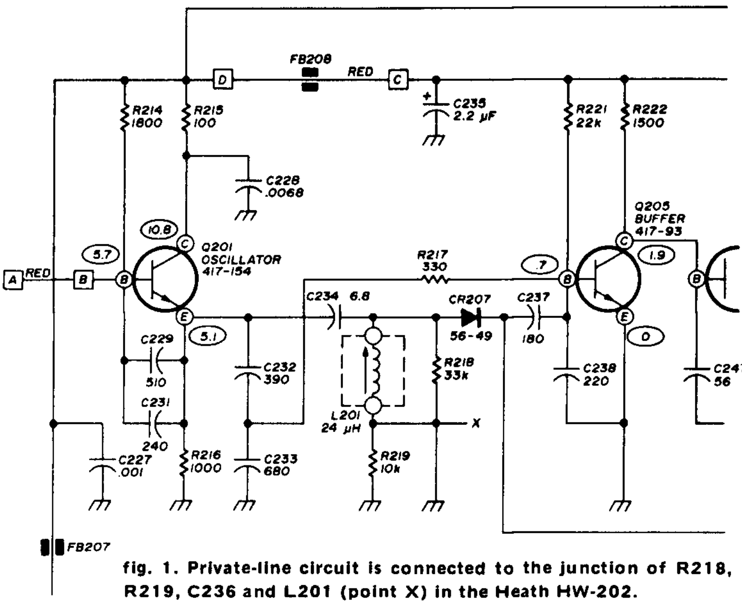


fig. 1. Private-line circuit is connected to the junction of R218, R219, C236 and L201 (point X) in the Heath HW-202.

installation

Mechanical installation was rather simple. There are no holes to drill if the private-line board is installed in an enclosed space with foam surrounding it. Heath made it especially convenient by putting an extra hole in the transmitter circuit board printed-circuit foil at the junction of L201, R218, R219 and C236. **Fig. 2** shows the position of the unused hole on the circuit board.

The output of the private-line board is run through *shielded* wire to that hole. The shield is grounded to the underside of the board at the point marked X *(fig. 2)*. Voltage for the board can be taken from any 13.8-volt point in the unit. I personally chose taking the voltage off the hot side of the pilot lamp on the regulator/hash-filter circuit board.

If hash filtering is needed for the private-line board, as it was in my case, a 220-ohm series resistor, followed by a 500-µF capacitor to ground, will do the job. It may also be advisable to install a switch for the B+ supply to disable the private-line circuit when you're traveling out of town. At least a few repeaters in the Midwest will reject signals containing private-line tones. Instead of defacing your unit, it may be better to replace the present 2000-ohm squelch control with a push-pull switch type. I set mine to turn off the private-line in the "pulled" position. The squelch function is, of course, not affected.

The private-line unit is wrapped in electrical tape and mounted in foam

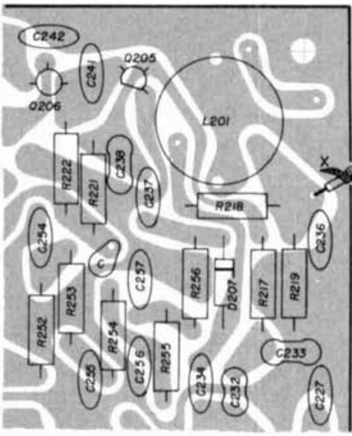


fig. 2. Upper right-hand portion of HW-202 transmitter printed circuit board. Center conductor of shielded cable from the private-line circuit is connected to a hole already in the circuit pad. The shield is connected to the grounded circuit trace.

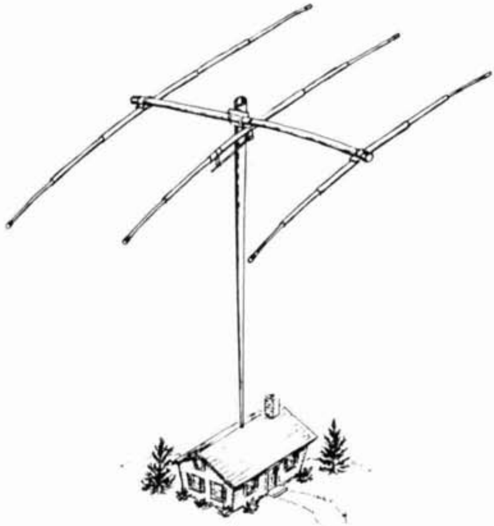
rubber behind the microphone input. Do *not* put the reed near the speaker magnet – reeds and magnets don't mix! The foam rubber is a good shock mount and eliminates the need for any holes in the HW-202.

summary

Performance of my unit is excellent. Private-line level is within one dB of nominal (550-Hz deviation) and there is no interference to the other forms of modulation used in the HW-202 (tone burst and touch-tone, as well as voice).

In conclusion, I would like to especially thank Keith Peterson, W8SDZ, member of the DART Repeater in Detroit, without whose help this project would not have been nearly as successful as it was.

ham radio



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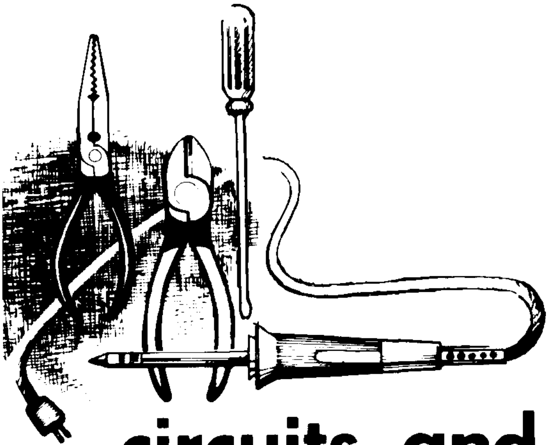

circuits and techniques

ed noll, W3FQJ

the dipole beam

Dreaming about beams makes no contacts. Commercial beams can be both expensive and troublesome. Don't forget that the dipole is a beam antenna, albeit bi-directional. It is not unidirectional like a beam but, then again, side rejection might be considerably better than that obtainable from a poor three-band beam. Often QRM off to the side is more the culprit than signal pick-up from the rear. Furthermore, the dipole provides a direct match to 50- or 70-ohm coaxial cable so you don't have to fiddle with any sort of matching arrangement. Matching arrangements restrict bandwidth and introduce loss.

A dipole for 10, 15 or 20 meters is a simple lightweight affair that can be rotated by arm-power or the smallest of TV antenna rotors. Your basic support can be a 10 x 6-inch (25.4 x 15.2-cm) piece of 5/8-inch (16-mm) plastic, **fig. 1.** Two pairs of U-bolts can be used to hold the two pieces of inexpensive aluminum tubing to the plastic. Another pair of heavier U-bolts holds the antenna to the mast.

Do you need the gain of a large beam? It can be useful for chasing down rare ones. On 10 or 15 meters the band is well open in a given direction or it is closed to sensible communications. A 1 or even 2 to 3 S-unit differential is unimportant in the majority of short DX contacts. The differential is even less important in stateside contacts. Things are tougher on 20 meters but this band is tough for everyone, even those with beams bursting with watts.

Then there is the matter of wind resistance and pocketbook load. The cost per foot of putting your dipole at a specified height is substantially less than that required to put a beam at the same altitude. The dipole is less likely to come down and, if it does, the financial loss or trouble are less than the catastrophe of a disjointed beam.

A split dipole arrangement can be used on more than one band through the use of a line tuner. My preference is for the T-network type (refer to the January, 1973, issue of *ham radio*, page 59). I have used one such 15-meter dipole for almost a year and have worked out pretty much where I wished to on 10 and 15 meters. The radiation pattern is still bidirectional, a standard figure-eight on 15 meters, extended and sharper figure-eight on 10 and a broad figure-eight pattern on 20. If a 20-meter dipole is used, the standard figure-eight pattern is obtained on 20 while a reasonable figure-eight pattern persists on 15. On 10 meters the pattern becomes a lobed affair.

hanging in there

If your dipole is mounted in an accessible location, such as immediately over a chimney or roof top, short sections of insulated wire can be hung on the end of a 15-meter dipole whenever you wish to

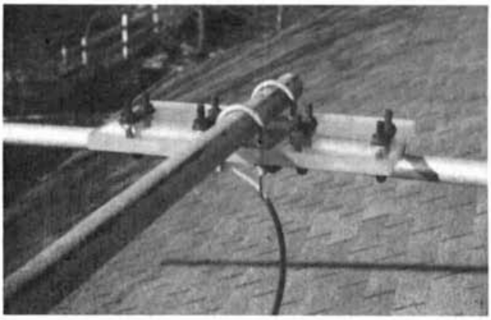

fig. 1. Basic construction of a rotatable dipole element showing the use of U-bolts and a scrap piece of plastic.

operate on 20 meters, **fig. 2**. This simple trick permits 20-meter dipole operation in a tight location and where there is inadequate turning radius (attic mount) for a full-length 20-meter dipole. A better plan is to have a pair of extensions (about 6-feet or 2-meters long) that can be telescoped into the dipole ends.

15-meter two-element

The same basic construction can be used for a two-element beam, **fig. 3**. If the reflector is spaced one-quarter wavelength in back of the dipole there is a minimal reduction in antenna impedance. Again, the T-network line tuner is a fine crutch if your transmitter is anemic and requires a perfect match (one of my transmitters is so inclined).

A line tuner permits this same antenna to function very well on 10 meters. The 15-meter reflector is not too long to provide some gain and directivity on 10. You can also load it on 20 meters with a tuner. If readily accessible, telescoping or hanging end pieces can be attached to the 15-meter dipole to resonate it on 20 meters, **fig. 4**. In this case the 15-meter reflector does some 20-meter *directoring*. Ingenuity is a fun substitute for high costs.

matching emancipation

The matching dispute has two avenues of thought. There is the complete-match phobia and the other. Personally I pursue the other. In most amateur practice antenna-to-line mismatches up to 3-to-1 have little effect on antenna system performance.

Where the mismatch hurts is at the transmitter. This is a matter of transmitter design and is a form of technical enslavement that hampers amateur antenna experimentation. Transmitter mismatches result in loss of output, rapid aging of the output stage, introduction of distortion components (sometimes) and other defects that could be avoided with the design of a wide-impedance output matching system.

In summary, the major ill effects of mismatched loads are a transmitter limitation and not one of antenna performance. The average amateur can overcome this limitation with the use of a tuner. The

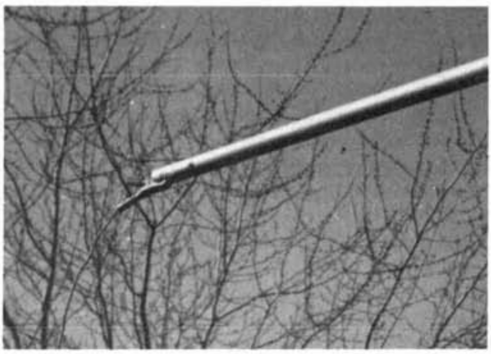

fig. 2. To add 20-meter operation to your 15-meter dipole simply clip an extension wire on each side.

manufacturer can overcome the problem with more versatile design or by including a tuner as part of the transmitter. In modern operation a tuner is not a bad idea at all because of its ability to suppress harmonics and other spurious radiation.

This problem has become very apparent to me because of the inquiries I have received with regard to matching the triangle antenna which has become increasingly popular on 40 and 80 meters

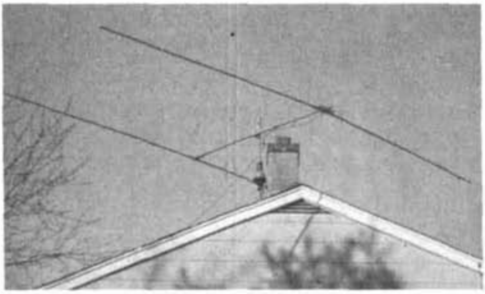

fig. 3. Simple arrangement for adding a reflector to a 15-meter dipole.

and is used on 20 to 160 meters as well. The question usually evolves around obtaining an exact match using a single triangle-driven element. The problem is more prevalent on 20 and 40 meters than on 80 and 160. On the latter two bands the average height of the configuration in terms of wavelength and feedpoint above ground is such that the impedance is inherently low once the antenna is resonated.

In the usual forty-meter installation the swr usually falls somewhere between 1.5:1 and 2.5:1. Such a mismatch has no significant effect on the performance of a triangle antenna but is a matter of concern to those worried about swr and possible influence on their transmitter. The swr at the transmitter can be reduced with the use of a 4-to-1 balun and experimentation with the overall length of the transmission line. The ultimate answer is the use of an antenna tuner.

The mismatch problem is a general one with full-wavelength closed antennas such as quad, delta and triangle. Tri-band beams can be an agonizing experience for the amateur striving for that perfect match although antenna performance itself is little affected by what is normally considered a serious mismatch.

In most amateur high-frequency antenna installations the practical length of the transmission line is such that very little loss is generated by the mismatch. Using high-quality low-loss coaxial line or open-wire types it is an insignificant quality.

The full-wave closed antenna performs well at low mounting heights, can be readily positioned and shaped to fit the mounting site and provides good performance despite necessary physical distortions away from straight-line mounting (see the May, 1973, issue of *ham radio*, page 66).

add 160 to your 80-meter inverted dipole

Many backyards that can handle an 80-meter inverted dipole are too confining for the construction of a full-size 160-meter antenna. However, by accepting some distortions away from the straight-on construction you can add 160-meter performance to your 80-meter antenna. The ends of your inverted dipole are low and in most situations it is practical to clip add-on sections that will provide 160-meter resonance. These can

fig. 4. Telescoping section converts a 15-meter dipole to 20 meters. On 20 meters the 15-meter reflector acts as a director.

be clipped on whenever 160-meter operation is desired, producing a very fine performing antenna. They can be run straight away, keeping them 7 feet (2 meters) or more above ground to permit pedestrian traffic. However, resonance can also be obtained by running these extensions at various angles to meet your property line, **fig. 5**. I have operated one successfully on 160 meters using the arrangement of **fig. 5B**.

To calculate the length of the extensions use the quarter-wavelength

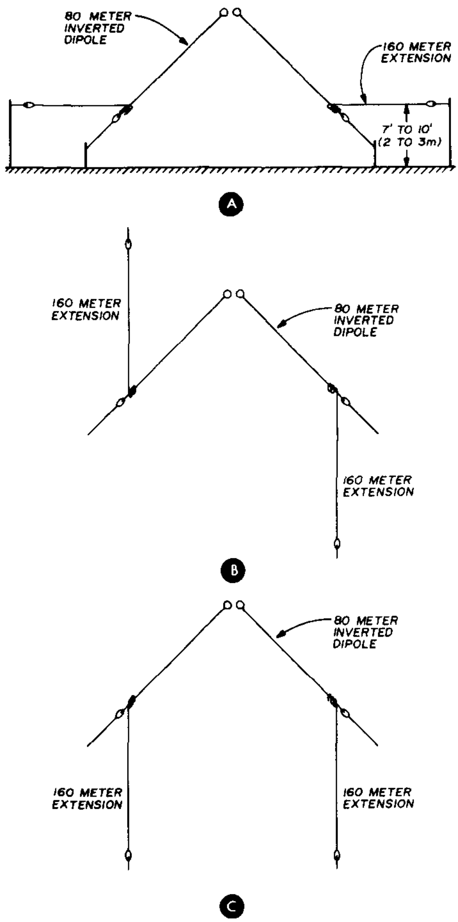

fig. 5. How to add 160-meter extensions to your 80-meter inverted dipole (see text for calculating extension length).

equation, substracting from it the leg length of your 80-meter inverted dipole.

$$L_{ext} = \frac{234}{F_{160}} - L_{80}$$

where L_{ext} is the length of the extension in feet, F_{160} is the desired operating frequency on 160 meters and L_{80} is the length of the 80-meter inverted antenna. This calculation may give you a longer overall length than required, but you can trim back both ends to establish resonance on the desired 160-meter frequency. Usually the extension needs to be cut back further as you fold the legs away from the straight-away plane of your 80-meter inverted dipole.

Using a T-network tuner I was able to load this antenna on every band, 10 through 160 meters. The performance of the configuration of **fig. 5C** was interesting in that it provided some additional directivity for operation on 10, 15 and 20.

ham-metrics

The English-speaking radio amateur has been talking metrics since the "below 200-meters" days. However, in calculating antenna length amateurs make the unnatural conversion to feet using the appropriate constants instead of going directly to antenna length in meters. The purchase of a meter stick or meter rule provides an easy introduction to the metric system when used in conjunction with antenna calculation. Typically, meter rules come in lengths of 10, 25 or 50 meters.

What is the wavelength of a 4-MHz wave? As you know, wavelength equals the quotient of propagation velocity (300,000,000 meters/second) over the frequency in hertz. Therefore:

$$L_{meters} = \frac{300 \times 10^6}{4 \times 10^6} = 75 \textit{ meters}$$

What is a half-wavelength at this frequency? A quarter wavelength?

$$\lambda/2 = \frac{75}{2} = 37.5 \text{ meters}$$

$$\lambda/4 = \frac{75}{4} = 18.75 \text{ meters}$$

What is the length of each quarter-wave segment of a half-wavelength dipole cut for 4 MHz considering an "end effect" of 4%?

$$\lambda/4 = 18.75 \times 0.96 = 18 \text{ meters}$$

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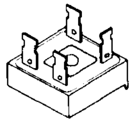

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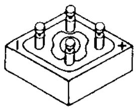

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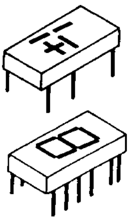

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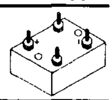

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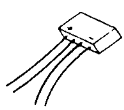

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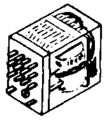

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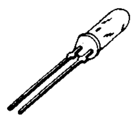

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basic equations

Three useful equations for making free-space wavelength calculations and one for calculating the length of one segment of a half-wave dipole are as follows:

$$\text{Length (free-space full wave)} = \frac{300}{f_{MHz}} \text{ meters}$$

$$\text{Half wavelength} = \frac{150}{f_{MHz}} \text{ meters}$$

$$\text{Quarter wavelength} = \frac{75}{\text{MHz}} \text{ meters}$$

$$\lambda/4 \text{ dipole element} = \frac{72}{f_{MHz}} \text{ meters}$$

For example, to find the length of a quarter-wave segment of a half-wavelength, 80-meter dipole for use on 3.6 MHz, the calculation is as simple as this:

$$\lambda/4 = \frac{72}{3.6} = 20 \text{ meters}$$

Get out your meter rule and cut your dipole. If, out of curiosity, you may want to know how long your antenna is in feet, you can use a conversion factor to determine this length. Why bother? You have already cut your antenna to resonate at 3.6 MHz. Using the metric system is not so much accepting the new as it is throwing away the old.

Recall the conversion you must make to inches when your answer is in decimal parts of a foot? No such foolishness is required with the metric system. Just remember that there are 10 decimeters in one meter and 10 centimeters in a decimeter (100 centimeters in one meter). All of which can be read directly from meter rules of various types. For example, calculate the length of each segment of a half-wave dipole for operation on 14060 kHz.

$$\lambda/4 = \frac{72}{14.06} = 5.12 \text{ meters}$$

Measure off two lengths of 5 meters, 12 centimeters and you have your 20-meter dipole.

ham radio

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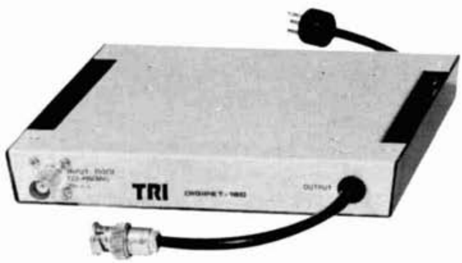

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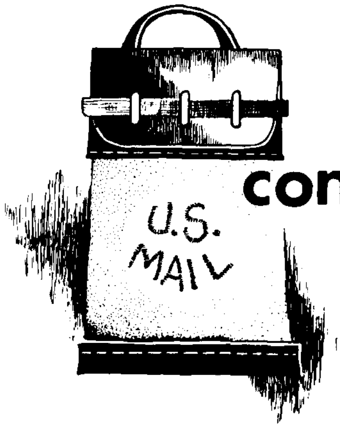
More Details? CHECK–OFF Page 94



comments

solid-state tube replacements

Dear HR:

I was highly interested to see the letter from Mr. Walter Loomis printed in the October, 1973, issue of *ham radio.* The company that I am employed with is in the business of manufacturing semiconductors, and one of our products is very similar to the Fetron that Teledyne produces.

Being active in vhf myself, and having numbers of tube-type communications receivers that use 6AK5 tubes, I gave our devices a try. The following observations may be useful to some of your readers.

1. Our first tube substitute was made in an epoxy block. The performance of this unit was slightly less than could be obtained by going to a ceramic substrate, as Teledyne had done. The main problem with the epoxy package was feedback capacitance, measured at about 0.05 pF.

2. The input to the device is the gate of a jfet. Gate currents of good jfets are in the picoamp range, while grid currents in vacuum tubes are in the hundreds of nanoamps, even when operated at several volts of negative bias. This may cause problems with biasing, since the vacuum-tube oriented engineer may have been counting on this electron interception current to provide some operating bias for the tube. The Fetron will not show this bias.

3. Input capacitance of the Teledyne device is specified at 8 pF maximum, which is twice what the tube will have. This will lead to all sorts of impedance-matching troubles, even if the tuned circuit can be resonated, at operating frequencies above approximately 50-MHz.

4. The gate of the input jfet looks like a diode connected from grid to cathode. This diode cannot be forward biased to much more than 0.8 volts. Consequently, applications depending on short pulses of high current (class-C frequency multipliers) cannot be expected to work well, as the tube replacement cannot be enhanced as the original vacuum tube could.

5. Many communications receivers use a small *positive* bias voltage applied to the bypassed end of the grid bias resistor for the rf amplifier. This voltage becomes a few volts negative on strong signals from agc circuitry in the i-f of the receiver. However, when the receiver is idling with no signal input, the positive bias voltage will flow through the resistor and forward bias the gate diode in the input jfet. This bias causes the jfet to show a marked decrease in g_m, usually to about 1/10 of the original value. Thus, a sort of "inverse agc" action results. Additionally, the input impedance decreases markedly under these conditions, further aggravating the low gain problem. The fix here is to short out the agc line.

I have tried these solid-state tube replacements in every receiver that I have using a 6AK5. The results were most encouraging. Our epoxy model tube replacement worked well at 6 meters, showing a 2-dB noise figure improvement over the tube. Gain was slightly higher (about 1 dB) than the tube. Equally good results were obtained in 70-MHz i-f amplifier service. At 147 MHz, results were not encouraging. The input circuit could not be made to tune to the operating frequency, and overall sensitivity was extremely poor. Switching to our ceramic substrate type of tube replacement did not provide too much improvement at first. However, an investigation of the problem revealed that good results could be obtained.

The noise figure of these devices is quite good, as the input transistor is a somewhat improved 2N4416. The feedback capacitance is extremely low *when the metal can is grounded.* Measurements show it to be approximately 0.01 pF under operating bias. The Y_{os} at the plate terminal is extremely low, about 0.05 micromho. This is much better than the tube. In short, this device can run circles around the tube it replaces, provided the circuit is adjusted to optimize the design of the stage for the replacement device.

After changing around the input and output matching networks of one receiver on 147 MHz, I obtained a sensitivity of 0.2 microvolts for 20-dB quieting (with a 20-kHz i-f bandwidth). Now I will never have to replace my rf stage again and I can expect long and trouble free maintenance of the sensitivity. But I had to work on the receiver to take advantage of the performance of the device.

I have also used the device in low-frequency applications such as limiters at 455 kHz. The tubes replaced here were 6BH6s and 6CB6s. In most of the places I tried them, the solid-state replacements worked as well as the original tubes. In some sockets, instability and oscillation was the result. The short between pins 2 and 7 is probably responsible for this.

An important point about fets that will probably limit the usefulness of solid-state tube replacements in a-m equipment (including ssb) is that the fet is a square law device, while the vacuum tube is only a poor approximation to square law, at best. Why is this? Because an a-m system almost always requires agc, and the agc action of a remote-cutoff pentode is very slow. Useful amplification can be obtained at −10 volts of grid bias with the remote-cutoff pentode, while the square-law jfet will long since have pinched off and become merely a low value capacitor connecting the input and output. On the other hand, since the jfet is such a good square-law device, it will provide excellent intermodulation distortion performance, which is a key requirement in today's equipment.

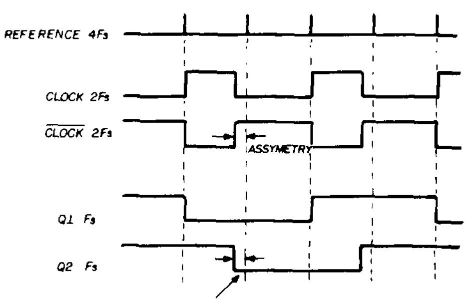
To summarize, I found the solid-state tube replacement to be a great device, but it is also one whose requirements must be understood to make it useful.

Bob Hale, WB6APU
Solitron Devices
San Diego, California

phasing receiver

Dear HR:

The Phase II communications receiver described by WAØJYK in the August, 1973, issue is a good example of the practical realization of well-known concepts that can only be achieved after a certain level of technical refinement is reached in the state of the art. Simple concepts often seem the most difficult to implement. I would like to add a few comments that I hope will lead to a better understanding of the digital phase-shifting technique.



The broad-band digital phase shifter used by WAØJYK, while only requiring a two-times frequency division instead of the more common four-times division, depends heavily on symmetry of clock transition timing to maintain phase quadrature. That is, for proper operation, the clock *transitions* must be periodic at 4-times F_{signal} since the basic two-stage shift or switch tail counter *must* have a clock component at four-times F_{signal} to deliver quadrature phase output at F_{signal}. The paraphase clock driver described in the article is a form of frequency doubling and causes one flip-flop to change state every clock transition. The timing diagram above is offered to demonstrate this point. Clock asymetry is exaggerated for illustration.

Asymetry of the clock transisions at twice the clock frequency may have several causes. Among them are junction saturation effects (minimal in ECL), non-linear junction capacitance, gain drifts, unequal path propagation delays and, in the circuit used in the article, comparator threshold and hysteresis. All of these effects are proportional to frequency. Differences in the Q1, Q2 flip-flop clock to output propagation times also add to departures from orthogonality but for the 10131 ECL dual D-type flip-flop used, these effects should be smaller than the clock asymetry problems described above. Phase shifts due to differential flip-flop output loading and mixer differences should also be considered.

In summary, if best use is to be made of the broad-band digital quadrature phase generator technique and optimum performance obtained from the high speed devices available, both flip-flops in the chain must see identical clock transitions. This requires, of course, a fundamental clock frequency of four times the desired signal frequency.

Julius M. J. Madey, W6FAW
Fairfax, California

Mr. Madey has pretty well assessed the problems involved. I think in a word the present design might be summed up as economical. The frequency coverage was already limited by the immediate demands of Air Force Mars and no big effort was made to extend the range above 10 MHz. ECL logic is non-saturating and therefore does not radiate as much noise and doesn't cause as much noise in the power supply as does saturating-type logic. The MC1035P is very economical and has proven to be a real workhorse. The comparator threshold and hysteresis that Mr. Madey mentioned no doubt exist in the MC1035 and enter into the poor performance obtained above 10 MHz. Clock symmetry of the MC1035P at 7 MHz looks very good on a Tektronix 545 but even that scope is not sufficient for a fair evaluation.

The MC1035P presents no particular problem in waveform symmetry below 10 MHz providing that either good crystals are used or a vfo is transformer coupled into the crystal socket with at least 1-volt p-p. With transformer coupling the vfo seems to preserve the symmetry of the vfo signal.

The big problem arises when the 10-MHz frequency limit is exceeded or the clock source is a synthesizer with a dissymmetrical output. Unfortunately, most inexpensive synthesizers and, most of all, those with TTL ICs have dissymmetrical outputs. One local amateur, well versed in digital logic, presented a convincing argument for including the digital phase shifter right in the synthesizer.

Perhaps someone will develop this synthesizer with the quadrature output and the Phase II receiver can be pushed to 20, 15 and even, possibly, 10 meters. It seems the digital phase shifter is sort of the thorn on the road to success. There is no way around it but perhaps we can help each other over it.

G. Kent Shubert, WAØJYK

motorola test set

Dear HR:

I want to congratulate you on Dave Maxwell's fine article, "Test Set for Motorola Radios." I have built it and we are now using it to align USAF MARS Motorola units on a MARS repeater frequency in the Las Vegas area.

Two items of interest. The Amphenol microphone jack may be converted to a Motorola microphone jack by sawing a wider guide slot in the Amphenol unit. The Simpson 1212 50-μA meter has an internal resistance of 5300 ohms. Any meter in the 5000-ohm range may be used.

Oscar Heinlein, W7BIF
Boulder City, Nevada

all-channel frequency synthesizer

Dear HR:

The article, "Inexpensive All-Channel Frequency Synthesizer for Two-Meter FM," by Jim Fahnestock, WØOA, which appears in the August, 1973, issue of *ham radio* could cause people some serious problems.

Several years ago I wrote a technical report* on spurious outputs from mixers, and I wondered about such outputs when I read WØOA's article. I used the programs described in that report to analyze the frequency sets he used, and then wrote a rather simple program in BASIC to do a somewhat more thorough analysis. The results of the latter calculation are enclosed for your information (see **table 1** below).

table 1. Possibly troublesome spurious mixer products of WØOA's frequency synthesizer using the 4 x 4 matrix. These products are based on an output at 146.28 MHz, using a fundamental frequency of 6095 kHz and 24-times multiplication (F1 = 8125 kHz, F2 = 2030 kHz).

M	N	Frequency	M+N
1	1	6095 kHz	2
1	7	6085 kHz	8
2	5	6100 kHz	7
2	11	6080 kHz	13
3	9	6105 kHz	12
3	15	6075 kHz	18
4	13	6110 kHz	17

I used an output frequency of 146.28 MHz as a basis since that is our local repeater input frequency. The fundamental frequency for a multiplication of 24 is 6095 kHz. Note that for M + N = 7 there is a spurious output only 5 kHz away. There is another spurious output 10 kHz away from the desired signal for M + N = 8. These are both apt to result in strong spurious outputs from the fm transmitter since there is no way of removing them by filtering.

In the article Mr. Fahnestock suggested that "mixing problems arise" when using a 5 x 5 or a 6 x 6 matrix of frequencies. I did all the necessary calculations for a 6 x 6 matrix and analyzed it with the BASIC program. Interestingly enough, the mixing problems are down by two orders of M + N, and the closest spurious frequency goes out to 7.5 kHz away. This is an extremely doubtful practical solution, but it would be many dB better than the 4 x 4 matrix.

It is possible that these schemes would work if digital mixers were used, but I am by no means certain since I have no experience with them. It seems unlikely that the author considered them with his emphasis on various filters.

Bill Wageman, K5MAT
Los Alamos, New Mexico

*William E. Wageman, "Analysis of the Spurious Frequency Response of Mixers," Los Alamos Scientific Laboratory document LA-4296-MS, December, 1969.

vhf fm in england

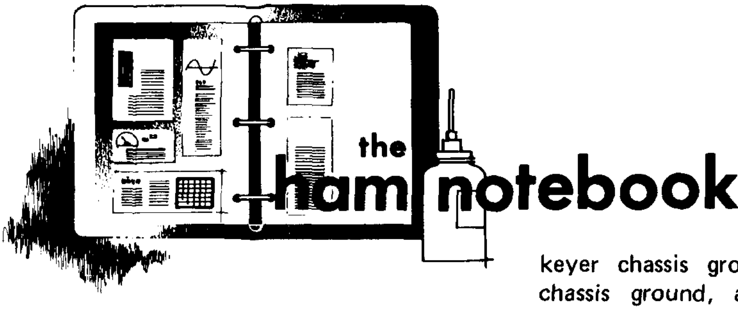
Dear HR:

Here in England vhf fm has really gotten going. We have several fm groups similar to the local one (Southern FM Group). One fm repeater is in operation, licensed as GB3PI, and several more are under construction and waiting license approval.

The Southern Group two-meter repeater with which I am involved will use a modified Storno 613 solid-state base station. The control and ID equipment will be built by club members. Output power will be 25 watts, fed into an Antec antenna (7-dB gain). Input frequency will be 145.125 MHz; output will be at 145.725 MHz. Callsign will be GB3SN.

I have found the repeater articles in *ham radio* to be very useful in designing the control equipment for GB3SN and I would like to thank the American authors. I would be interested to hear from Stateside repeater groups with a view to discussing mutual problems.

Rodney V. Smallwood, G8DGR
11 Wilmot Walk, Wash Common
Newbury, Berks, England



npn transistor switching for electronic keyers

Many amateurs own solid-state keyers which use a pnp transistor switch to key a transmitter's grid-block keying circuit. Such keyers can easily be adapted for use with solid-state transmitters which require a switching positive voltage above ground to negative ground. There are several ways to do this. First, a relay can be inserted between the keyer and the transmitter. Although this method is generally used, the noise of the relay is often found objectionable. Then, too, a relay takes quite a bit of power; this is an important consideration when a keyer is battery-operated.

Second, the pnp transistor in the keyer can be used to do the switching by reversing the leads to the transmitter key jack, but this necessitates separation of keyer chassis ground from transmitter chassis ground, a practice not to be recommended. Third, an npn transistor switch can be inserted between the keyer and the transmitter. The keyer would be plugged into the npn switch built into a small Minibox and the latter plugged into the key jack of the transmitter. The keyer, the npn switch and the transmitter would all have a common chassis ground but different electrical grounds. This method is highly recommended. The little device is inexpensive to construct and can be put together in very little time. The schematic and parts list are given in **fig. 1**.

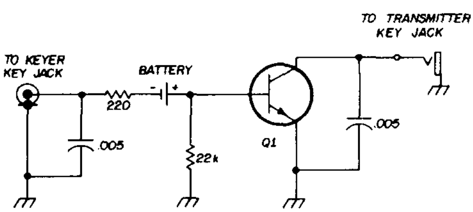


fig. 1. Circuit for keying low positive voltages to negative ground using keyer with pnp switching transistor. Battery is a single penlight call; capacitors are disc ceramics; resistors are ½ watt; Q1 is a Motorola 2N3904 or similar npn transistor.

Those amateurs who have keyers with npn switches but would like to use solid-state switching instead of a relay to control grid-block keying circuits in their transmitters can also use this conversion method. Only two changes in the device described here are necessary to adapt it for that purpose. The polarity of the battery must be reversed and the npn switching transistor in the adaptor must be replaced by a Fairchild 2N4888 or similar pnp transistor.

C. Edward Galbreath, W3QBO

cleaning files

If you do much work with aluminum you will find that your files and cutting tools rapidly plug up with aluminum filings. These filings are usually quite difficult to remove, especially from the finer toothed files. Don't throw these plugged files out, you can make them look like new again with this simple method. Just soak the files in a solution of warm water and common household lye. The lye will dissolve the aluminum particles and the hydrogen bubbles released by this reaction will help to dislodge any metal particles that may also be plugging the file. When working with lye it is wise to wear rubber gloves and avoid spilling or dripping the solution on any surface that you don't want harmed.

Pete Walton

build 5/8-wavelength 144-MHz antennas from CB mobile whips

The popularity of 3-dB gain mobile antennas for two-meter fm communications has very little to do with their steep cost. The following is an easy method for converting inexpensive CB whips to 5/8-wave antennas that are identical to commercial gain antennas. In many cases 27-MHz CB whips may be obtained at half the cost or free, depending on your sources of supply.

Before you run over to your local CB supplier, or patronize the want ads, be sure that you can identify the proper type of whip for this conversion. These whips appear to be identical to commercial vhf whips (148-158 MHz), the difference being in the hidden resonator within the epoxy sleeve antenna base. CB whips should dip to around 27 MHz on a grid-dip meter.

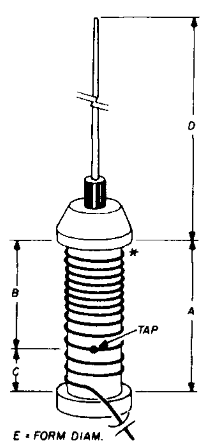


fig. 2. Modifying a CB mobile whip to 5/8-wavelength whip for two-meter fm. No. 12 wire is used for the winding. Connection at top of winding is on other side (point marked with asterisk).

Some epoxy sleeves may be pried loose from the antenna with the aid of a sharp screwdriver. Others may have to be unscrewed by gripping the chrome fittings at both ends with two pairs of pliers. Removing the core should be easily accomplished by friendly persuasion from a hammer and soft wood rod. Getting past the ring of sealer may require some additional effort and patience (five minutes).

Winding the new coil may require five minutes or so of your time. The diameter of the coil forms may vary slightly among manufacturers, but present no problem. Just make sure that your tap remains at three turns above the ground end. The only compensation which may be necessary is extending or shortening the radiating element by a few inches. Two

table 1. Typical dimensions of CB mobile whips (see fig. 2).

	A	B	C	D	E
whip 1	2"	6½ turns (1½")	3 turns (½")	44"	5/8"
whip 2	1½"	7½ turns (1")	3 turns (½")	44"	1/2"

specific examples of typical whips are shown in **table 1** to eliminate any guesswork.

Most commercial whips are universal as far as roof mounts, hardware and threading are concerned. Manufacturers make available individual components which are contained within the antennas. By simply screwing together some of this hardware, your modified equipment can be made to accommodate your existing equipment.

Also, you don't have to limit yourself to available commercial whips. The coil winding data is applicable to plexiglass rods, and readily available coaxial connectors for the homebrewer. You'll find, as I did, that none of the specifications are critical.

There is no noticeable difference between my commercial whip and the modified antenna. Both exhibit significant advantage over quarter wave whips. The swr is 1.05:1.

Robert Harris, WB4WSU

new products

digital in-line rf wattmeter

The new Bird 4371 *Thruline*® Directional High-Power Wattmeter is the first digital insertion instrument for measuring forward or reflected CW power in coaxial transmission lines. It accurately measures power flow under any load condition from 25 to 520 MHz and from 1—1000 watts in six ranges. Insertion vswr in 50-ohm systems is a low 1.1 and accuracy is ±5% of full scale. Model 4371 is also the first high-power directional wattmeter which the user can calibrate in the field to known rf power standards, eliminating weeks of transit for periodic certifications.

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Available from Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139. For more information, use *check-off* on page 94.

RTTY handbook

The new *Teleprinter Handbook* published by the RSGB and available from HR Books is one of the most complete books ever published on the subject. Under preparation by G3LLZ and G3NTT for more than three years, this welcome handbook fills a long-standing need of RTTY operators. The *Teleprinter Handbook* covers all aspects of RTTY communications, including complete mechanical, operating and lubrication data on machines made by Teletype Corporation, Creed and Siemens as well as the surplus TT4 and TG7-A. Information is also included on teleprinters, perforators, reperforators, signal generators and distortion measuring sets.

The chapter on terminal units includes complete construction information for several popular RTTY demodulators including the W6FFC's Mainline ST-5 and ST-6 and TT/L-2. Several commercial units are also described including the popular AN/URA-8B. This chapter also has some good practical information on diversity reception, regenerative repeaters, phase-locked loops and frequency-shift keying. The auxiliary equipment chapter covers such diverse subjects as polar relays, mercury-wetted relays, and electronic and mechanical speed controls. The section on polar relays provides complete operation, maintenance and applications information for a large variety

More Details? CHECK—OFF Page 94

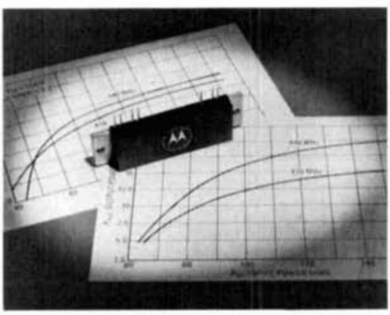
of currently-available polar relays including the Western Electric 215A and 255A.

The chapter on keying discusses FSK and AFSK techniques and includes circuits for several keying units, both commercial and amateur built. Various filter circuits are discussed in detail, with design and construciton information for both passive and active types. The test equipment section provides information on various RTTY measurements, and the control systems chapter tells you how to set up your own RTTY station. The *Teleprinter Handbook* also covers RTTY operating procedures and contests.

One of the most useful sections in the *Teleprinter Handbook* is the glossary of commercial equipment, which provides a short rundown on practically every unit of available RTTY equipment. The appendix provides much practical data including complete design information for simple LC filters based on 88-mH toroids, gear speed identification and teleprinter compatibility.

This book is probably the best teleprinter handbook ever published, and combines a lot of hard-to-find RTTY information all in one place. Highly recommended for RTTY enthusiasts. Hard cover, 13 chapters plus appendix and index, 324 pages, $14.95 from HR Books, Greenville, New Hampshire 03048.

432-MHz hybrid rf amplifiers



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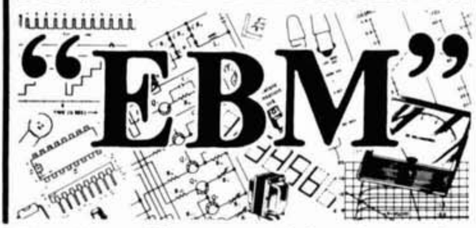

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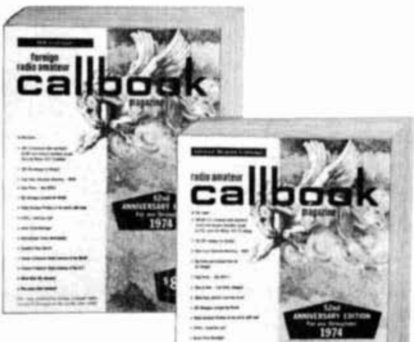

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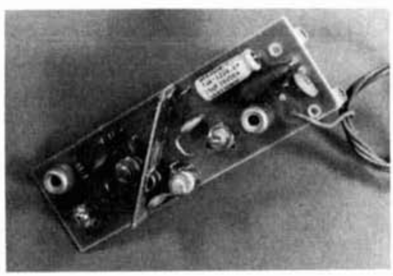
amplifier modules introduced by Motorola offer more than 18 dB power gain at 432 MHz. Designated the MHW709, (7.5-watt) and the MHW710 (13.0-watt) uhf power modules, these are complete amplifier units capable of covering the 400 to 470 MHz frequency range.

Both units operate from a 12.5-volt dc supply which is common in most mobile applications. The MHW 709 delivers 7.5-watts output with a driving power of approximately 100-mW for a power gain of 18.8 dB. A full 13.0-watts can be produced by the MHW710 with only 150-mW of driving power; this represents a 19.4-dB power gain. The frequency range is covered in two bands with the units designated MHW709-1 and MHW710-1 for 400 to 440 MHz, and MHW709-2 and MHW710-2 to cover the 440 to 470 MHz range.

Harmonic suppression is at least -40 dB down across the frequency range with all spurious outputs more than 70-dB below the desired signal. Input impedance is 50 ohms for both modules, and operation with a 20:1 load mismatch produces no damage to the unit.

For more information contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, P.O. Box 20924, Phoenix, Arizona 85036, or use *check-off* on page 94.

450-MHz preamp



Topeka FM Engineering has introduced a new 450-MHz preamp that features a dual-gate mosfet in the front end. The mosfet provides superior cross-modulation performance and reduced spurious responses. The new preamp, the model 450M, is built on low-loss epoxy

More Details? CHECK—OFF Page 94

circuit board, and has a voltage gain of approximately 15 dB with a typical noise figure of 4.5 dB. The circuit board is silver plated for maximum efficiency, and is shielded on both sides for maximum isolation of the input and output circuits.

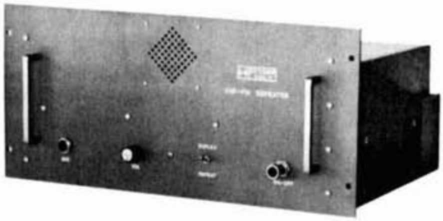
The 450M preamplifier is designed to operate from 10 to 15 volts dc, but may be used with higher supply voltages when the accessory HF450PK adapter is used. Similar preamplifiers are available for 406 to 470 MHz. The preamplifier is priced at $29.95, complete with instructions and mounting hardware. The HF450PK power supply adapter is $1.25. For more information, write to Topeka FM Engineering, 1313 East 18th Terrace, Topeka, Kansas 66607, or use *check-off* on page 94.

hard-to-find electronic components

Radio Shack has added more than 2,000 hard-to-find electronics items to its parts and test equipment line with a special "Qwick-Fill" electronic parts catalog. The 52-page Radio Shack Electronic Parts Catalog is available on request from any Radio Shack store. It includes a variety of special-purpose tubes, transistors, readouts, PC and IC equipment, relays, resistors, capacitors, potentiometers, transformers, sophisticated test instruments, connectors, power supplies and other items previously unavailable from Radio Shack or other electronics stores in most localities.

Through a special arrangement with Allied Electronics, Radio Shack's sister company, orders may be placed at any of more than 1800 participating Radio Shack stores. Allied Electronics is known as one of the nation's leading electronic parts suppliers for business, industry and scientific users. There is no minimum order requirement and only a $1.00 shipping and handling charge. Orders are delivered to the Radio Shack store for pickup by the customer. For more information use *check-off* on page 94.

DYCOMM



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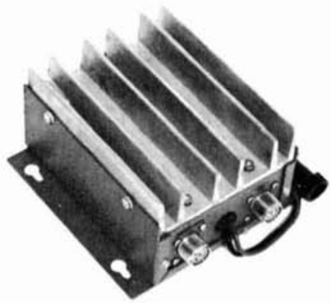

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For Technical Details
See Ad in March Ham Radio for Super E.
See Ad in December Ham Radio for Super D.

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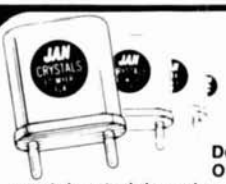

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The HAL ID-1A is available factory assembled for $115 or as an assembled board/kit, including all parts external to the board except the cabinet for $75. Look into it. Better yet, order yours today.

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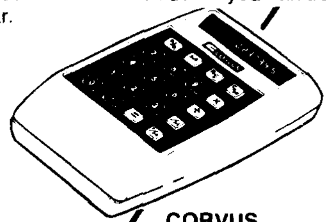

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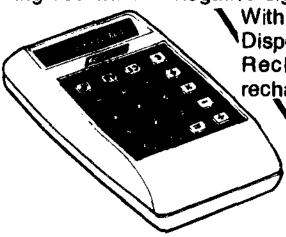

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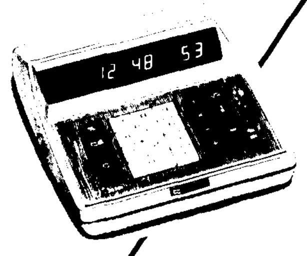
CORVUS 305 DESK-TOP CALCULATOR WITH CLOCK & CALENDAR



full 7 function (+, -, x, ÷, %, square root and reciprocal) with 10 digit readout, selectable 2-4 or floating decimal, automatic, constant and many features. But when the 305 isn't being used as a calculator it's a digital clock and calendar that keeps perfect time as long as it's plugged in.

Model 305 .... **$99.95**
Model 300 w/o clock **84.95**

ADD $2.75 TO EACH ORDER FOR SHIPPING.

Send check, money order, Bank Americard No., Mastercharge No. or American Express No. to: (Residents of Texas, Illinois, New Jersey & California include 5% sales tax)

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ELECTRONICS · COMPANY

P.O. Box No. 1050C
Garland, Texas 75040

FOR RUSH ORDERS CALL TOLL-FREE 800-527-4642 (In Texas call 214-348-8800)

More Details? CHECK—OFF Page 94

IRON POWDER R.F. TOROID CORES

CORE SIZE	-41 Mix Green 'HR' 20 kc - 100 kc μ=75	-3 Mix Gray 'HP' 50 kc – 1 mc μ=30	-2 Mix Red 'E' 500 kc – 30 mc μ=10	-6 Mix Yellow 'SF' 10 mc - 90 mc μ=8	-10 Mix Black 'W' 30 mc -150 mc μ=7	-12 Mix Grn-Wh 'IRN-8' 60 mc – 200 mc μ=5	Outer Diameter (inches)
T-200	$2.50	$2.75	$3.00	$3.50			2.000
T-130	1.50	1.75	2.00	2.50			1.300
T-106	.95	1.00	1.00	1.50			1.060
T- 94	.70	.75	.75	.95			.942
T- 80	.55	.60	.60	.80	.90		.795
T- 68	.45	.50	.50	.65	.75		.690
T- 50	.40	.45	.45	.50	.60	.65	.500
T- 37	.30	.40	.40	.45	.45	.55	.370
T- 25	.25	.30	.30	.35	.40	.45	.255
T- 12	.25	.25	.25	.25	.25	.35	.125

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Use Amidon Ferrite Beads for Parasitic Suppression, Shielding, Noise Suppression, Spike and Transient Clipping, RFI Suppression, Antenna Loading and for Special Inductors. The Regular 3 mm bead accepts up to #18 wire. The Husky 7.5 mm bead accepts #12 AWG. Each Husky bead exhibits an inductance of 1.25 Microhenry. Permeability Factor: 900

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Husky Beads .... Package of 12, $3.00

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DRILLED PCB ONLY $ 6.50
RTTY AFSK Gen. All Shifts & CW I.C. 9 VDC @ 2ma $7.25
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LIMITER PREAMP For High Z Mike 9 VDC @ 1ma $5.30
PRODUCT DETECTOR For Your Receiver 9 VDC @ 1ma $3.95
"S" METER KIT Less 1ma Meter 6.3VAC $5.25
SWR METER, Stripline, Less 200μa Meter $3.25
WWV CONVERTER 3.5-4.0 MHz Output 9 VDC @ 5ma $5.75
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7-11 MHz Output, Less 43 MHz Xtal
CW KEYING MONITOR, RF Keyed, Less Spkr. 9 VDC @ 9ma $5.20
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Wired & Tested Less 2 ea 6CW4 Nuvistors
DRILLS, #54, 56, 58 or 60 (each) $.50
Finest Quality for PCB's, Made in USA Three For $1.25

EXCEPT AS NOTED ABOVE, ALL KITS ARE NEW, 100% SOLID STATE, AND COME COMPLETE WITH AN UNDRILLED G-10 PCB (PRINTED CIRCUIT BOARD) AND ALL PCB MOUNTED COMPONENTS. KITS ARE LESS POWER SUPPLIES, CHASSIS, AND ENCLOSURE HARDWARE. SEND SELF-ADDRESSED, STAMPED ENVELOPE FOR COMPLETE DATA SHEET AND SCHEMATIC.

SATISFACTION GUARANTEED. RETURN IN 30 DAYS FOR REFUND. ALL KITS POSTPAID. INCLUDE 50¢ HANDLING CHARGE. WASHINGTON RESIDENTS ADD 5.3% SALES TAX.

Martex Corporation

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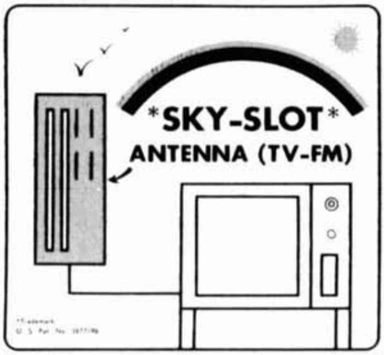

TV-FM-220 MHz FM ANTENNA NEW

ROLLABLE (with perfect color band-width). Just unroll and stick on wall. Uses no masts, no rabbit ears, no dangerous plugging to AC lines, and NO SIGNAL SPLITTERS since the VHF-UHF-FM terminals (VHF usable for 220 MHz FM) available at the same time. (New 1971 slotted design with U.S. Patent S/N 3577196). Made of decorative foil/plastic yet rugged for use in yacht or RV. Optional: Plain burlap or printed etching on burlap of either desert, galaxy or leaved branch. Antenna size: 18" x 48" unrolled, works behind picture, against metal! Price is only $15 for model with following gain: (VHF - 9 dB), (UHF - 12 dB), (FM - 6 dB); gains above dipole.

Order postpaid AIRMAIL, insured, from

ANTENNA DESIGN CO.

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More Details? CHECK—OFF Page 94

How's Your FM I.Q.?

Try Your Hand At Answering These Few Simple Questions About the World of FM.

WHO FIRST OFFERED A PACKAGE DEAL ON THE CLEGG FM27 WITH A $100 SAVINGS?

See Feb. 1973 QST Magazine

	List
Clegg FM 27	$479.50
Dycomm 35-0	175.00
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	684.45

Your Price $584.45 (fob store)
You Save $100.00!!

WHO HAS AVAILABLE FOR IMMEDIATE DELIVERY A FULLY SYNTHESIZED 2 METER TRANSCEIVER OF "MOTOROLA QUALITY" (AT AN AMATEUR PRICE) WITH FULL 144-147.99 MHz COVERAGE IN 10 kHz STEPS?

WHO HAS OVER 38 MAJOR BRANDS FOR THE FMer INCLUDING ANTENNA SPECIALISTS, CLEGG, CUSH CRAFT, GENAVE, ICOM, MOTOROLA, SBE, DYCOMM, CDR, AND RP.

WHO SELLS MORE AMATEUR ONLY FM EQUIPMENT THAN ANY OTHER DEALER/DISTRIBUTOR IN THE WORLD.

WHO CAN OFFER OVER 35 YEARS OF COMBINED AMATEUR FM EXPERIENCE WHEN IT COMES TO FILLING YOUR FM NEEDS?

WHO NORMALLY SHIPS YOUR ORDER THE SAME DAY AS RECEIVED AND CAN CHARGE YOUR BANKAMERICARD OR MASTER CHARGE?

To See How Well You Did Check Page 91 For Answers.

Advertisement

SB36 SSB CW TRANSCEIVER

an experience in relaxed operating enjoyment

- Full coverage, SSB/CW, 80-40-20-15-10 meter bands.
- 500W p.e.p. power input.
- 6-digit frequency counter reads to 100 Hertz.
- Selectable USB/LSB/CW.
- Receiver Incremental Tuning.
- Double conversion receiver with fractional uV sensitivity.
- Panel multi-meter: "S" units, plate MA, RF out, ALC.
- 51 transistors (6-MOSFETS), 32 diodes, 6 vacuum tubes.
- 115/230VAC supply, speaker, in matching assembly cabinet.

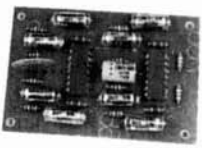
WRITE FOR COMPLETE BROCHURE.

SB-36—personalized station and—with half-a-thousand watts p.e.p. input, a table-top powerhouse. But much more! Picture the precision, 6-digit counter that reads out your frequency as fast as you can spin the free-turning, counter-weighted tuning knob. And see how RIT (Receiver Incremental Tuning) obsoletes older, zero-beat-only transceiver operation—prevents leap-frogging. Push the RIT button—zero-in received sigs ±7kHz—without effecting transmit frequency! Also, change sidebands without beat note shift or retuning. It takes **two** 9MHz crystal lattice filters to do this job. SB-36 has 'em.

SBE

LINEAR SYSTEMS, INC.
220 Airport Blvd., Watsonville, CA 95076

CW FILTER



New Model CWF-2BX—$19.95. Ready to use. Please include $1.00 postage

Model CWF-2—$12.95, Kit. $14.95 Wired, tested, guaranteed. Please include 55c postage.

- **Get Razor Sharp selectivity from any receiver or transceiver.**
- **Extremely high skirt rejection**
- **Drastically reduces all background noise.**
- **No audible ringing.**
- **No impedance matching.**
- **Ultra modern active filter design uses IC's for super high performance.**

We have what we think is the finest CW filter available anywhere. The 80 Hz selectivity with its steep sided skirts will allow you to pick out one signal and eliminate all other QRM and QRN. Simply plug it into the phone jack or connect it to the speaker terminals of any receiver or transceiver and use headphones, small speaker, or speaker amplifier. Better yet, connect it between any audio stages to take advantage of the built in receiver audio amplifier.

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

SPECIFICATIONS

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

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

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

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NOW-------**An extremely rugged, weather-proof BALUN!**
- **Full 2KW, 3-30 MHZ., 1:1 or 1:4 ratios.**
- **Special Teflon insulation, May be used with tuned lines and tuners.**
- **With dipole insulator and hang-up hook.**

ONLY $9.95ppd. (state ratio)
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COUNTS & DISPLAYS YOUR TURNS

- *99.99 Turns*
- *One Hole Panel Mount*
- *Handy Logging Area*
- *Spinner Handle Available*

Case: 2x4"; shaft ¼"x3"
Model TC2: Skirt 2⅛"; Knob 1⅝"
Model TC3: Skirt 3"; Knob 2⅜"

PRICES	POST PAID
TC 2 —	$7.00
TC 3 —	$7.50
Spinner (-S) —	$1.00

Add $0.75 for Air or UPS

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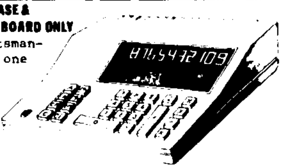
More Details? CHECK-OFF Page 94

TABLE TOP CALCULATOR CASE & KEYBOARD ONLY



Attractively designed! Excellent craftsmanship. Case and keyboard (designed as one unit). Cabinet is made of high-impact plastic beige color with black bezel and amber window. Keyboard consists of a 3-position slide switch and 25 keys, 5 of which are used for memory function. 20 keys gray, 5 keys orange. All keys mounted on one printed circuit board. This modular unit is well suited for our calculator chips. Ideal for the CT5005. Case and Keyboard Complete: ONLY $15.95

RCA NUMITRON

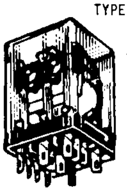
EACH............$ 5.00

SPECIAL: 5 FOR $20.00

DR2010

Potter & Brumfield

TYPE KHP RELAY 4 PDT 3A CONTACTS



24 VDC (650 coil)....$1.50

120 VAC (10.5 MA coil).... 1.75

CMOS

CD4001	$ .75	CD4011	$ .75
CD4002	.75	74C20	.75
CD4010	1.00	74C160	3.25

CT5005 CALCULATOR

This calculator chip has a full four-function memory, which is controlled by four keys, +M (adds entry into memory), -M (subtracts entry from memory), CM (clear memory, without clearing rest of registers), RM (read memory or use as entry).

- 12-Digit display and calculate
- Fixed decimal at 0, 1, 2, 3, 4, or 5
- Leading zero suppression
- 7-Segment multiplexed output
- True credit sign display
- Single 28-pin chip

CHIP AND DATA.........ONLY $14.95
DATA ONLY (Refundable)....... 1.00

5001 CALCULATOR

40-Pin calculator chip will add, subtract, multiply, and divide. 12-Digit display and calculate. Chain calculations. True credit balance sign output. Automatic over-flow indication. Fixed decimal point at 1, 2, 3, or 4. Leading zero suppression. Complete data supplied with chip.

CHIP AND DATA........ONLY $9.95
DATA ONLY (Refundable).... 1.00

All ICs are new and fully-tested; leads are plated with gold or solder. Orders for $5 or more will be shipped prepaid. Add 35¢ for handling and postage for smaller orders; residents of California add sales tax. IC orders are shipped within 2 workdays--kits are shipped within 10 days of receipt of order. $10.00 minimum on C.O.D.s (phone in).

MAIL ORDERS:
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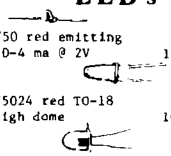
MONEY BACK GUARANTEE ON ALL GOODS!

SEND FOR FREE FLYER LISTING 100's OF MONEY-SAVING BARGAINS!

BABYLON ELECTRONICS

LED's



MV50 red emitting 10-4 ma @ 2V	$ .20 / 10 FOR 1.25
MV5024 red TO-18 high dome	$ .35 / 10 FOR 2.95
MV10B visible red 5-7 ma @ 2V	$ .30 / 10 FOR 2.50

Memory IC 74S206

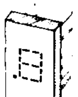
1x256 Bipolar TTL RAM (45 ns). One bit wide by 256 deep random access memory. Three-chip enable lines. Output is complement. Complete data supplied.

$3.95 ea. 10 for $32.00

3-Amp Power Silicon Rectifiers Marked Epoxy Axial Package

PRV	PRICE	PRV	PRICE
100.....	$ .10	800.....	$ .30
200.....	.15	1000....	.40
400.....	.18	1200....	.50
600.....	.23	1500....	.65

MAN 1



7-Segment, 0-9 plus letters. Snaps in 14-pin DIP socket or Molex. Operates with IC voltage requirements. Long operating life. ONLY $3.25

7400 Series DIP

7400	$ .25	74L51	$ .30
74H00	.35	74H51	.35
7401	.20	7453	.20
74H01	.35	7454	.25
7402	.35	74L54	.35
7403	.30	74L55	.35
7404	.28	7460	.20
74H04	.35	74L71	.30
7405	.28	7472	.40
7406	.70	74L72	.50
7408	.35	7473	.60
74H08	.35	74L73	.75
7410	.25	7474	.65
74H11	.35	74H74	.80
7413	1.25	7475	1.40
7417	.40	7476	.60
7420	.25	74L78	.80
74L20	.35	7480	.65
74H20	.35	7483	1.00
74H22	.35	7489	4.00
7430	.25	7490	1.20
74H30	.35	7492	.90
74L30	.40	7493	1.15
7440	.25	7495	1.15
74H40	.35	74L95	2.00
7441	1.25	74107	.70
7442	1.20	74121	1.25
7446	1.75	74154	2.30
7447	1.50	74163	2.00
7448	1.50	74192	2.50
7450	.25	74193	1.50
74H50	.35	74195	1.00
7451	.25		

RECTIFIERS

VARO FULL-WAVE BRIDGES

V5447	2A	400V	$ .90
V5647	2A	600V	1.10
MR810 Rect.	50V	1A	.10

Special 811: Hex Inverter

TTL DIP Hex Inverter; pin interchangeable with SN 7404. Parts are brand new and are branded Signetics and marked "811."

Data Sheet Supplied

EACH.......$ .30
10 FOR..... 2.50
100 FOR.... 23.00
1000 FOR... 220.00

MAN 3

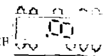
0-9 plus letters. Right-hand decimal point. Flat-pack type case. Long operating life. IC voltage requirements. Ideal for pocket calculators!

EACH $1.25
10 OR MORE 1.00

MAN4

Seven-segment, 0-9 plus letters. Right-hand decimal point. Snaps in 14-pin DIP socket or Molex. IC voltage requirements. Ideal for desk or pocket calculators!

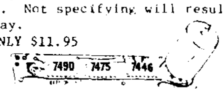
EACH.......$2.75
TEN OR MORE 2.50 EACH



CD-2 Counter Kit

This kit provides a highly sophisticated display section module for clocks, counters, or other numerical display needs. The unit is .8" wide and 4 3/8" long. A single 5-volt power source powers both the ICs and the display tube. It can attain typical count rates of up to 30 MHz and also has a lamp test, causing all 7 segments to light. Kit includes a 2-sided (with plated thru holes) fiberglass printed circuit board, a 7490, a 7475, a 7447, a DR 2010 RCA Numitron display tube, complete instructions, and enough Molex pins for the ICs . . NOTE: boards can be supplied in a single panel of up to 10 digits (with all interconnects); therefore, when ordering, please specify whether you want them in single panels or in one multiple digit board. Not specifying will result in shipping delay.

COMPLETE KIT, ONLY $11.95
FULLY-ASSEMBLED UNIT $15.00



Boards supplied separately @ $2.50 per digit.

LINEARS

NE540	70-Watt power driver amp........	$2.00
NE555	Precision timer.................	1.50
NE560	Phase lock loop DIP.............	3.25
NE561	Phase lock loop DIP.............	3.25
NE565	Phase lock loop TO-5............	3.25
NE566	Function generator TO-5.........	4.00
NE567	Tone decoder MINI DIP...........	4.00
NE567	Tone decoder TO-5...............	3.00
NE5558	Dual 741 op amp MINI DIP........	1.00
709	Popular Op Amp DIP..............	.45
710	Voltage comparator DIP..........	.75
711	Dual comparator DIP.............	.40
723	Precision voltage regulator DIP.	1.00
739	Low noise op amp DIP unmarked...	1.00
741	Op amp TO-5/MINI DIP............	.55
747	Dual 741 op amp DIP.............	1.00
748	Op Amp TO-5.....................	1.00
CA3018	2 Isolated transistors and a Darling-ton-connected transistor pair...	1.00
CA3026	Dual differential amp...........	1.00
CA3045	5 NPN transistor array..........	1.00
LM100	Positive DC regulator TO-5......	1.00
LM105	Voltage regulator...............	1.25
LM302	Op amp voltage follower TO-5....	1.25
LM308	Op Amp TO-5.....................	2.00
LM309H	5V 200 MA power supply TO-5.....	1.00
LM309K	5V 1A power supply module TO-3..	2.00
LM311	Comparator TO-5.................	1.75
LM370	AGC amplifier...................	2.00
LM380	2-Watt Audio Amp................	1.75
LM1595	4-Quadrant multiplier...........	2.00
8038	Sine square triangle function generator........................	4.95
MC1536T	Op Amp..........................	2.00

CALCULATOR KIT

Kit contains NORTEC calculator chip 4024,21 driver transistors, 8 LED's, 7 segment readout by EXITON, 3 LED for overflow, low btry, overrange, a keyboard by FLEX KEY. All this $45 value for only .... $16.50 2 for $27

BASIC CLOCK KIT

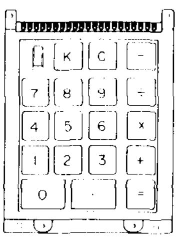
We furnish clock chip CT7001 and 4 first line LED readouts .3 inch hgt. You furnish the misc. minor parts. A $35 value for $19.95

LIQUID CRYSTAL WATCH

Build a wrist watch or desk clock with this liquid crystal display. Recently written up in several magazines. Display with special socket reads hours and minutes .... $15 2 for $25

UNDERWATER LISTENING

Brand new by OLIN. Use it for a swim pool monitor-alarm, use it on lake or ocean listening to underwater noises, fish, etc. Complete with hydrophone, 50 ft. mike cable, speaker-amplifier console. Operates from 115 volts AC or 15 volt dry cells. 12 lbs. $25

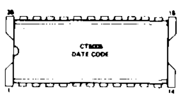


$5.00 CALCULATOR KEYBOARD 2/$9.00

Brand new key boards for hand held calculators. Two styles available.

One for use with calculator chip CAL TEX 5001-5002-5012 or MOSTEK 5010-5012. Another for use with Gen. Inst. chip C500.

Priced at ~~$8.00~~ each or two for ~~$15.00~~.



CT 5005 CALCULATOR CHIP

Single MOS chip with all logic required for 12 digit 4 function desk top calculator with extra storage register for memory or constant. Multiplexed 7 segment outputs for LED, Incandescent, Fluorescent, or Gas Discharge displays. Brand new, bargain priced, with specs.

$8.00 each, 2 for $15.00

AM-FM RADIO $5.50

Due to the West Coast ship strike they came in too late for the customer. Now it's your bargain. Use it as is or build it into your own cabinet, desk, wall, etc. All built, ready to use, with AC supply. To make it portable all you do is power it with a couple of "D" cells. Fully assembled solid state chassis with AC power supply, less speakers. Covers full AM as well as FM broadcast. The price . . . an astounding meager $5.50

313,344 CORE MEMORY $125.00

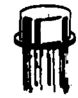
From SPECTRA computer, visually OK. 64 x 68 x 4 x 18 core stack. Figures out to 35K Byte.

CLOCK/CALENDAR ALARM CHIP

These large scale integrated (LSI) chips eliminate literally thousands of components or hundreds of chips in the construction of a clock. For most applications only a single supply and a minimum of components are required.

☐ 7001 Chip - Features 28/30/31 day calendar, 12/24 hour clock, 24 hour clock, 24 hour alarm, snooze alarm, 6 digit display, direct drive to luminescent anode tubes or LED segments, single transistor interface with Sperry displays. Segment and digital outputs can be "wire" or "D" to share calculator displays .... $9.95 each, 2/$18.00

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

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.

CALCULATOR CHIPS

5001 LSI (40 pin) Add, subtract, multiply divide 12 digit. With data $6.95
Data alone .... 50 cents

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CMOS HEX INVERTER, dual inline package. 3-18 volt range, dual diode protection against static charge. Dielectrically isolated complimentary MOS. $1.00 each 12 for $10.00

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Dual 16 bit memory, serial MOS by Philco TO-5 case, brand new with 2 page specs.
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2048 bit MOS LSI random access memory NEC 6003. All inputs except clock are TTL compatible. 2048 word by 1 bit. 22 pin ceramic dual-in-line. With specs.
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Another SUPER SCOOP by Meshna. Brand new RCA packaged, considered obsolete by RCA but what an exotic opto-electronic device for the sophisticated experimenter. Only several hundred on hand. Values shown are approx. as each diode characteristic varies. Each is marked with correct value.

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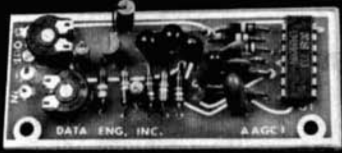

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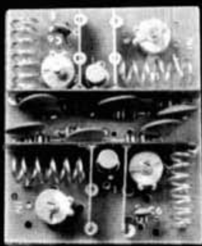

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QSL'S — BROWNIE W3CJI — 3035B Lehigh, Allentown, Pa. 18103. Samples with cut catalog 35¢.

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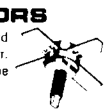

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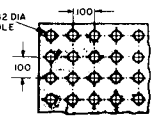

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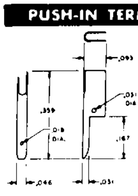

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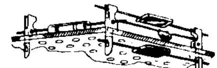

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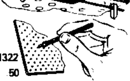

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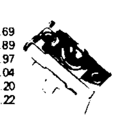
405	1.01	425	.85	462	.52	468	1.20
406	1.04	426	1.01	463	.64	469	1.22
420	58	427	1.12				



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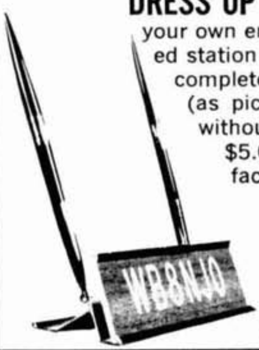

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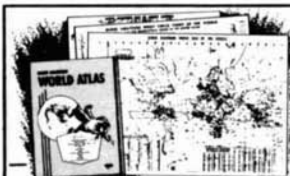

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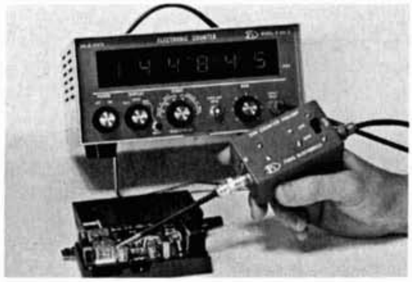

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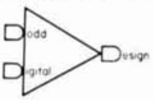

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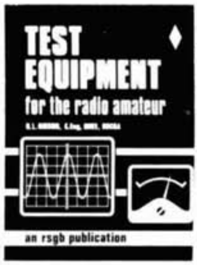

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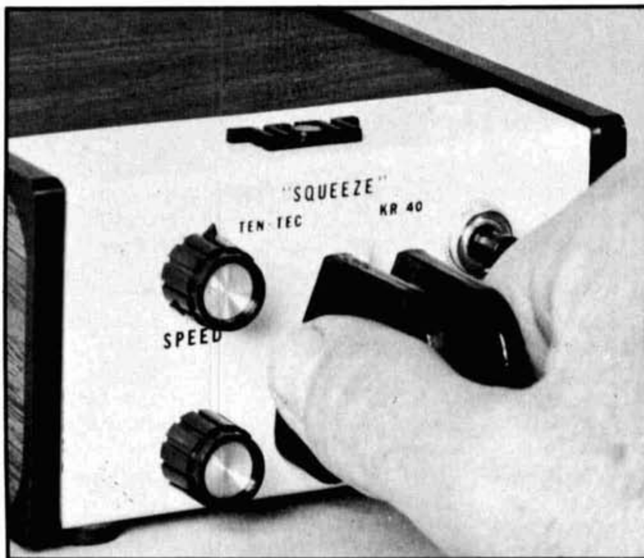

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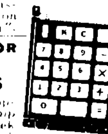

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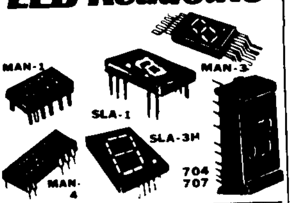

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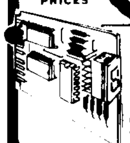

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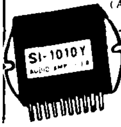
707*	.33 h.	Litronics
704**	.33 h.	Litronics
SLA-1*	.33 h.	Opcoa

* Pin-for-pin MAN-1, ** Pin-for-pin MAN-4, elec. char. same



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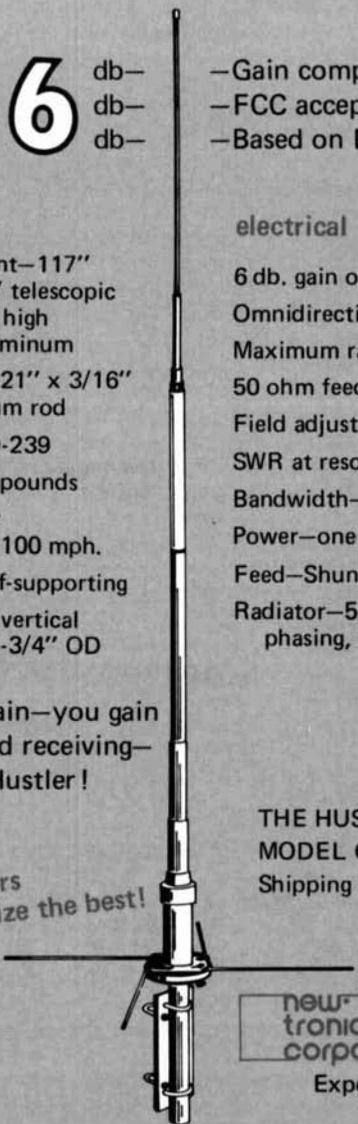

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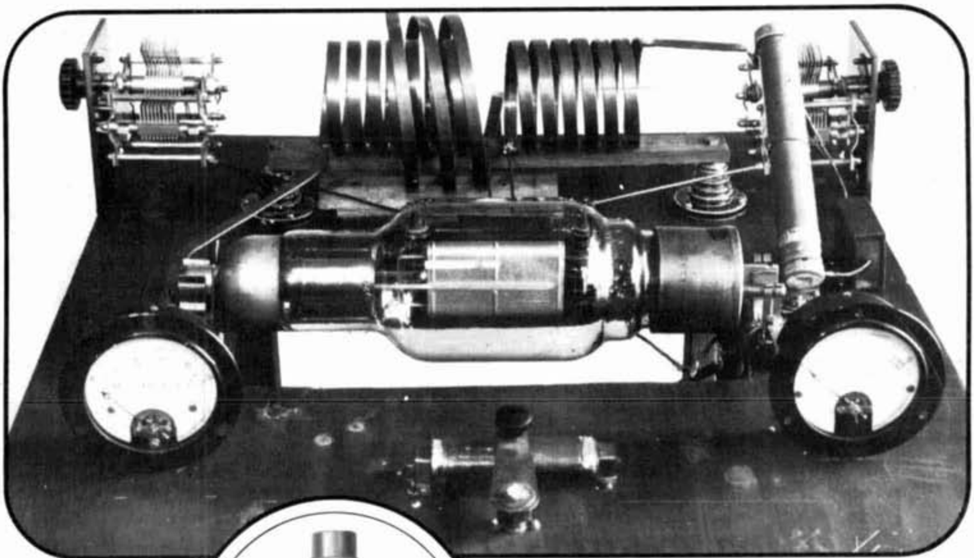
Ed Willis, 6TS, did it in 1925. He exchanged 20 meter signals across the continent—in daylight—with John Reinartz, 1XAM. Both Ed and John used 204 tubes for this record-breaking QSO.



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